



Introduction to Tube Amplifier Theory: 10.02.15

Featuring the AX84 P1-eXtreme Amplifier

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The P1-eX amplifier is a simple three-stage vacuum tube electric guitar amplifier. As outlined in Fig. 1, it consists of two preamp stages driving a power amp stage. Depending on the choice of output tube, this amp can deliver between 7 and 15 watts into a 4-, 8-, or 16-ohm load.

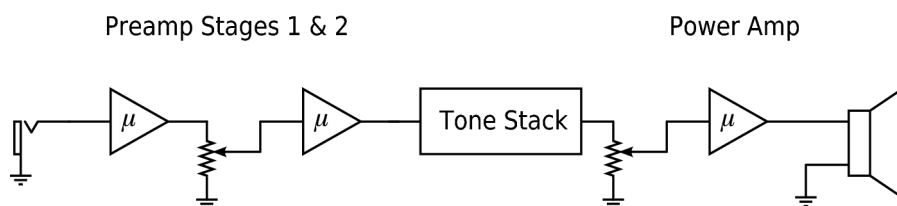


Fig. 1: P1-eX block diagram

Placed between the two preamp stages is a preamp volume control. In a similar manner, the bass, middle, and treble tone controls and a master volume control are placed between the preamp and the final power amp stage. The guitar amplifier performs two primary functions: One, it amplifies the small voltages and currents produced by the guitar pickup into a signal powerful enough to drive a speaker. Two, it shapes the frequency response, tonality, and distortion characteristics of the raw guitar signal into a form pleasing to the musician.

To better grasp how a guitar amplifier accomplishes these functions, let's take a walk through the inner workings of the AX84 P1-eXtreme amp (P1-eX). For reference, there is copy of revision 06.03.16 of the schematic provided in Appendix 1 at the end of this document. To fully understand how a guitar amp works, even a simple one like the P1-eX, you must know how to read schematic diagrams, and understand what things like resistors and capacitors are. You also need some knowledge of basic algebra and electronic theory.

HOW THE HECK DO TUBES WORK ANYWAY?

So you run to the bookstore and buy a book on basic electronics. Hnm..., no tube chapter. Even some good tube amp books fail to explain how tubes work. Many different types of tubes have been developed over the years, but guitar amps generally use only three types: diodes, triodes and pentodes.

VACUUM DIODES

Let's start with the vacuum diode, because it is the simplest type of vacuum tube. Inside the glass bottle, there are a few metal parts: the filament, the cathode, and the plate. The filament is sometimes called the heater, because that is exactly what it does—it heats the cathode. Like the filament in a light bulb, the tube filament is a thin length of wire that gets hot when electricity flows through it.

As the filament heats the cathode, it emits electrons from its surface in a process called [thermionic emission](#). With the cathode warmed up, a cloud of electrons gathers around the cathode, as shown in Fig. 2(b). The electrons represented in this drawing have no place to go because the plate is negatively charged with respect to the cathode. In matters of electricity, like-charged objects repel each other, so the negatively charged electrons just hang around the cathode like farm boys dreaming of big city lights! But consider Fig. 2(c). Things change when the plate is made positive with respect to the cathode. Now the electrons stream off from the cathode to the plate, causing current to flow¹ in the plate-cathode circuit. If you think about this behavior, you see that the vacuum diode functions as a one-way valve that allows current to flow in only one direction. It is this function that makes it useful for converting alternating current (AC) into direct current (DC).

In vacuum tube amplifiers, vacuum diodes are often used to rectify the AC line voltage into a DC voltage suitable for powering the plates of other tubes in the circuit. Typical examples include the [EZ80](#) and the [5U4](#) types. Solid-state silicon diodes behave in a similar manner and are often used in place of vacuum diodes in tube amps because of their lower cost and improved reliability. Yet, there are some sonic gains to be had from using a vacuum rectifier in push-pull power amplifiers due to the greater overall load-dependent voltage drop introduced by the rectifier. This is sometimes called “sag.” Single-ended amplifiers, such as the P1-eX, present roughly the same load to the power supply whether they are at idle or running flat out, so the voltage sag from the tube rectifier does not occur. Consequently, a vacuum tube rectifier adds cost and complexity, but little sonic value, to amps like the P1-eX. While some early versions of the P1 amp used tube rectifiers in their power supplies, the P1-eX relies on solid-state diodes for mains rectification. Besides their function as rectifiers in power supplies, diodes also are used in many other types of circuits, including automatic volume controls, compressor/expander circuits, and radio signal detectors. We won't spend any more time on vacuum tube diodes, though we will discuss the P1-eX power supply later on.

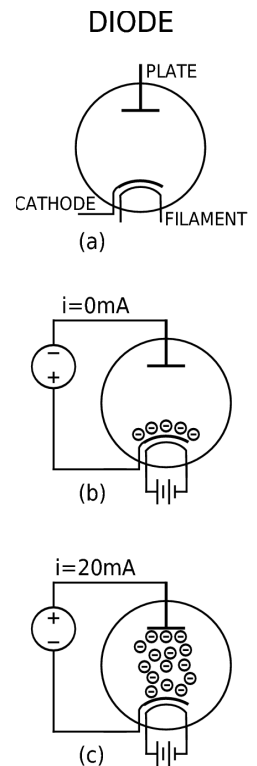


Fig. 2: The diode, forward and reversed biased

¹ A note on conventional electric current vs. electron flow: [Conventional electric current](#) is defined as the flow of positive charge in a circuit. This is because the conventional definition of electric current was developed before the discovery of the electron and its negative charge. Back in the early days, electronics experimenters knew that invisible bits of charge flowed through the wire in a circuit, but had no way of determining the direction the bits traveled. They took their best guess. Bzzzzt! Wrong! By the time the not-so ancients realized the mistake, it was too late. By convention, most electronics texts refer to electric current and mean “the negative flow of electron current.” There are exceptions in electronics texts as well as in texts from other fields. Some treat current as flowing from positive to negative and some as flowing from negative to positive. Additionally, some materials allow current flows in both directions but carried by different particles. Do not let this situation confuse you. From the perspective of vacuum tube design, the direction of the current flow rarely matters, as long as you treat it consistently. The important thing to remember is that current is a measure of the volume of electrons flowing past a particular point in a circuit. The greater the volume, the more useful work (or damage) you can do with it.

VACUUM TRIODES

Things get much more interesting when you add an electrode between the cathode and plate of a vacuum diode, creating a triode (Fig. 3). The 12AX7 commonly found in guitar amplifiers contains two triodes inside its glass envelope. This new electrode is called the control-grid (or grid, for short) and is usually constructed of a mesh of thin wires positioned between the plate and cathode—very close to the cathode. If you were to connect the grid to the cathode, the tube would behave much like a diode. Most of the electrons would flow right past the grid on their way to the plate. Do not try this. Most triodes are not designed to be operated in this manner.

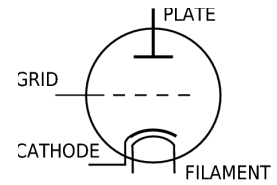


Fig. 3: Vacuum Triode

When charge on the grid is made more negative with respect to the cathode, the electron flow from cathode to plate starts to get pinched off. Fig. 4 illustrates this property. The device in this drawing is conceptual, so don't go hunting for a real triode that produces these results. In Fig. 4(a) the grid voltage is -1V , so it is not very repulsive to electrons and many of them successfully travel to the plate, which allows the plate circuit to draw 10mA . As the grid voltage becomes more negative with respect to the cathode, as shown in Fig. 4(b), more of the electrons are repelled by the growing negative charge on the grid, and the number of electrons that pass through the grid is reduced. In this example, the current in the plate circuit falls to 5mA . As the grid voltage is driven even more negative than the cathode, the electron flow is reduced further until at some negative voltage the current cuts off completely, as illustrated in Fig. 4(c). When the triode is cut off like this, no current can flow in the plate circuit, and increasing how negative the grid is with respect to the cathode beyond this point has little effect on the behavior of the triode.

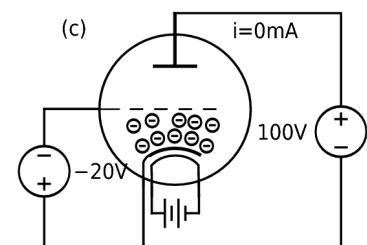
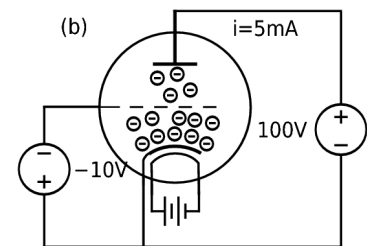
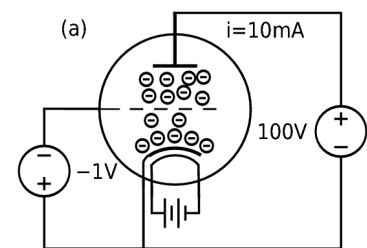


Fig. 4: Current falls as grid voltage goes more negative with respect to cathode

Every AC signal cycles up and down with respect to some level of DC bias voltage. If your house is like mine, the main AC voltage is about 125V_{RMS} (**root-mean-square**), meaning the voltage cycles back and forth in a sine wave from a positive peak of $+177\text{V}$ to 0V to a negative peak of -177V . That zero in middle indicates it has a DC bias voltage of 0V (or it should have if your wiring is correct). In a similar manner, the signal on the grid has a DC bias voltage. It could be 0V , like your house mains, or it could be some positive or negative DC voltage. That DC bias voltage, when applied to the grid, is called “**grid bias**.” It is also sometimes referred to as the zero-signal grid voltage because it is the voltage the grid sits at when the signal is reduced to zero AC volts. A negative DC bias voltage sufficient to reduce plate current to zero is called the “**cut-off bias**.”

When the grid voltage becomes the same as the cathode voltage, the grid bias is said to be at 0V , and the onset of grid-conduction occurs. Grid conduction means that current flows in the grid circuit as well as in the plate circuit. Although most small-signal triodes do not tolerate sustained operation in the

grid-conduction region, many power triodes and other sorts of tubes are designed for it. Continued increases in grid bias voltage allow even more current to flow in the plate and grid circuits, up to a limit. The point where plate and grid currents no longer increase, regardless of increases in grid bias voltage, is called the *saturation point*. The saturation point is determined by the cathode's ability to emit electrons at a given operating temperature, which is governed by the cathode's material and construction.

Between the cut-off point and the onset of grid-conduction, a triode behaves as a roughly linear device. Small-signal triodes are typically designed to operate in this linear region. Between grid-conduction and saturation, a triode's behavior becomes very non-linear because the grid is drawing off progressively more of the current flow. For a given triode stage, we ensure operation in this linear range by setting the zero-signal grid bias about halfway between the cut-off point and the onset of grid-conduction. So what does it mean to set the bias voltage of a stage and how do we do that? The answer to that question is in the next section.

BIASING A 12AX7 GAIN STAGE

Before we begin the discussion of setting the zero-signal bias voltage, let's establish three common abbreviations that will appear in the equations below:

E , meaning [electromotive force](#) measured in volts.

I , meaning [electric current](#) measured in amperes.

R , meaning [electrical resistance](#) measured in ohms.

The relationship among these three electrical properties is defined by [Ohm's law](#):

$$E = I \times R \text{ which can also be expressed as } I = \frac{E}{R} \text{ or } R = \frac{E}{I}$$

Don't let the math throw you. Ohm's law basically says that the current through a given circuit element times the resistance presented by that element gives the voltage drop across that element. The cool thing about Ohm's law is that if you have two of the values, you can figure out the remaining value. Suppose you measured the voltage with respect to the circuit ground just before and just after a 1K-ohm resistor and observed a 12 volt difference between the two. The resistor is said to drop 12 volts. You can plug the voltage and resistance values that you know into Ohm's law, as follows:

$$I = \frac{E}{R} = \frac{12}{1000} = .012 \text{ A} = 12 \text{ mA}$$

You can see that 12mA flows through the resistor. That means that the circuit following the resistance is drawing 12mA. Ohm's law is a powerful tool that will help us analyze how the P1-eX amplifier functions.

Take a look at Fig. 5. It shows the first preamp stage from the P1-eX schematic. This is an example of a resistance-capacitance coupled, grounded-cathode voltage amplifier. It operates such that the voltage across plate load resistor (R22) is an amplified and inverted copy of the grid signal voltage at pin 7. The signal enters the amplifier via J2, where you plug in the guitar. R25 provides a load across which the signal from the pickups can form. The grounded circle around the lead from the input jack to R24 indicates shielded wire intended to prevent electromagnetic field (EMF) interference from mixing with the signal. R24 is a grid-stopper resistor and its purpose is to roll-off high frequencies to further limit EMF interference. Typically, it is mounted directly on the tube socket to further reduce the chance of EMF noise.

Some designers prefer to reduce the value of R24 because of its effect on the high-frequency response of the preamp. Others remove it all together. In such cases, it is important that the ground at the input jack serves as the signal ground to chassis ground tie point. Otherwise, the lead from the input jack to the first preamp grid can act as a radio antenna.

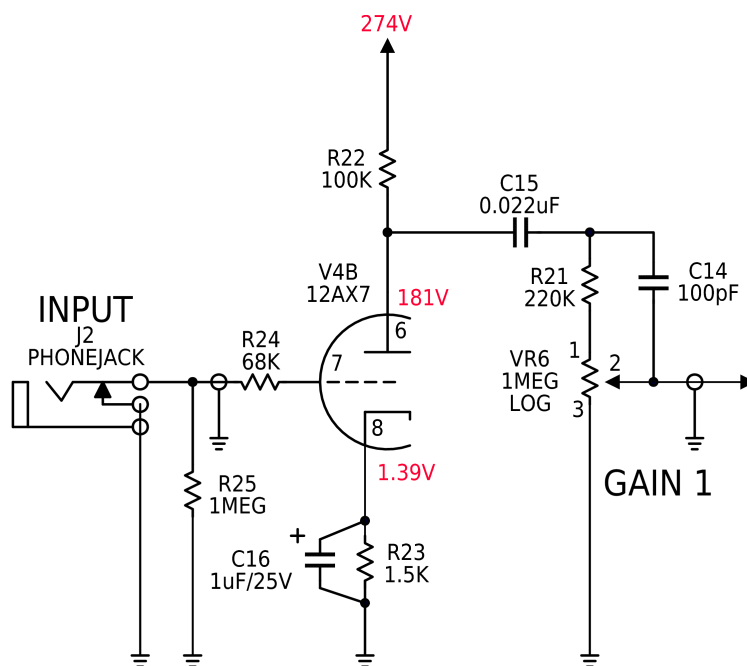


Fig. 5: Stage one of the P1-eX preamp is a 12AX7 gain stage

Let's think about what is happening in that circuit. First take a look at the I_p - E_p graph from the 12AX7 data sheet, shown in Fig. 6. It looks confusing at first, but this graph can tell you all sorts of things about the tube's behavior in a circuit once you learn how to unpack it. The curved lines that slope upward to the right are called "plate curves." Each one represents the behavior of the triode when the grid is at a specific bias voltage (0V, -0.5V, -1V and so forth). Recall in Fig. 4, that as the grid voltage became more negative, the plate current dropped. You can see that behavior in the 12AX7 triode. At a constant

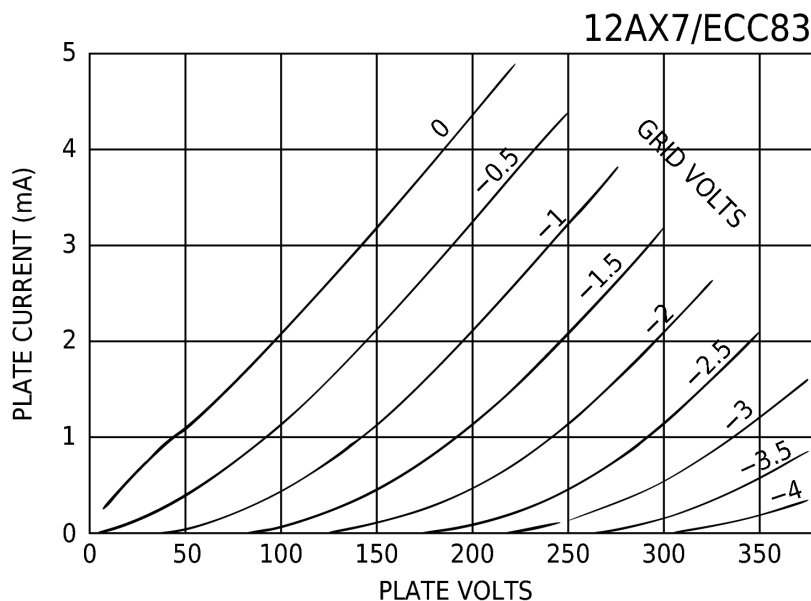


Fig. 6: 12AX7 I_p - E_p graph

150V on the plate, the plate current is about 3.2mA when the grid is at 0V, about 1.2mA when the grid is at -1V, and about 0.2mA when the grid is at -2V. At a constant 150V plate volts, the triode cuts off with a grid bias voltage somewhere between -2V and -2.5V.

The signal coming into the grid of the triode is an AC signal. Its voltage is swinging up and down around the DC bias voltage at the frequency of the notes being played. It's going positive with respect to the bias voltage and then negative with perhaps a 1V-to-2V peak-to-peak swing. If this grid signal swings sufficiently negative to cause the current in the plate circuit to cut off, then there would be no voltage drop in R22 (remember Ohm's law $E=R*I$), and the voltage at the plate would equal the power supply voltage, which is 274V for stage one of the P1-eX preamp. You can see that a grid voltage of approximately -3.6V is necessary to cause the triode to cut off when the plate is at 275V. Grid-conduction occurs when the grid voltage is 0V, so we want a bias point about midway between these two points; somewhere between -1V and -2V would be fine.

When no signal is present at the input, the grid of V4B is connected to ground through resistors R24 and R25, so the DC voltage (E_g) at V4B's grid is 0V. The problem is how to make the grid voltage negative. We could use a fixed negative DC source like a battery or a separate power supply. But we don't have to resort to those means. Remember, the grid only has to be *negative with respect to the cathode*. If we can raise V4B's cathode voltage to about 1.4V, we can accomplish the same thing as lowering the triode's grid from 0V down to -1.4V.

One convenient way to do this is to place a resistor between the cathode and ground. This raises the cathode above ground by the voltage dropped across the cathode resistor. This technique for setting the bias voltage is called self-bias or cathode-bias. Look back at the schematic in Fig. 5. The voltage across R22 is:

$$E_{R22} = (274V - 181V) = 93V$$

Using Ohm's law, the current through R22 is:

$$I_{R22} = \frac{E_{R22}}{R22} = \frac{93V}{100K\ \Omega} = 0.93\ mA$$

So, 0.93mA flows in the plate circuit of V4B. The current in the cathode circuit of a triode is the sum of the currents in the plate and grid circuits. Because our triode is operating below the onset of grid conduction, nearly zero current flows in the grid circuit. For practical purposes, the current flowing in the cathode circuit is equivalent to the current in the plate circuit. Therefore, using Ohm's law again, we can multiply I_{R22} times the cathode resistance (R_k) to determine V4B's cathode voltage. R_k in the schematic is identified as R23 and has a value of 1.5K ohms:

$$E_{k(V4B)} = I_{R22} * R_{k(V4B)} = 0.93mA * 1.5\ K\ \Omega = 1.40V$$

Remember that the bias is the grid voltage referenced to the cathode voltage. We can easily calculate the bias voltage of V4B by subtracting cathode voltage of V4B from its grid voltage. Recall above that at zero-signal the grid voltage is 0V, so that gives:

$$E_{grid-bias(V4B)} = E_{g(V4B)} - E_{k(V4B)} = 0 - 1.4 = -1.40V$$

That's good; 1.40V is right where we want it to be.

Wasn't it convenient that we knew what the plate voltage was? If you were designing your own preamp stage from scratch, you would have to calculate the zero-signal plate voltage and corresponding bias voltage. There are a few ways to do this. Some use computers to model the circuit in PSPICE (models for popular tubes are available on [Duncan Munro's website](#)). Others prefer the challenge of doing the math by hand (ouch!). I'm a visual sort of engineer and like drawing, so I'll show you how to draw a load line.

A load line is simply a graphical representation of the circuit that shows the relationship between the plate load resistor (R22) and the triode (V4B) in the context of the I_p - E_p graph from the data sheet. Suppose you had already decided that your B+ was to be 275V and your load resistor R22 was to be 100K. Because R22 is a fixed resistance, it will plot as a straight line. One end of that 100K load line would be at the cut-off point where plate current is zero and plate voltage equals B+ (275V, 0.0mA). To find the other end of the 100K load line, use Ohm's Law to divide B+ (275V) by the value of the plate load resistor (100K). This gives the plate current of 2.75mA, such as it would be when the the load resistor is dropping the entire B+ voltage and the plate is at 0V (0V, 2.75mA). Fig. 7 shows this line in red on the 12AX7 I_p - E_p graph. Its upper left terminal point is at (0V, 2.75mA) on the graph, and its lower right terminal point is at (275V, 0mA) on the graph. You can do this with any value of load resistor and B+ voltage.

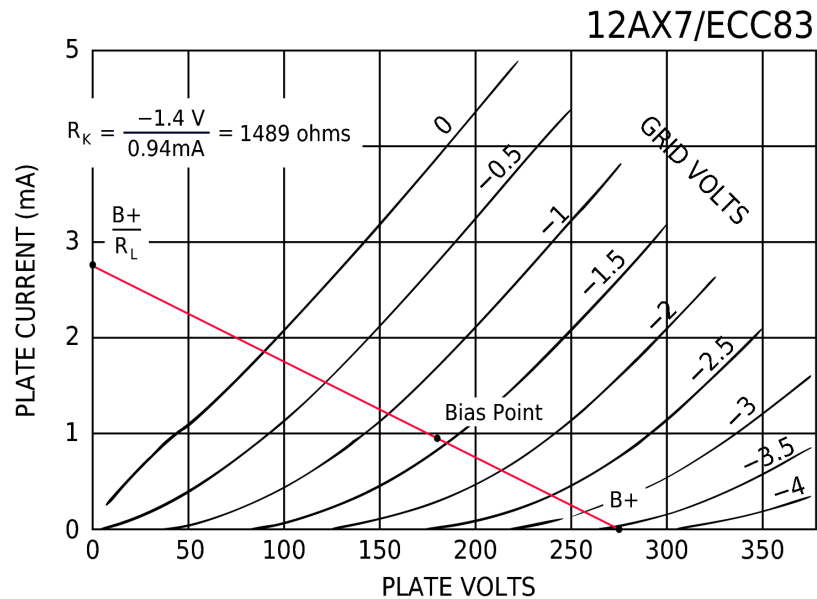


Fig. 7: A 100K load line with an operating point at -1.4V grid bias

Now let's assume we want a grid bias voltage of -1.4V. How do we figure that out? There is no plate curve on the I_p - E_p graph for a grid voltage of -1.4V. It's easy: We extrapolate the value. Imagine that there were plate curves for -1.1V, -1.2V, -1.3V, and -1.4V on the graph. Because the device is linear in this region, those curves would be more-or-less evenly spaced between the -1V and -1.5V plate curves. That means that the -1.4V plate curve would intersect the R22 load line up from the -1.5V intersection at about 1/5 the distance between the -1V and -1.5V plate curves. I've drawn a dot there and labeled it "Bias Point." The bias point is at approximately (180V, 0.93mA) on the graph. This means that the zero-signal plate voltage will be about 180V, with a idle current of about 0.93mA. We know we want the cathode to be about 1.4V above ground and we know the idle current; therefore, we can calculate the desired cathode resistance using Ohm's law:

$$R = \frac{E}{I} = \frac{1.4V}{0.00093} = 1505 \Omega$$

After an amp is built, of course, it is easy to measure the actual plate and cathode voltages and calculate the current from those values. But if you are designing a stage from scratch, then you can get there from a load line. Suppose you wanted to use a 75K-ohm load resistor with a B+ of 250V and the grid bias set at -1.2V. Try working that out on the 12AX7 I_p - E_p graph. Is the load line going to be steeper or flatter than the 100K load line? If you said steeper, you are on the right track. See if you can determine what cathode resistance would be necessary to achieve that bias voltage. What would the quiescent plate voltage be at that bias point? Don't look before you try to work this out, but there is a graph of that load line and operating point in Appendix 2 at the back of this document.

The choice of bias point affects the tonal and distortion characteristics of a given preamp stage. You may want to tune your amp for different tonal qualities by modifying the bias point of either the first or second preamp stages. In general, making the bias less negative with respect to the cathode (increasing the current through the triode) makes for a warmer tone but affords less clean headroom. Alternatively, making the bias more negative with respect to the cathode makes for a cooler tone. Tastes vary, so as you advance from building an existing design like the P1-eX to creating your own designs, you may wish to experiment with different operating points. You can even tweak an existing design with load changes and operating point changes. Note, however, that you can only push the operating point so far in either direction before you have to start rethinking the overall design of the stage and perhaps the entire amp.

AMPLIFYING THE INPUT SIGNAL

We set the bias point to keep the tube operating in the linear part of the curve. In plain English, this means the tube circuit is designed so that an input signal that doesn't try to push the grid into conduction or past the cut-off voltage will get amplified with minimal distortion. There is always a slight amount of distortion because a vacuum tube is not a perfectly linear device. We won't worry about this here; it would be too complicated to add this non-linearity into the calculations below.

The two preamp stages in the P1-eX circuit amplify the input signal in series before it is delivered to the output stage or power amp (V1). So the question is, how much amplification occurs in the preamp? Now that we know the operating point of the first triode stage, we can determine its voltage gain. To do this, we need to understand three operational parameters that apply to triodes: the *amplification factor*, *transconductance*, and *plate resistance*.

The amplification factor is a measure of how well the triode amplifies the grid voltage; it is the ratio of the change in plate voltage to the change in grid voltage. On tube data sheets it is usually denoted by the symbol μ :

$$\mu = \frac{\text{change in plate voltage}}{\text{change in grid voltage}} \text{ (plate current held constant), or more formally } \mu = \frac{\Delta E_p}{\Delta E_g}, I_p \text{ const}$$

Of the three triode parameters, it is the most constant across various operating points. So much so that we generally treat it as a constant and take the value given in the data sheet. Because it is a ratio of two voltages, it has no unit of measurement.

Transconductance (or mutual conductance) is the ratio of change in plate current to the change in grid voltage. In the linear operating region, it is roughly constant, but fluctuates considerably as the operating

point moves out of this region. This ratio is loosely a measure of how efficiently the grid controls the plate current. It is usually denoted in tube data sheets by the g_m :

$$g_m = \frac{\text{change in plate current}}{\text{change in grid voltage}} (\text{plate voltage held constant}) \text{ or more formally } g_m = \frac{\Delta I_p}{\Delta E_g}, E_p \text{ const}$$

The unit of transconductance is called a *Siemens*, or in the older texts the *mho*. While *Siemens* is now the international standard name for the unit of conductance, it was not widely used at the height of tube design. Because many of the tube data sheets available on the Internet use the term *mho*, so will we.

The third primary operating parameter of interest is plate resistance (r_p , some texts refer to this as r_a .) or the ratio of the change plate voltage to the change plate current. It helps to think of a triode as a variable resistor. When the grid voltage is made more positive, more current flows through the tube. This means the effective resistance between the plate and cathode decreases. Make the grid voltage more negative, and this resistance increases. Plate resistance, like any other resistance, is measured in ohms :

$$r_p = \frac{\text{change in plate voltage}}{\text{change in plate current}} (\text{grid voltage held constant}) \text{ or more formally } r_p = \frac{\Delta E_p}{\Delta I_p}, E_g \text{ const}$$

Looking at the similarity of these parameters with respect to each other, you might wonder if they are somehow related. They are, in fact, mathematically related, and we will take advantage of that later. First, however, let's examine the triode parameters in a bit more detail.

The μ , g_m , and r_p values for a particular tube are usually specified at a couple of different plate voltages in the tube's data sheet. For instance, a 12AX7 data sheet I have indicates:

Plate Voltage	100	250	VOLTS
Amplification Factor (μ) also called “mu”	100	100	
Transconductance (g_m)	1250	1600	μ MHOs
Plate Resistance (r_p)	80K	62.5K	OHMS

While the amplification factor of most triodes can usually be considered a constant regardless of the operating point, the same is not true for transconductance and plate resistance. Because the plate voltage at our operating point (181V) is neither 100V or 250V, we have to determine the transconductance and plate resistance for our quiescent operating point. There are a variety of ways to estimate these values for a given operating point, but to really see what they mean for the tube, it's a good exercise to plot them on the plate curves.

Consider the enlarged view of the 12AX7 I_p - E_p plate curves around our operating point that is shown in Fig. 8. The first thing we do is draw a vertical line through the operating point, making sure it intersects both the -1V and -2V grid voltage curves. This represents the constant plate voltage needed for calculating the transconductance value. Next, we plot horizontal lines from the intersections at the -1V and -2V grid curves to obtain the plate currents at those grid voltages. These are marked I_a and I_d on the drawing.

At E_{g1} (-1V), the plate current I_a is 1.73mA and at E_{g2} (-2 V), the plate current I_d is 0.30mA. Plugging those values into the transconductance formula gives:

$$g_m = \frac{\Delta I_p}{\Delta E_g} = \frac{I_a - I_d}{E_{g1} - E_{g2}} = \frac{1.73 \text{ mA} - .31 \text{ mA}}{-1 \text{ V} - (-2 \text{ V})} = 1420 \mu \text{ mho}$$

Calculating plate resistance takes a similar approach. First, plot a grid curve through the operating point that is roughly parallel to the nearest specified grid curve. This represents the -1.4V grid voltage that must stay constant for the plate resistance calculation. Next choose two voltages around the operating point, say 175V and 185V. Those are marked as V_a and V_b on the graph. Next draw vertical lines from those voltages to intersect our -1.4V grid curve and obtain the associated plate currents at I_b and I_c .

At V_a (175V) on the -1.4V grid curve, the plate current I_c is 0.88mA and at V_b (185V) on the -1.4V grid curve, the plate current I_b is 1.03mA. Plugging those values into the plate resistance formula gives:

$$r_p = \frac{\Delta E_p}{\Delta I_p} = \frac{V_b - V_a}{I_b - I_c} = \frac{187 \text{ V} - 173 \text{ V}}{1.05 \text{ mA} - 0.86 \text{ mA}} = 73.6 \text{ K } \Omega$$

As you can see, it is relatively easy to plot the transconductance of a tube on the plate curves. All it takes is drawing a vertical line through your quiescent operating point. Accurately plotting a plate curve through an arbitrary operating point is much more difficult and generally not worth the trouble because, as mentioned earlier, the three parameters are interrelated. Specifically, the amplification factor of a triode is the product of its transconductance and its plate resistance. Mathematically, this is expressed as:

$$\mu = g_m r_p \text{ which is also equivalent to } g_m = \frac{\mu}{r_p} \text{ or } r_p = \frac{\mu}{g_m}$$

Recall that the amplification factor (μ) of a triode is relatively constant over a large range of operating conditions. Therefore, if you know either one of the other operating parameters, it is simple to calculate the missing parameter. Since transconductance is easier to plot accurately, we typically plot it for the desired operating point and then plug it and the triode's amplification factor into the above formula and solve for the plate resistance.

$$r_p = \frac{\mu}{g_m} = \frac{100}{.001420} = 70422 \Omega \text{ (We will round this to 70K ohms.)}$$

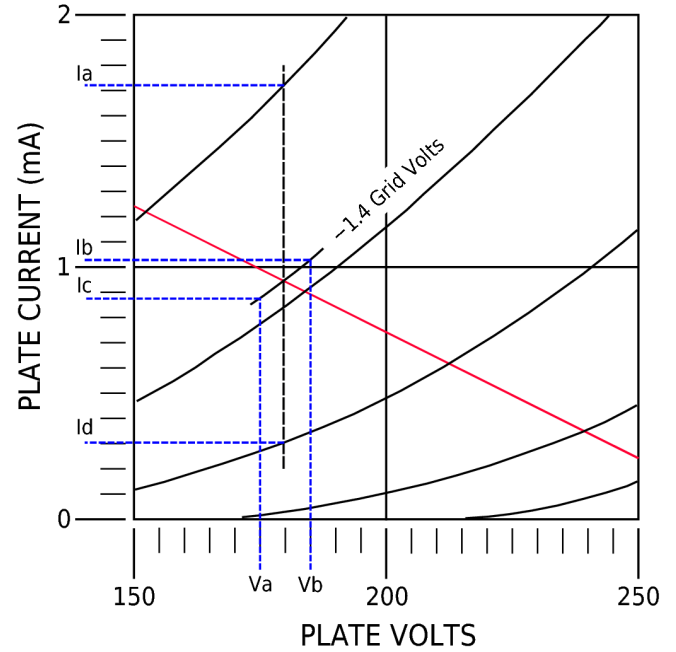


Fig. 8: Calculating transconductance and plate resistance on the I_p - E_p graph

Once we have the plate resistance and the transconductance, we calculate the voltage gain for this sort of circuit using this formula:

$$Gain = \frac{r_p \parallel R_l \parallel R_g}{\frac{1}{g_m} + R_k}$$

where:

g_m is the transconductance of the triode

r_p is the plate resistance of the triode

R_l is the plate load resistance (note: some documents refer to this as R_a or R_p)

R_g is following grid resistance

R_k is the unbypassed resistance in the cathode circuit

The symbol “ \parallel ” is shorthand to indicate resistances in parallel. The plate load resistance is simply the value of the plate resistor (R22 of the preamp stage one). The following grid resistance is the total resistance of the circuit connected to the plate of the tube, not including the plate load resistor. The cathode resistance is the unbypassed part of the cathode circuit.

Since we are talking about AC signals, let's just imagine that C15 has no resistance at all to the AC output signal (in actual practice, its resistance varies with frequency). The only reason C15 is in the circuit is to block the 181VDC present at the plate of V4B, and pass only the AC signal. That said, the following grid resistance seen by V4B is equal to 1.2M-ohms (R21+VR6).

The cathode bypass capacitor, C16, allows a path for AC signals to bypass R23. Consider what happens if C16 is not present. When an AC voltage is applied to the grid, it causes the current flowing through the tube to change. Without C16, the varying current through the tube would cause the voltage drop across R23 to vary, which would cause the grid bias voltage would vary with the signal. Such degenerative feedback from cathode to grid reduces the gain of the stage. The presence of C16 prevents that feedback for signals above a certain frequency. Though we won't discuss how to calculate it just yet, the 1 μ F value of C16 causes the first stage to have an effective cathode resistance of 0 ohms to all AC voltages above 200Hz. Plugging these values into the gain formula gives:

$$Gain = \frac{70K\Omega \parallel 100K\Omega \parallel 1.2M\Omega}{\frac{1}{1400\mu MHOS} + 0\Omega} = 55.7$$

So an AC signal of 100mV with a frequency above 200 Hz on the grid of V4B produces a 5.57VAC signal on the plate. At the junction of R21 and the top of VR6 the maximum signal strength is about 4.3VAC. Signals below 200Hz are somewhat attenuated because of the changes in grid bias we just mentioned.

The value of C16 has a marked effect on the stage's low-frequency response. For example, doubling C16 to 2 μ F causes the stage to be fully bypassed at all frequencies above 100Hz. Doubling C16 again to 4 μ F

lowers the bypass frequency down to 50 Hz. Going the other way and halving the value of C16 to 0.5 μF bypasses the cathode only for frequencies above 400 Hz. This is another place in the P1-eX where you may want to experiment with different component values. Generally with a guitar amplifier, you want some bass cut in order to keep the low frequencies from becoming muddy when the amp is driven hard.

COUPLING CAPACITORS

In the last section, we assumed that C15 presents infinite resistance to DC, and zero resistance to AC. High DC resistance is very important when a capacitor couples amp stages.

Consider what happens if some of the DC voltage makes its way through C15 into the next stage. Stage two of the preamp is designed so the grid is at -0.7V potential (with respect to the cathode) when no input signal is applied. If a positive DC voltage were applied to the grid, it would alter the bias and would require adjusting the cathode resistor to maintain the desired bias. So, if some of the DC voltage present at the plate of V4B were to get past C15, it would complicate the design of stage two. While some designs operate with direct coupling between the stages, we won't explore those here.

Thus, C15 needs to block DC voltages but let the audio signal through. Used in this way, a capacitor is sometimes called a “DC blocking capacitor” and more frequently called a “coupling capacitor” because it couples the output from one amplifier stage to the input of another stage.

The combination of C15 and VR6 creates a [high-pass filter](#). Lower frequency signals encounter more resistance than higher frequency signals. In the P1-eX, C15 is selected such that the cut-off frequency of this high-pass filter is below the frequency range of a guitar. If you substitute a smaller value capacitor for C15, you decrease the bass response of the amp. Increasing the value of C15 gives increased bass response—that is, lower frequencies are less attenuated by the coupling capacitor.

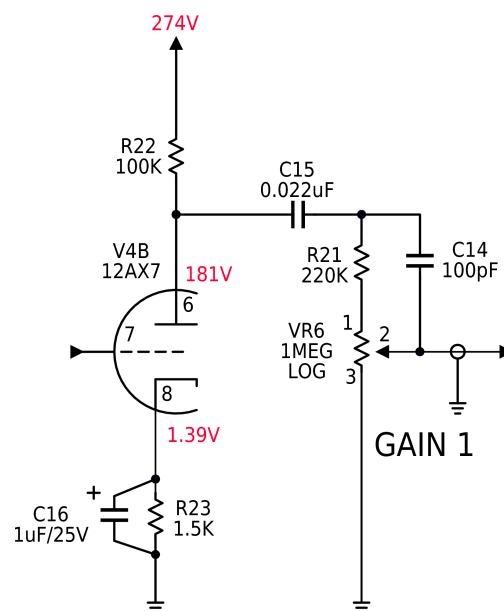


Fig. 9: Coupling capacitors

VOLUME CONTROL

R21 and the potentiometer VR6 form a variable voltage divider in the circuit, the output of which feeds the input grid of the second stage. At the lowest setting of VR6, the grid of V4A is connected directly to ground and all of the signal from the first stage of the preamp is shunted to ground. At the highest setting, the AC voltage applied to the grid of V4A is about 78 percent of the AC voltage present at V4B's plate.

R21 limits the stage-one gain before feeding it to stage two. But wait a minute! Isn't gain good? Up to a point, yes. But too much gain can drive the next stage into a particularly nasty sounding form of distortion called [blocking distortion](#)—a very non-musical sound that is sometimes called “farting out.”

I'll leave it to you to guess what this sounds like. R21 attenuates the signal a little to prevent this from happening. The greater the attenuation, the more noticeable it appears at higher frequencies, C14 is added to allow a greater proportion of the signal's high frequency energy into stage two. This helps brighten the stage and keep it from sounding muddy.

GAIN MANGEMENT

Besides the approach used in the P1-eX, there are various other ways to manage gain between stages. You can insert a voltage divider after the volume potentiometer, as shown in Fig. 10. With such an arrangement, the bright cap bypasses the top half of the divider. This example shows the 50 percent divider used in the P1 and HO projects.

A second approach to gain management is to use a split-load plate resistor. This technique, shown in Fig. 11, places the divider in the plate load. The main advantage of a split-load is that it attenuates the signal without changing its tone as much as other divider methods, thus reducing the need for bright caps. Refer to Steve Ahola's article, [“Split-Load Plate Resistors”](#) for more information.

A third approach is to eliminate the interstage attenuation and employ a lower-gain tube, such as a 12AU7. The disadvantage of approach is that it reduces the gain of both stages, which may not be desirable. There are also dissimilar triodes available, such as the 12DW7, that contain a low-gain and a high-gain triode in the same envelope. These present additional opportunities for experimentation.

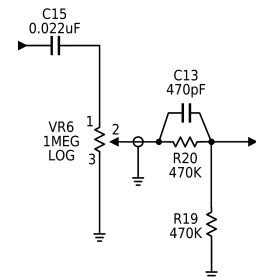


Fig. 10: Voltage divider

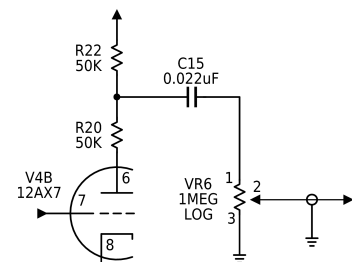


Fig. 11: Split-load plate resistors

PREAMP STAGE TWO

Using the same techniques described earlier, we can easily determine the bias current and voltage for V4A. Since the schematic shows the voltages, I will not bother with the math. Feel free to try your hand at the calculations and see if you come up with the same voltages as shown in the schematic. Fig. 12 shows a load line for stage two with a load resistance of 120K ohms and a quiescent operating point at 136V on the plate and -0.7V on the grid.

There are two significant differences between this load line and that of the first stage. The first is that it is less steep than the 100K-ohm load in the first stage. This results in slightly greater voltage gain for the stage, which overcomes some of the losses introduced by the tone stack that follows this stage.

The second difference is that the quiescent operating point is closer to the onset of grid conduction. Rather than setting the quiescent plate voltage about halfway between the plate voltage at the cut-off point and the plate voltage at grid conduction, the bias on the second stage sets the quiescent plate voltage about one-third the way between the plate voltage at the cut-off point and the plate voltage at the onset of grid conduction. This means that large signals tend to compress due to grid clamping before they begin to clip due to the signal trying to push beyond the cut-off voltage. This produces a smoother form of distortion when compared to clipping that many consider pleasing to the ear. To understand why this is so, let's take a look at what happens in the stage as the grid signal swing grows larger than 0.7V peak-to-peak. In other words, its peak voltage is greater than the grid bias voltage, as shown in in Fig. 13:

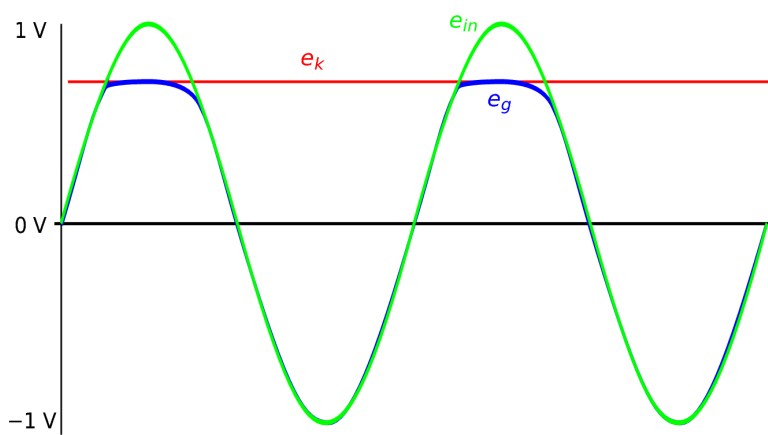


Fig. 13: Grid clipping due to an overdriven signal.

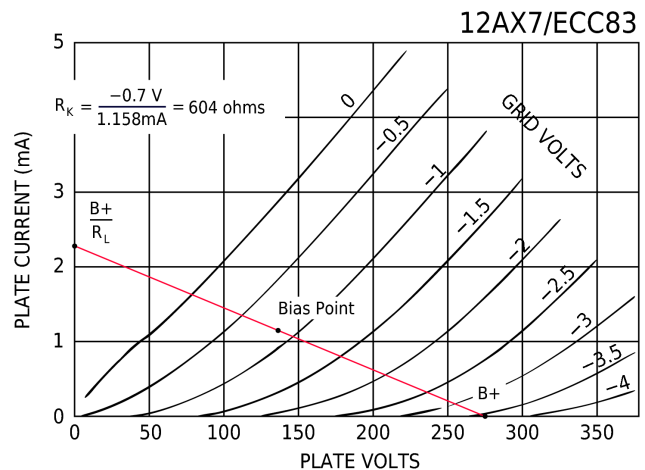


Fig. 12: 120K load line with the operating point at 136V/1.158mA

The curves in Fig. 13 represent the original input signal voltage (e_{in}), the level in volts that the cathode is lifted above the circuit ground (e_k) and the resultant clipped signal voltage that appears at the grid of V4A (e_g).¹ As you can see, the negative-going swing of e_g tracks closely with the input signal. On the positive-going portion of e_g , however, things get interesting. As the grid voltage approaches and passes the cathode voltage (that is, as the grid becomes positive with respect to the cathode), the grid must draw direct

¹ Waveforms in Fig. 13 and Fig. 14 courtesy of Merlin Blencowe.

current to track with the positive-going swing of the input signal. While there are plenty of electrons flowing out of the cathode that could go into the grid, recall that the DC circuit through which those electrons must flow includes a 1 M-ohm volume potentiometer at the input of the stage. This resistance limits the total current that can flow through the grid to a fraction of a milliampere. Because of this grid current limitation, the positive-going swing of the e_g is clipped. Such grid clipping of the input signal is inverted and amplified on the plate load, as shown in Fig. 14b:

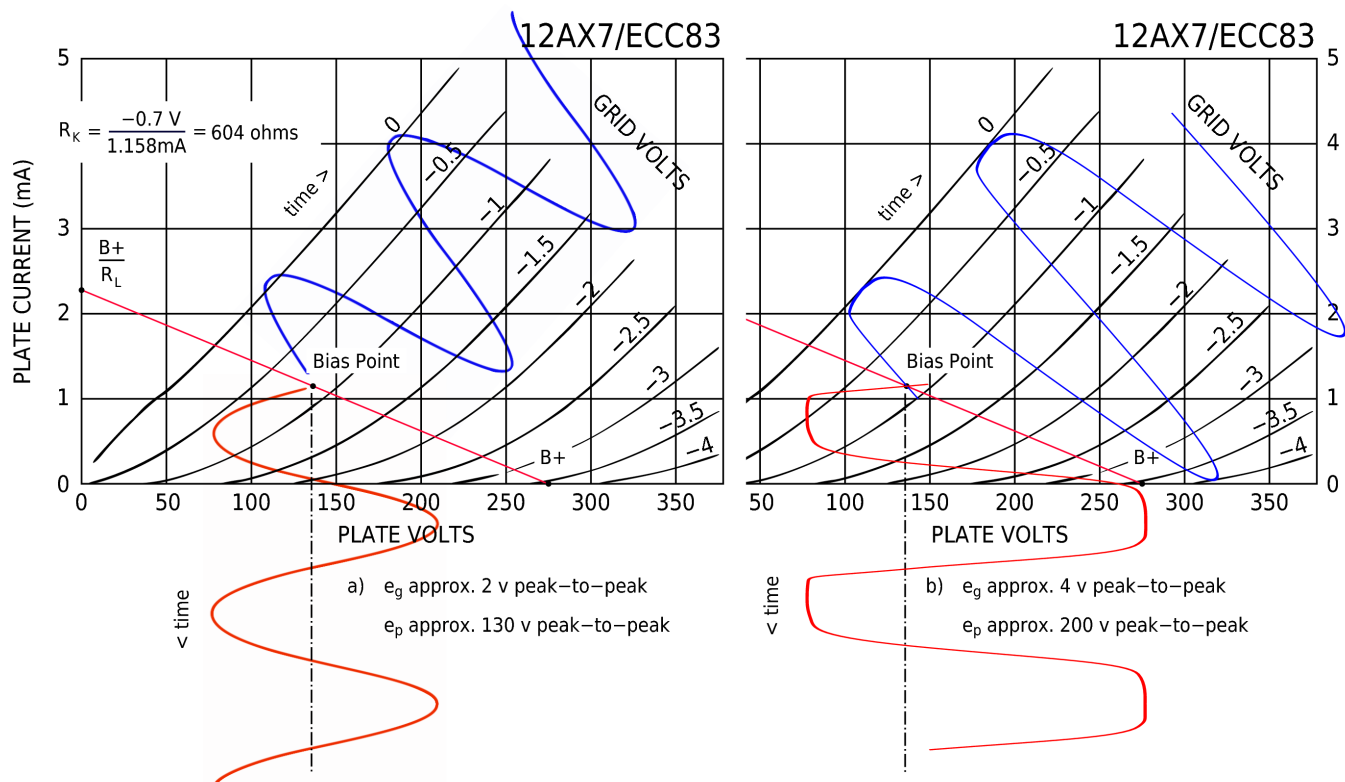


Fig. 14: Relationship between grid input and plate output signals

While the negative swing of the input signal to an resistively loaded stage is not limited by the grid, the inverted positive swing of the output signal is limited by the maximum power supply voltage available to the tube. If the grid voltage swings far enough negative with respect to the cathode, then plate current cuts off and the output signal clips as shown in Fig. 14b.

Notice in Fig. 14a, and similarly in Fig. 14b, how the downward swing of the plate signal from 136V at the bias point to 75V is shorter in amplitude than the upward swing from 136V to about 210V (or 275V in Fig. 14b)? This is called compression and is the result of the combination of the nonlinear behavior of the tube and choice of operating point.

All three forms of distortion—grid clipping, plate clipping and compression—have their place in guitar amplification and each contributes different harmonic characteristics to the signal. Thus, the selection of operating points and supply voltages determines an amplifier's characteristic tonal qualities. When designing a preamplifier it may be desirable, as is done in the P1-eX design, to bias some stages warmly by reducing the cathode resistance. In other cases, a designer might choose to bias a stage coldly with much larger cathode resistances to push the operating point much closer to the cut-off point. As you progress beyond a simple two-stage preamplifier, you will find plenty of room for experimentation.

Let's move on to calculating the gain of the second stage. As with the first stage, we determine the transconductance and plate resistance at the operating point. Recall the values from the data sheet:

Plate Voltage	100	250	VOLTS
Transconductance (gm)	1250	1600	MHOS
Plate Resistance	80K	62.5K	OHMS

As with stage one, the plate voltage of V4A falls between 100V and 250V, so we must estimate again. To do that we use the same graphing technique used for the first stage, but with a 120K-ohm load line and an operating point at 136V on the plate and -0.7V on the grid. This gives a transconductance value of 1600 and plate resistance of 62.5K.

Stage two of the P1-eX preamp (shown in Fig. 15) differs from stage one in that it has a tone stack in parallel with the plate load. It seems like it would be difficult to determine the effective resistance of this circuit. Well, it is, and that makes calculating the gain of this stage a bit more cumbersome, especially if we want to consider how the stage gain changes at different frequencies. However, we can approximate its effect on the gain of the stage if we make some assumptions.

Remember the formula for gain:

$$Gain = \frac{r_p \parallel [R_l \parallel R_g]}{\frac{1}{g_m} + R_k}$$

Hey! There are a couple of parentheses there that weren't there before. That's deliberate. We are going to pull that term out and calculate it separately. Here is the term:

$$R_l \parallel R_g$$

This term represents the effective resistance of the circuit attached to the plate of V4A. Once we have the effective load resistance, we plug it into the formula above. Here is how to calculate the effective load:

$$R_{effective} = (R17 \parallel R_g)$$

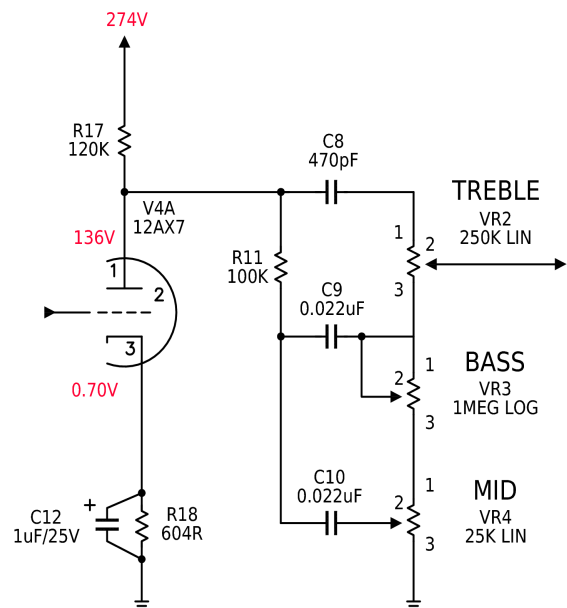


Fig. 15: P1-eX stage two of the preamp is followed by the tone stack.

Plug in the value for R17:

$$R_{effective} = (120K \Omega || R_g)$$

Now take a leap of faith. Just ignore the grid resistance of the following stage, as if the tone stack does not exist and the output impedance is infinite. Then, the equation above reduces to:

$$R_{effective} = 120K \Omega$$

Plug the numbers into the formula for gain:

$$Gain = \frac{65.2 K \Omega || 120 K \Omega}{\frac{1}{1600 \mu MHOS} + R_k}$$

The 1μF of bypass capacitance (C12) sets the half-boost frequency at about 283Hz (see the next section for details about how we determine the half-boost frequency), so for audio frequencies significantly above the half-boost frequency we get a gain of:

$$Gain = \frac{65.2 K \Omega || 120 K \Omega}{\frac{1}{1600 \mu MHOS}} = 67$$

As the frequency falls significantly below the half-boost frequency the gain gradually drops to a minimum of 34, which you calculate by assuming the cathode is not bypassed:

$$Gain = \frac{65.2 K \Omega || 120 K \Omega}{\frac{1}{1600 \mu MHOS} + 604} = 34$$

At the half-boost frequency, the stage gain is about 50; and around the half-boost frequency the gain shifts from the minimum to the maximum. A fully bypassed gain of 67 is considerably greater than the actual gain of stage. To see why, we'll have to add in the effect of the tone stack on the stage. Before we look at the tone stack let's see how to calculate the half-boost frequency. The math in the next section may be more difficult than you are willing to take on right now. If you want to skip it and move directly to the tone stack discussion, you can do so without worry. There is nothing in the tone stack section that depends on knowing how to calculate the half-boost frequency.

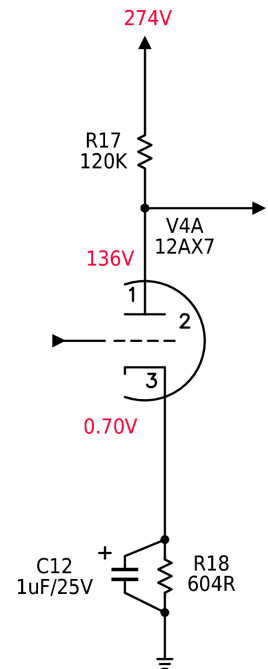


Fig. 16: Stage two treated as if the output load was an infinite impedance.

CALCULATING THE HALF-BOOST FREQUENCY

As mentioned earlier, it is often desirable to apply some bass cut to guitar amplifiers. By using a relatively small cathode bypass capacitor we only partially bypass the cathode, leaving the lower frequencies unbypassed and therefore attenuated or “shelved” to a lower level. The technical name for this is a *low-shelf filter*. It is so named because the frequency response graph (Fig. 17) looks a little like a shelf.

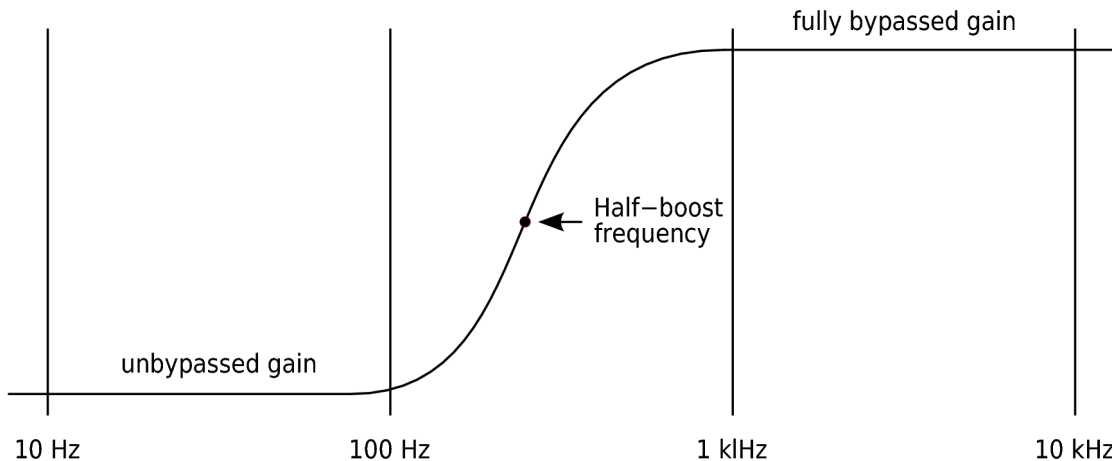


Fig. 17: Each stage forms a low-shelf filter with the half-boost frequency around 200-300 Hz

Merlin Blencowe (“Merlin” on the AX84 bulletin board) and David Ivan James have published a handy formula for determining the half-boost frequency of a plate-loaded triode gain stage. Given values for the plate resistance, the load resistance, the cathode resistance, cathode bypass capacitance, and the amplification factor of the tube, you can calculate the frequency at which the gain is half way between the un-bypassed gain and the fully bypassed gain. This formula is:

$$f_{halfboost} = \frac{\sqrt{1 + \frac{R_k(\mu+1)}{2(R_L+r_p) + \frac{1}{2}R_k(\mu+1)}}}{2\pi R_k C_k}$$

where:

- μ is the amplification factor of the triode
- r_p is plate resistance of the triode
- R_p is the plate load resistance or the effective load resistance ($R_p \parallel R_g$)
- R_k is the cathode resistance
- C_k is the cathode bypass capacitance

Do not let the apparent complexity of this formula throw you. Given what we already know about these amplifier stages, it is a simple substitution problem to solve for the half-boost frequency. Let's work through it using the stage two data from the previous pages. From the data sheet, we know the

amplification factor of the 12AX7 triode is 100. We calculated the 12AX7 plate resistance at the stage two operating point to be 65.2K. The plate load resistor is 120K and the cathode resistor is 604 ohms. The cathode bypass capacitance is 1 μ F. Plugging these values into the formula gives:

$$f_{halfboost} = \frac{\sqrt{1 + \frac{R_k(\mu+1)}{2(R_L + r_p) + \frac{1}{2}R_k(\mu+1)}}}{2\pi R_k C_k} = \frac{\sqrt{1 + \frac{604*(100+1)}{2*(120K + 62.5K) + \frac{1}{2}*604*(100+1)}}}{2\pi * 604 * .000001}$$

Simplifying these values, we get:

$$f_{halfboost} = \frac{\sqrt{1 + \frac{61004}{365000 + 30502}}}{0.0038} = \frac{\sqrt{1.1542}}{0.0038} = 282.72 \text{ Hz}$$

For those readers who are mathematically inclined and wish to know how to derive this formula, please refer to Blencowe, M. and James, D. I. "Choosing Cathode Bypass Capacitors," *AudioXpress*, August 2008, pp. 19-20.

You might want to experiment with different values of C_k in this formula and see how they affect a given gain stage's half-boost frequency. It is also an interesting exercise to study the cathode bypass values of various production amplifiers and see how they compare. Modifying successive stage cathode bypass capacitances is often used to shape the overall frequency response of an amplifier.

INCLUDING THE TONE STACK RESISTANCE IN GAIN CALCULATIONS

What happens to the gain of this stage if we do not ignore the tone stack circuit?

The simplest approximation of the tone stack is to imagine that C8, C9, and C10 behave as short circuits to AC voltages, effectively removing the bass control (VR3) and top half of the middle control (VR4) from the circuit. The capacitors in the tone stack do not actually behave that way. Instead, the frequency of the signal plays a large role in the total impedance of the stack, but this simplification gives an acceptable approximation. Suppose also that the tone controls are set at their midpoints. Applying these assumptions gives the equivalent circuit shown at the bottom of Fig. 18.

The total AC resistance of this approximate circuit is:

$$R_{\text{tone-stack}} = (R_{11} \parallel VR_2) + \frac{VR_4}{2}$$

Now, plug in the values:

$$R_{\text{tone-stack}} = (100K \parallel 250K) + \frac{25}{2} = (100K \parallel 250K) + 12.5 = 83.9K$$

The effective resistance on the plate turns out to be:

$$R_{\text{effective}} = (120K \parallel 83.9K) = 49.4K$$

Plug the numbers into the bypassed and unbypassed gain formulas to obtain:

$$Gain_{\text{bypassed}} = \frac{65.2 K \Omega \parallel 49.4 K \Omega}{1} = \frac{28.1 K \Omega}{625} = 44.9 \quad \text{and} \quad Gain_{\text{unbypassed}} = \frac{65.2 K \Omega \parallel 49.4 K \Omega}{1} = \frac{28.1 K \Omega}{1229} = 22.9$$

As you can see, this is a significant difference from the gain numbers we obtained without considering the tone stack. Additionally, the change in effective load moves the half-boost frequency up a little to 297Hz. Recall, however, that this is only an approximation. Capacitors do not act as dead shorts to AC voltages, but rather as frequency-dependent impedances. So the attenuation from the tone stack is not quite so pronounced.

There are some questions you might want to think about: What is the effect on gain if the controls are set at their lowest points? Or the highest points? What is the effect of raising or lowering the value of the R11?

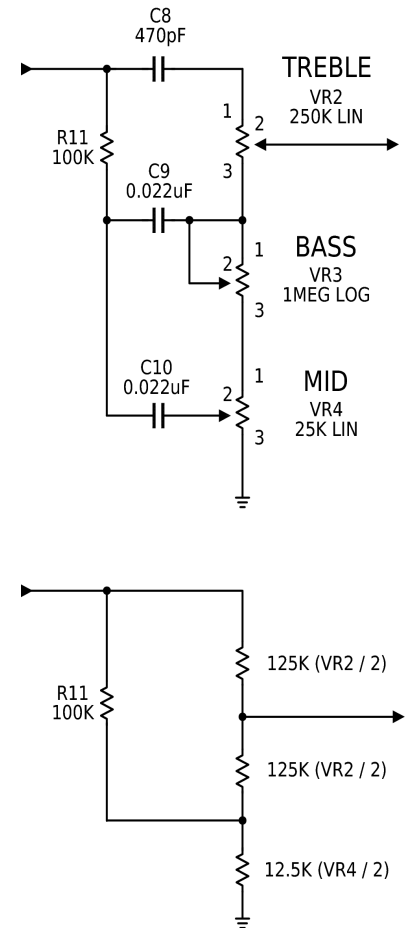


Fig. 18: A model of the tone stack that assumes C9 and C10 behave as AC shorts and each control is centered

tone stack (THE LONG COMPLICATED EXPLANATION)

If you have read many schematics of tube guitar amps, the tone control circuitry should look familiar because many amp companies have used this arrangement of components for their tone controls. The tube amp gurus call this circuit a FMV tone stack, which stands for Fender/Marshall/Vox. The FMV tone stack has been duplicated in hundreds of guitar amps in the last half century.

Technically speaking, the tone stack is a set of passive filters, each with its own -3db point. Another way to think of the tone stack is to call it a passive network with frequency-dependent insertion loss. When a signal is inserted into this circuit, some frequencies are attenuated more than others.

Let's first discuss two types of passive RC filters—high-pass and low-pass.

First the high-pass filter: Imagine you have input and output connections whose signals share the same ground. Wire a capacitor between the input and output connections, and put a resistor between the output connection and ground. This creates an RC high-pass filter that has a -3db cut-off frequency of:

$$F_{cutoff} = \frac{1}{2\pi RC}$$

The -3db point refers to the frequency at which the output signal is 3db lower than the input signal. In a high-pass filter, lower frequencies are attenuated more.

A low-pass filter is simply the opposite of a high-pass. Again, imagine you have input and output connections with a common ground. Wire a resistor between the input and output and place a capacitance between the output and ground. This creates an RC low-pass filter that has a -3db cut-off frequency of:

$$F_{cutoff} = \frac{1}{2\pi RC}$$

Yes, this is exactly the same formula as above. Convenient, huh?

If you feed the output of a high-pass filter into the input of a low-pass filter, you get a band-pass filter.

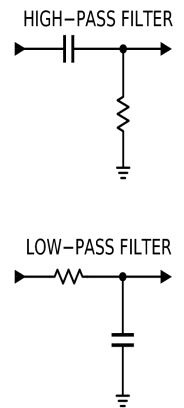


Fig. 19: Filters

Now, let's dissect the P1-eX tone stack, starting with the treble control.

Imagine that R11, C9, and C10 are removed from the circuit. You are left with one capacitor between the input and output (C8), and a set of potentiometers across the output and ground (VR2, VR3, and VR4). The resistances of the potentiometers add up and can be thought of as a single resistor. So we are left with the simple, high-pass RC filter shown in Fig. 20.

Now consider what is going on at VR2. Signals with a frequency greater than the -3db cut-off frequency pass right through the capacitor to the top terminal of VR2. Frequencies lower than the -3db cut-off frequency are noticeably attenuated. Remember how the volume control works? It is a resistive voltage divider. The same thing happens with VR2 (the treble control). When the wiper is closer to the top of VR2, the signal that just made its way through the high-pass filter is attenuated less. Moving the wiper toward the bottom leg of VR2 causes more attenuation of the high-pass filter output.

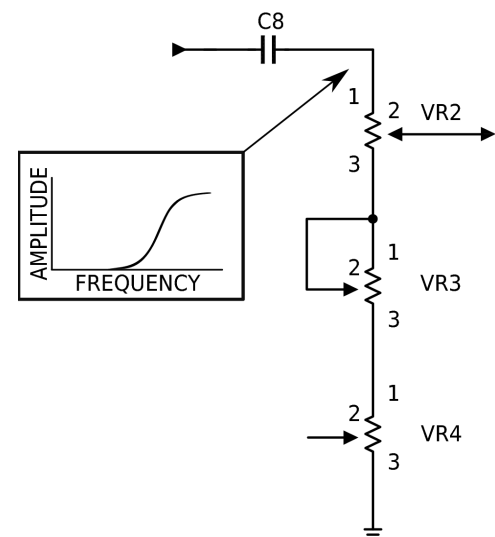


Fig. 20: The treble control in isolation

Take a close look at VR3 (the bass control). As the wiper moves toward the bottom leg of the potentiometer, the resistance of VR3 is reduced. This causes the -3db cut-off frequency to increase as VR3 rotates counterclockwise (moving the wiper closer to the bottom leg).

Imagine that C8, C10, and VR2 are removed from the tone stack circuit, as shown in Fig. 21. What remains is actually two filters in series. R11 and C9 form a low-pass filter, and C9 and the combination of VR3 and VR4 form a high-pass filter. Here again, the -3db cut-off frequency of the high-pass filter depends on the setting of VR3. At the highest bass control settings, the -3db cut-off frequency of the high-pass filter created by C9, VR3, and VR4 is well below the frequency of the guitar's low E string. Lower settings of the bass control raise the -3db cut-off frequency, attenuating the bass.

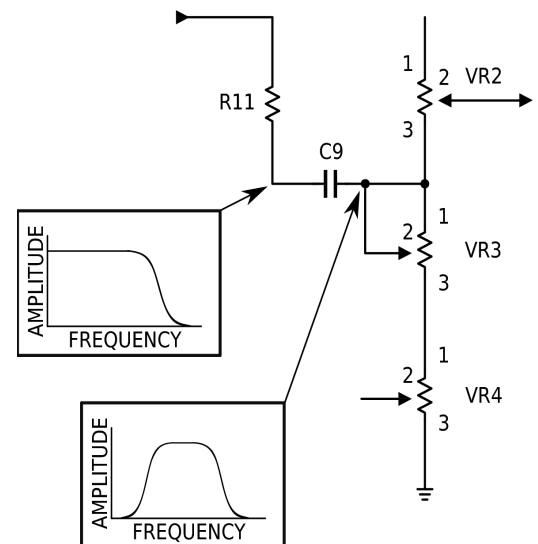


Fig. 21: The bass control in isolation

Moving on to VR4 (the middle control), we have a circuit almost identical to the bass control. Imagine that C8, C9, VR2, and VR3 are removed from the circuit, as shown in Fig. 22. R11 and C10 form a low-pass filter, just like the bass control circuit. C10 and VR4 create a high-pass filter, but with a much higher -3db cut-off frequency than the high-pass element in the bass control circuitry. The difference here is what happens to the output of this combination of filters. Think of it this way: The output of the filters that make up the mid control is present at the wiper of VR4. Move the wiper of VR4 towards the top leg (mid control up), and less of this output is shunted to ground through VR4. Turning down the mid

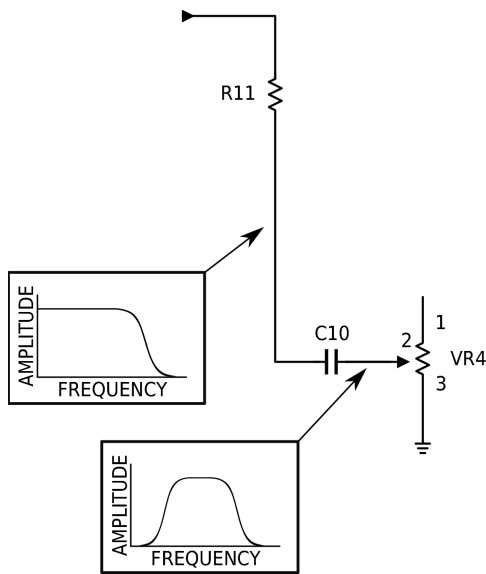


Fig. 22: The middle control in isolation.

control shunts more filtered signal to ground, attenuating the mid frequencies.

Not only does the mid control attenuate mid frequencies, it also attenuates the overall level of the output signal. Consider what happens when VR4 is turned all the way down, essentially shorting one side of C10 to ground. In this setting, C10 effectively shunts most of the signal present at the lower leg of R11 to ground. And for practical purposes, the lower leg of VR2 is also shunted to ground through the combination of C9 and C10, at least for AC signals. Thus, the mid control affects the overall output of the tone stack.

We made a huge assumption here when we ignored some of the tone stack components to analyze the high, mid, and bass control circuits separately. Doing this made it easier to see how each section worked, but we ignored the fact that the filters interact with each other. This interaction is far too complex to explain here. If you are curious about it, you can find a more thorough analysis of the Fender-style tone stack in Richard Kuehnel's excellent book *Circuit Analysis of a Legendary Tube Amplifier: The Fender Bassman 5F6-A*. Excerpts from this book are available at the Pentode Press website:

www.pentodepress.com.

TONE STACK (THE SHORT SUMMARY)

Treble Control (VR2): This potentiometer acts as a balance control between the output of a high-pass filter formed by C8 and the three potentiometers, and the output of the complex filter created by R11, C9, C10, VR3, and VR4.

Bass Control (VR3): This potentiometer sets the lower -3db cut-off frequency of a band-pass filter formed by R11, C9, VR3, and VR4. It also affects the -3db cut-off frequency of the treble control circuit.

Mid Control (VR4): This potentiometer controls the attenuation of the band-pass filter formed by R11, C10, and VR4. It also acts as a variable attenuator for the tone stack output.

VACUUM PENTODES

A discussion about pentode tubes has to start with an examination of the limitations of triode tubes. In the talk about triodes, we learned that inside a triode there are a few metal pieces, separated from each other by a vacuum. Now, consider that a capacitor is made by separating two conductive surfaces by an insulating dielectric. A vacuum is a sort of dielectric. So inside the triode, there are a couple of virtual capacitors, one between the plate and grid, and one between the grid and cathode. These internal capacitances are called inter-electrode capacitances.

Looking at the 12AX7 data sheet, we can find the inter-electrode capacitance specifications:

	Triode Unit 1	Triode Unit 2	
Grid to Plate	1.7	1.7	pf
Grid to Cathode	1.6	1.6	pf
Plate to Cathode	0.46	0.34	pf

The ability of a triode to accurately amplify high frequency signals is limited by this capacitance. We do not have to worry much about this, since a guitar amplifier is not designed to amplify radio frequencies, but the desire to build radio amplifiers in the VHF, UHF and higher frequency bands led researchers to look for ways to reduce inter-electrode capacitances.

Another limitation of triodes is that plate current depends not only on grid voltage, but also on plate voltage. For example, suppose you lower the grid voltage in a triode circuit. As we learned above, lowering the grid voltage results in an increase in plate current. And Ohm's law dictates that the voltage across the load resistor must increase as current through it increases. Now, the supply voltage is a constant, so the plate voltage must decrease as the voltage across the load resistor increases (the two voltages must add up to the supply voltage). Lowering the plate voltage results in reduced plate current; gain is limited in a triode because of this. In a preamp stage, reduced gain presents no problem because we have to discard some of the gain between stages to prevent blocking distortion. However, in an output stage, interaction between plate voltage and plate current is not good—it only reduces the maximum output power of the amp. What was needed was a way to decouple the output power from the plate voltage.

The efforts to reduce inter-electrode capacitance and to increase output power led to the development of the tetrode. A tetrode has an extra element called the screen grid fixed between the grid and plate. The screen grid solves both problems with the triode. Adding an element between the grid and plate reduces the distance between electrodes inside the tube which, in turn, reduce the inter-electrode capacitance. This allows the tetrode to successfully amplify much higher frequencies.

Furthermore, the screen grid allows the plate current to operate independently from the plate voltage. How does it do this? Remember our discussion of diodes and triodes. The electrons streaming off the cathode are attracted to the plate, because the plate has a positive voltage potential compared to the

cathode. You could say that the plate exerts a pulling force on the electrons. In a tetrode, the screen grid is operated at a voltage that is slightly less than the plate voltage. Now, because the screen grid is physically closer to the cathode than the plate, the screen grid exerts more pull on the electrons than the plate. Some of the electrons actually hit the screen grid, but most pass right through on the way to the plate. Thus, plate voltage has very little effect on electron flow in a tetrode. The steady voltage of the screen grid provides a nearly constant pulling force on the electrons. As a result, the grid has almost complete control over plate current, regardless of changes in plate voltage.

While we are talking about electrons whizzing around inside a tube, now is a perfect time to discuss the main limitation of a tetrode. Imagine what happens to an electron inside a tetrode tube. It gets knocked out of the cathode because of the heating effect of the filament. Then, the pull of the screen grid makes it speed away from the cathode. It accelerates past the grid and the screen grid. The short trip is over when it smashes into the plate. This impact actually causes some electrons to be knocked out of the atomic structure of the plate. For every electron that smashes into the plate, two or three electrons are knocked out of the plate.

We're talking about particle physics here! Impress your friends: Tell them that small particle accelerators power your amp.

This phenomenon is called secondary emission. Triodes do not suffer much from secondary emission. Electron velocities in triodes are lower and there is typically little else that is positively charged to attract those stray electrons, so they quickly fall back to the plate electrode. In a tetrode, however, the screen is positively charged, so it tends to attract those electrons as well. This is particularly true when a large grid signal causes the plate voltage to temporarily drop below the screen voltage. Under such conditions, the slow-moving electrons from secondary emission are attracted to the screen grid and do not make their way back to the plate. This reduces plate current and increases screen grid current, reducing gain and clipping the peak of the amplified signal.

To address this problem, researchers at Philips placed an additional element called a suppressor grid between the screen and plate and located very close to the plate, creating the pentode. The suppressor grid is usually connected to the cathode, often internally, which puts it at a voltage potential much lower than the plate. It does not interfere with the flow of electrons from the cathode, because those electrons are traveling at such a high velocity that they pass right through. The slower-moving secondary emission electrons, however, are easily repelled by the suppressor grid straight back to the plate. The effect handily eliminates the problems associated with secondary emission.

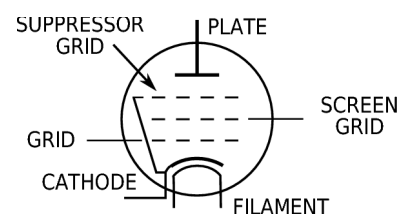


Fig. 23: The pentode

OUTPUT STAGE

The output stage of the P1-eX employs an EL34 power pentode as a cathode-biased, single-ended, transformer-loaded power amplifier. We have already discussed cathode biasing. Single ended describes an output configuration where a tube (or a set of tubes in parallel) is connected to the one terminal of a load and the other terminal of the load connects to ground (often via the power supply).

stage. Because all the current in the tube must flow through the cathode resistor, dividing the cathode voltage by the cathode resistor value determines the cathode current. Cathode current in tube V1 equals plate current plus screen grid current. Looking at the voltages on the schematic, you can see that our screen current is about 10mA (a 10V drop across the 1K resistor R8).

The plate current is the cathode current minus the screen grid current, as follows:

$$I_p = I_k - I_{g_2}$$

Ohm's law tells us that the cathode current is equivalent to the voltage drop across the cathode resistor divided by the value of the cathode resistor:

$$I_k = \frac{E_k}{R_k}$$

Substituting this last formula into the plate current formula shown in the previous formula gives:

$$I_p = \frac{E_k}{R_k} - I_{g_2}$$

To determine the screen current, you measure the voltage drop across the screen resistor R8 and divide that voltage by the value of R8. Using the measurements shown on the schematic, we see there is a 10V drop across R8. That means 10mA of current are flowing through the screen grid. Plugging the voltage across the cathode resistor, the value of that resistor and the screen current into the previous formula gives the plate current:

$$I_p = \frac{27.2}{380} - .010 = .062 = 62.0 \text{ mA}$$

To determine the plate dissipation in watts, you then subtract the cathode voltage (E_k) from the plate voltage (E_p) and multiply the result by the plate current:

$$\text{Plate Dissipation} = (E_p - E_k) \times I_p = (386 - 27.2) \times .062 = 22.2 \text{ W}$$

The data sheet indicates that the maximum plate dissipation for an EL34 is 25W. Assuming the voltages on the schematic are the voltages measured in the actual circuit, the plate dissipation is about 22W. At these voltages, the amp is running at approximately 88 percent of the maximum plate dissipation. In principle, you can run a single-ended output stage at 100 percent of its maximum plate dissipation because this output structure has a relatively constant current draw from zero signal to maximum signal. In practice, you'll want to back it down a bit because tubes aren't perfectly linear. This setting is fine for the EL34.

Do not assume that your amp will behave exactly like this. It is very important that you measure the actual voltages and do the math. Production tubes wander considerably from the published specifications, so it is important to check a tube in circuit and measure its dissipation. For example, an EL34 with a higher gain than the published specifications will draw more plate current for a given grid voltage. It could be damaged if plate dissipation exceeds 25W. To correct for such a situation, you

increase the value of the cathode resistor to decrease the plate current. Different power dissipation in an output stage produces different tonal responses. Ultimately, you may wish to bias by ear to obtain the tone you prefer. This is fine as long as you don't violate the tube specifications. Bear in mind that higher currents generally result in shorter lifespans for output tubes.

OUTPUT TRANSFORMER

The output transformer is one of the least understood components in a tube amp. The construction is fairly simple: two coils of wire wound around a magnetic core. The coil on the input side is called the primary. The coil on the output side is called the secondary.

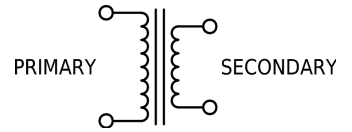


Fig. 25: A transformer

The name transformer is descriptive of its function—it performs a transformation of the voltage and current from the input to the output. For instance, in the P1-eX, the output transformer converts the high-voltage, low-current output from the output tube into a low-voltage, high-current signal suitable to drive a loudspeaker. Some basic transformer formulas follow. As you can see, the turns ratio is equivalent to the voltage and current ratios.

$$\frac{E_{AC-Primary}}{E_{AC-Secondary}} = \frac{N_{Turns-Primary}}{N_{Turns-Secondary}} \quad \text{and} \quad \frac{I_{AC-Secondary}}{I_{AC-Primary}} = \frac{N_{Turns-Primary}}{N_{Turns-Secondary}}$$

The formulas above assume an ideal transformer with no losses, but they are close enough for our purposes here.

In a tube amp, the most important job of the output transformer is to reflect the low speaker impedance into the high impedance load needed by the output tube. A transformer alone has no primary or secondary impedance; it has an impedance ratio. If you look in a tube amp parts catalog, you will see the primary and secondary impedances specified as ratios according to this formula:

$$\frac{Z_{Primary}}{Z_{Secondary}} = \left(\frac{N_{Turns-Primary}}{N_{Turns-Secondary}} \right)^2 \quad \text{or} \quad \frac{N_{Turns-Primary}}{N_{Turns-Secondary}} = \sqrt{\frac{Z_{Primary}}{Z_{Secondary}}}$$

Suppose you buy a transformer with a 5K-ohms to 8 ohms impedance ratio. The turns ratio is:

$$\frac{N_{Turns-Primary}}{N_{Turns-Secondary}} = \sqrt{\frac{5000 \Omega}{8 \Omega}} = \sqrt{625} = 25$$

If you were to use a speaker an impedance of 4 ohms, the impedance reflected back on the primary would be 2.5K-ohms.

$$\frac{N_{Turns-Primary}}{N_{Turns-Secondary}} = \sqrt{\frac{2500 \Omega}{4 \Omega}} = \sqrt{625} = 25$$

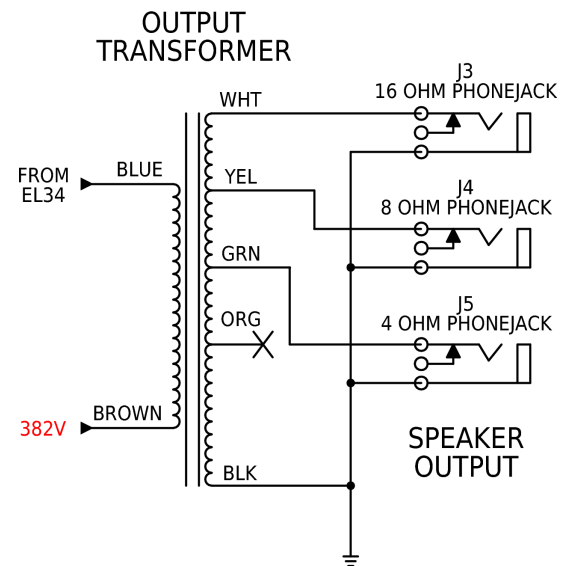


Fig. 26: The output transformer wiring.

The output transformer shown in the P1-eX schematic has a 16 ohms secondary with additional taps at 8 and 4 ohms. Assuming you connect your speaker to the appropriate tap, the reflected primary impedance is 5K-ohms. This primary impedance was not chosen by accident. Rather the value is taken from typical design examples in EL34 data sheet for operating points similar to that used in the P1-eX output stage.

POWER SUPPLY

The power supply in the P1-eX is an example of a full-wave rectifier supply. The power transformer has three secondary windings. The red winding is center-tapped and provides a high-voltage source for the B+ power to the preamp and output stage. The green 6.3VAC winding provides the filament supply. The yellow 5-VAC winding is intended to heat a vacuum tube rectifier. The stock P1-eX design has no tube rectifier and does not use the yellow winding.

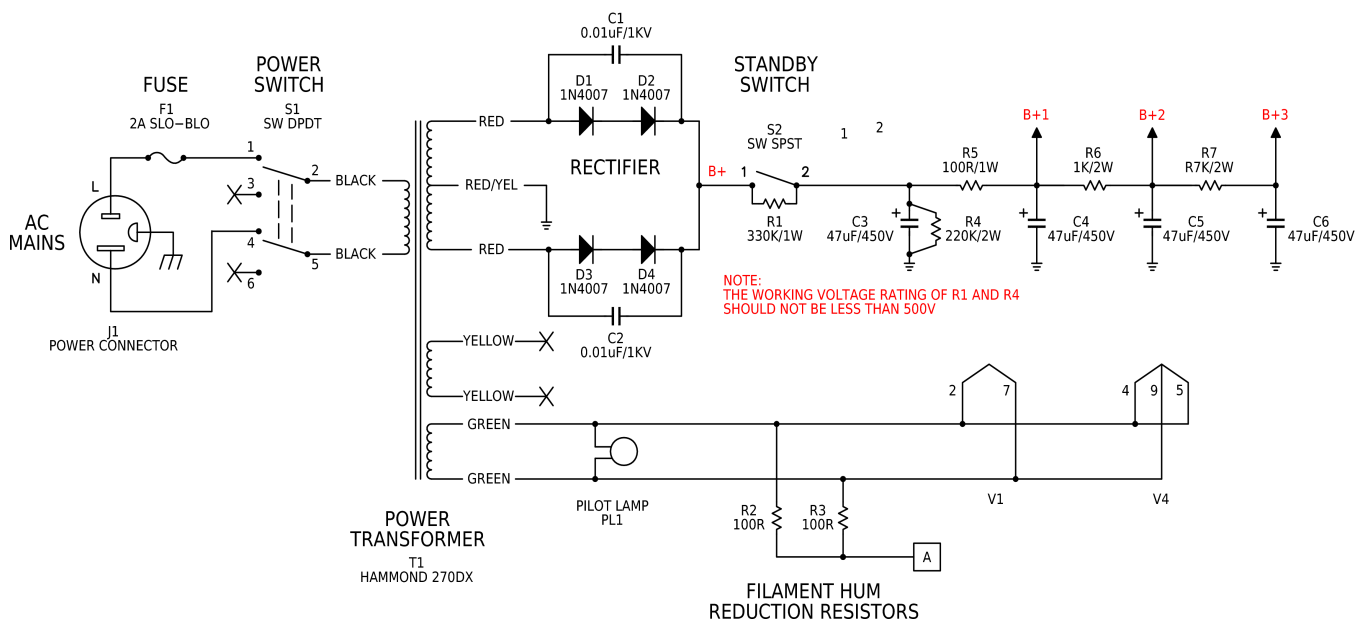


Fig. 27: The P1-eX power supply circuit from mains to plate including filament supplies

First, let's look at the B+ supply. The center tap of the power transformer's high-voltage secondary winding is connected to ground. This ground connection references the secondary signal to zero volts, so it is alternating current around the circuit's zero-voltage point. In the case of the P1-eX, the voltage swings from 0V to +400V, back down to 0V, then to -400V, and back to 0V. In the US, each cycle takes 1/60 of a second. Perhaps you are questioning the value of 400V, because the Hammond transformer specifications say the secondary voltage is 275VAC. Transformer secondaries are generally specified by their RMS AC output voltage. RMS is a measure of voltage delivered into a load over time rather than a measurement of an instantaneous peak voltage. To get the peak voltage you must multiply the RMS voltage times the square root of 2 (approximately, 1.414):

$$V_{peak} = V_{RMS} \times \sqrt{2} = 275 \times \sqrt{2} = 389V$$

Hmm... 389V. So what accounts for the extra 11V? For their "classic" series of transformers, Hammond assumes a line voltage of 115VAC, which was a US standard back in the day. Since then, US wall

voltage has climbed to about 120VAC. That 5V RMS difference on the primary side accounts for a higher peak voltage of about 405V on the secondary side. Subtract a couple of volts for the drop across the diodes and a few more volts for DC losses in the secondary winding, and you have a B+ value of about 400VDC. You'll also find the same proportional voltage increase in the other secondary windings.

Each side of the high-voltage secondary winding connects to the anode of a series string of two 1N4007 silicon rectifier diodes. The capacitors across each diode set snub out some of the high-frequency switching noise that the diodes introduce into the power signal. The cathodes of these two diode strings are tied together to form a full-wave rectifier. Arranged this way, the diodes form a sort of one-way valve so only the positive voltage swings are allowed through. The two out-of-phase AC voltages are turned into one bouncing DC voltage. What does that mean exactly? Consider the following diagram:

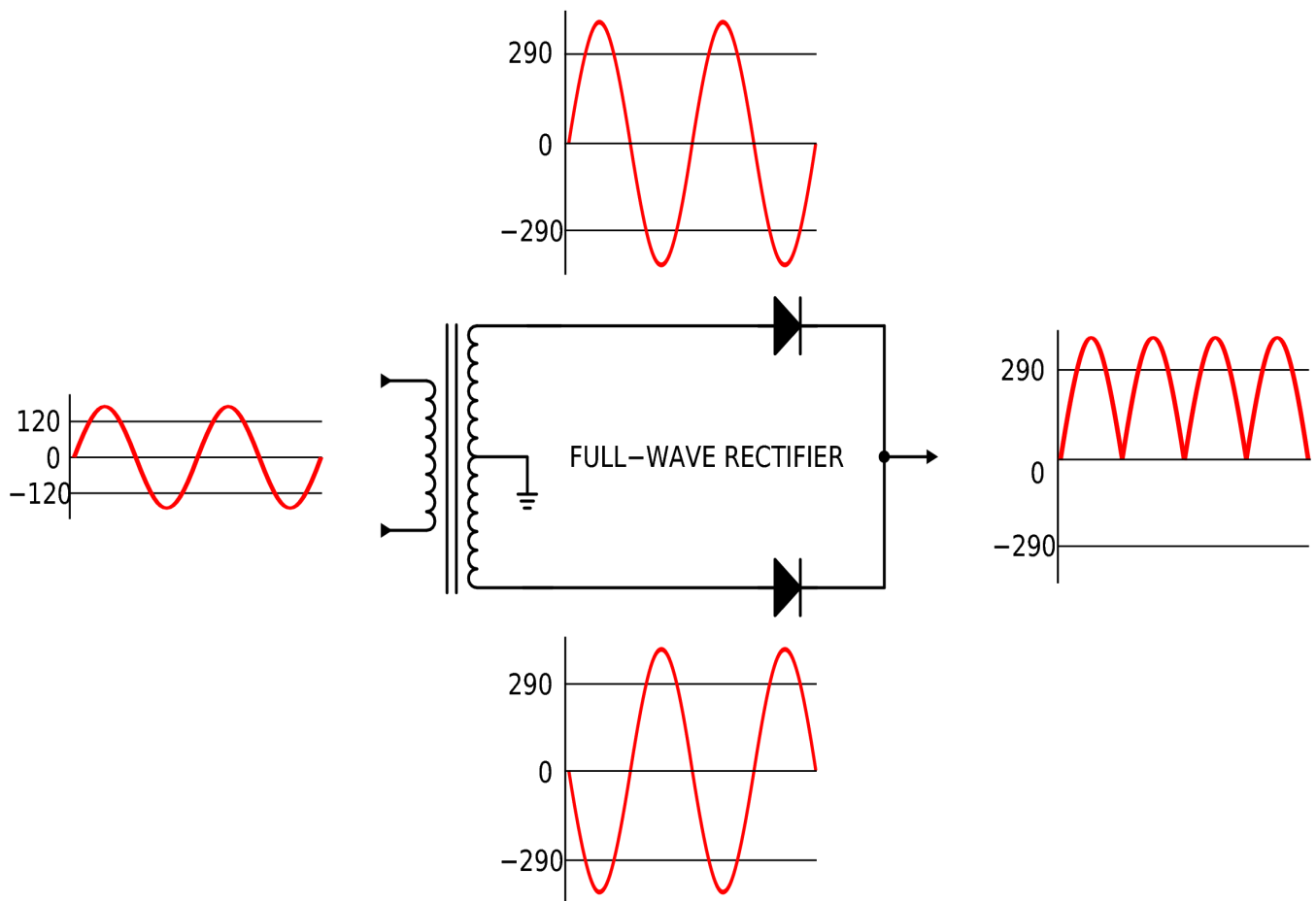


Fig. 28: Rectifying AC into pulsating DC

The line voltage is connected to the primary. The transformer steps it up to about 287VAC. Notice how each side of the secondary winding is out of phase with the other such that when one half of the winding is swinging positive, the other is swinging negative. Each diode allows only the positive swinging half through, so when they combine after the diodes, the signal forms a pulsating DC voltage.

The standby switch immediately following the diodes allows you to delay turning on the B+ to the plates until after the filaments have warmed up. Consider what happens when a tube amp is turned on with the

standby switch closed. The filaments inside the preamp and power tubes are cold, so no current flows from the plate to cathode. With silicon diodes as the rectifier, the B+ voltage jumps to full value almost instantly when power is applied. Remember when we determined the voltages present on the plates of the preamp tubes by multiplying current by the value of the plate load resistor. Well, when no current is flowing through the tube, there is no voltage drop across the plate load resistor. So, shortly after the power switch is thrown, the voltage at the plates rise to full power supply value. The coupling capacitors are subjected to the full B+ value, which is why all coupling caps in a tube amp should be able to withstand the full B+ voltage. Additionally, bias voltages are at 0V, because no current through the tubes means no current through the cathode resistor.

Once the cathode heats up enough, current will begin to flow through the tube. When the amp is first turned on, however, there is full B+ on the plate, and nearly zero bias voltage. This quickly settles as the tubes warm up, but for a short period of time they are stressed beyond their specified limits. The standby switch is intended to address these issues, but it brings its own set of issues to the design. One of the problems is that toggle switches are rarely rated for DC operation at typical B+ values, which means the switch itself will be operated beyond its specifications. Another issue is that running a tube amplifier for long periods without any voltage on the plates is also harmful to the tubes. The *Radiotron Designers Handbook*, 4th Ed. recommends no more than 15 seconds on standby. For this reason the 330K resistor R1 was added. There is much debate about whether the standby switch adds any value to the design.

The bouncing output voltage at the cathodes of those two diodes adds a tremendous amount of 120 Hz hum to the B+. This hum must be filtered out. C3 serves as the first step in that filtering process. Capacitors used for this purpose are called filter capacitors.

R4, which is wired to parallel C3, is the [bleeder resistor](#). It is a safety device that quickly drains the current stored in the filter capacitors when the amp is switched off. Some amplifiers also use a bleeder resistor to provide a slight voltage regulation effect by ensuring there is always some load on the power supply. R4 does not provide much regulatory effect because its current draw is very small.

After the first filter cap, there is an RC low-pass filter built from R5 and C4. These two components smooth the B+ voltage even more, and drop the voltage slightly. The output of this filter supplies the main B+ to the output stage. Downstream from this first low-pass filter is another RC filter built from R6 and C5. R6 lowers the voltage a tad, and C5 smooths the voltage. The output of this filter supplies power to the screen grid of the EL34. Finally, another RC filter is built from R7 and C6. The output of this filter supplies power to the preamp section.

There are two primary reasons why the power supply is designed with successive stages of filtering. The first is that you want the preamp power supply to be very quiet. In order to function properly, the first stage of an electric guitar preamp must be sensitive to very small signals. It doesn't care whether the source of those signals is a guitar pickup or the power supply. Any noise or hum on the power supply ends up being greatly amplified. The second reason for multiple stages of filtering is to decouple the power stages from each other. Without this decoupling, a low-frequency feedback path exists from the power stage to the preamp stages. Low-frequency oscillation can start via this path and cause the amplifier to make a putt-putt-putting sound known as motorboating.

FILAMENT SUPPLY

The 6.3VAC winding of power transformer supplies power to the tube filaments. When designing an amp, it is important to select a power transformer that can provide the current required by all the filaments. The Hammond 270DX can supply more than enough current for this little amp.

If you have examined schematics of various tube amps, you may have noticed 100-ohm resistors connected between each side of the filament supply and ground. These resistors provide a ground reference for the filament voltage. It is important to provide a reference point for the filament supply, but there is a problem when this reference is the same as the amplifier's ground reference.

To understand this, we have to review some tube theory. Remember that a heated cathode emits electrons. In fact, the cathode has a coating of material that is very good at spewing off electrons when heated. Even though the filament does not have this same coating, it does emit a field of electrons during normal operation. Because the filament is connected to 6.3VAC, this field has a 60Hz signature. For at least part of each 60Hz cycle, it has a potential that is less than the voltage present on the cathode in our 12AX7 stages. During this time, the cathode attracts some electrons from the field around the heater, which in turn are mixed into the signal the grid and passed on to the plate. The end result is that power supply hum is introduced into the preamp stages.

The P1-eX design uses a clever technique to avoid this problem. Instead of having the filament supply referenced to ground, it is referenced to a positive voltage taken from the top of the cathode resistor on the EL34, which is approximately 25V. This elevates (or biases) the heater voltage sufficiently above the preamp tube's cathode voltage to prevent heater noise from entering the preamp. An alternate way to elevate the heaters is to use two series resistors across C4 to form a voltage divider with a tap to obtain 10-50VDC heater reference.

That concludes our discussion of the power supply.

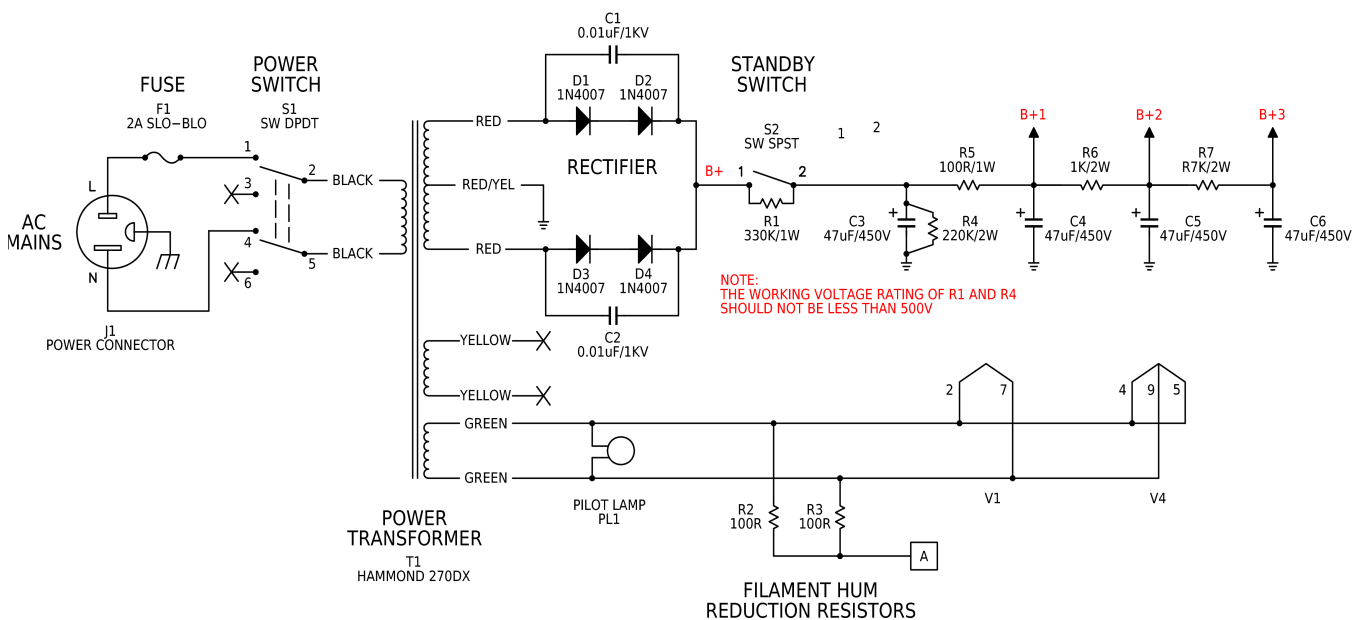


Fig. 29: The P1-eX power supply revisited

A FEW FINAL WORDS

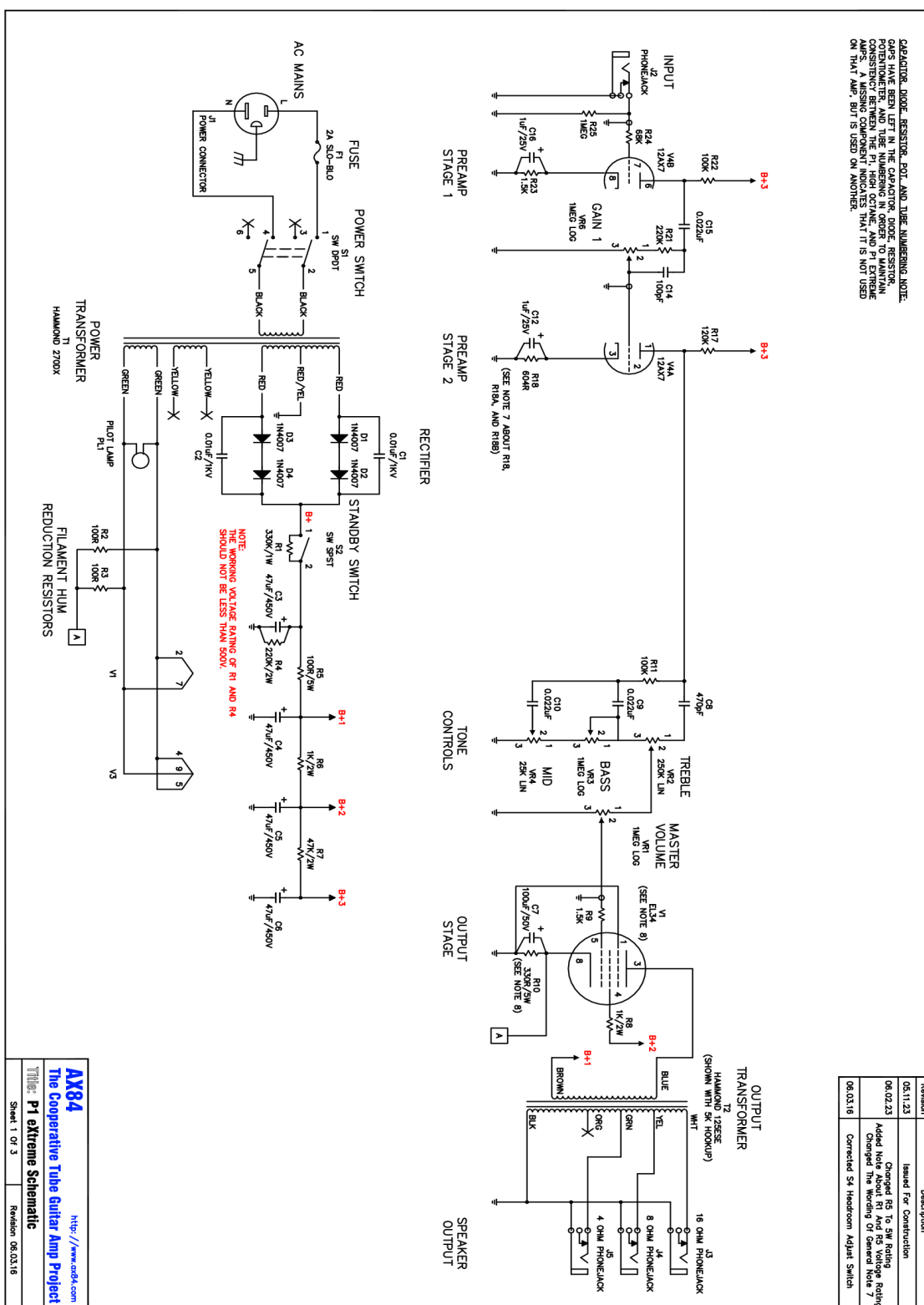
Building your own amplifier is an exiting and rewarding project. Nothing quite compares to the satisfaction felt the first time you power up your amp and it works. This document has offered a basic overview of the ins and outs of the P1-eX design in the hopes that it will demystify what goes on inside our favorite household particle accelerator.

A note about safety: The voltages inside an electric guitar amplifier are potentially deadly. Please take time to read up on how to work safely before attempting to build any electronics projects. Many good resources on this topic can be found for free online. Two sources of note are the Drifter Amps guide, found at <http://www.drifteramps.com/safety.html> and Aiken Amp's safety guidelines at <http://www.aikenamps.com/Safety/Tips.html>. This information could save your life. Read it!

This overview has provided just a taste of the vacuum tube amplifier design process. It is not the end of the story. There are many other design aspects to be explored: pentode stages, direct-coupled stages, push-pull output stages, transformer coupled stages, cathode followers, alternate passive tone controls, active tone controls, feedback amplifiers, multi-channel amplifiers, switching techniques, effects loops, various on-board effects such as reverb, tremolo and vibrato, and host of other topics. We aren't going to try to cover those at all. We hope that this brief document has given you a solid baseline knowledge about tube amplifier theory. You will find that the more you learn about the topic, the more there is to learn. That is a good thing. There is a wide universe of circuit variations to try and the tonal possibilities are nearly limitless.

As you study other amplifier designs, remember the concepts you learned here: how to construct load lines, how to analyze the gain and frequency response of each stage in an amplifier, and how the coupling between stages causes them to interact. Those are powerful tools for understanding any vacuum tube amplifier. You are embarking on a thoroughly enjoyable quest. Never stop learning, be safe, have fun.

Appendix 1: P1-eX REV 06.03.16 Complete Schematic



Appendix 2: 12AX7 WITH A 75K OHM LOAD AT 250V

As promised, here is the load line for a 12AX7 with a 75K-ohm load and a B+ of 250V.

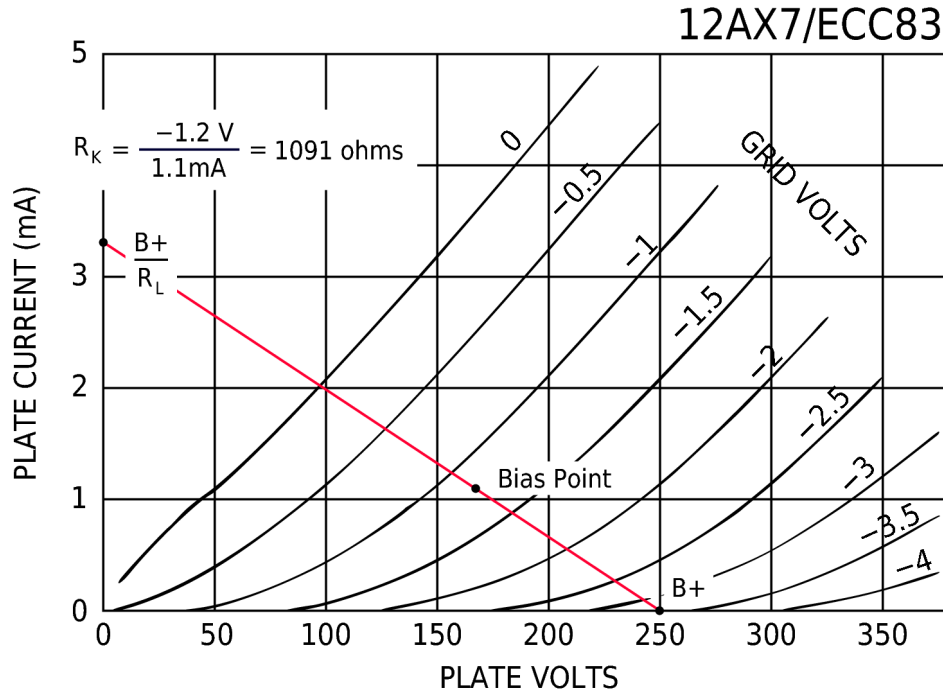


Fig. 30: A steep load line results in reduced gain at generally higher currents

The 75K-ohm load is steeper than a 100K-ohm load. With a B+ of 250V and a bias point at -1.2V on the grid, the zero-signal plate voltage is about 170V at 1.1mA. Gain is somewhat reduced by this change because the lower plate resistance does not allow for as large an output signal swing. To verify this, simply replace the plate load and cathode resistances in the gain calculations:

$$Gain_{\text{fully bypassed}} = \frac{70K\Omega \parallel 75K\Omega \parallel 1.2M\Omega}{\frac{1}{1400\mu MHOS} + 0\Omega} = 49.2 \quad \text{and} \quad Gain_{\text{unbypassed}} = \frac{70K\Omega \parallel 75K\Omega \parallel 1.2M\Omega}{\frac{1}{1400\mu MHOS} + 1.1K\Omega} = 19.4$$

Additionally, the steeper load line with a grid bias a little closer to 0V lowers the half-boost frequency from 199.4 Hz for the stock stage-one preamp, down to about 192.4Hz:

$$f_{\text{halfboost}} = \frac{\sqrt{1 + \frac{R_k(\mu+1)}{2(R_L+r_p) + \frac{1}{2}R_k(\mu+1)}}}{2\pi R_k C_k} = \frac{\sqrt{1 + \frac{1100*(100+1)}{2*(70.6K+70K) + \frac{1}{2}*1100*(100+1)}}}{2\pi * 1100 * .000001} = 192.4Hz$$

Appendix 3: Abbreviations and Symbols

Symbol	Meaning
E	Electromotive force measured in volts (V)
I	Electric current measured in amperes (A)
R	Electrical resistance measured in ohms (Ω)
mA	Milliampere, a unit of electrical current equivalent to 1/1000 of an ampere
μ	The amplification factor of an vacuum tube electrode
g_m	Mutual conductance between the grid and plate of a triode
$r_p (r_a)$	Dynamic plate (anode) resistance
$R_L (R_a)$	DC plate (anode) load resistance
R_k	DC cathode load resistance
R_g	DC grid resistance
G	Gain of an amplifier; that is, the ratio of output voltage to input voltage
V	Volt, the unit of electromotive force
Ω	Ohm, the unit of resistance

CREDITS (Dave Sorlien)

This document contains material from various posts from rec.audio.tubes and alt.guitar.amps by the following authors: Duncan Munro, Ned Carlson, Anonymous.

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CREDITS (Stephen Keller)

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While many people have contributed to the development of this amplifier, it is fitting to mention and thank Phil Rowley (aka Zaphod), who designed the original AX84 P1-eXtreme.

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