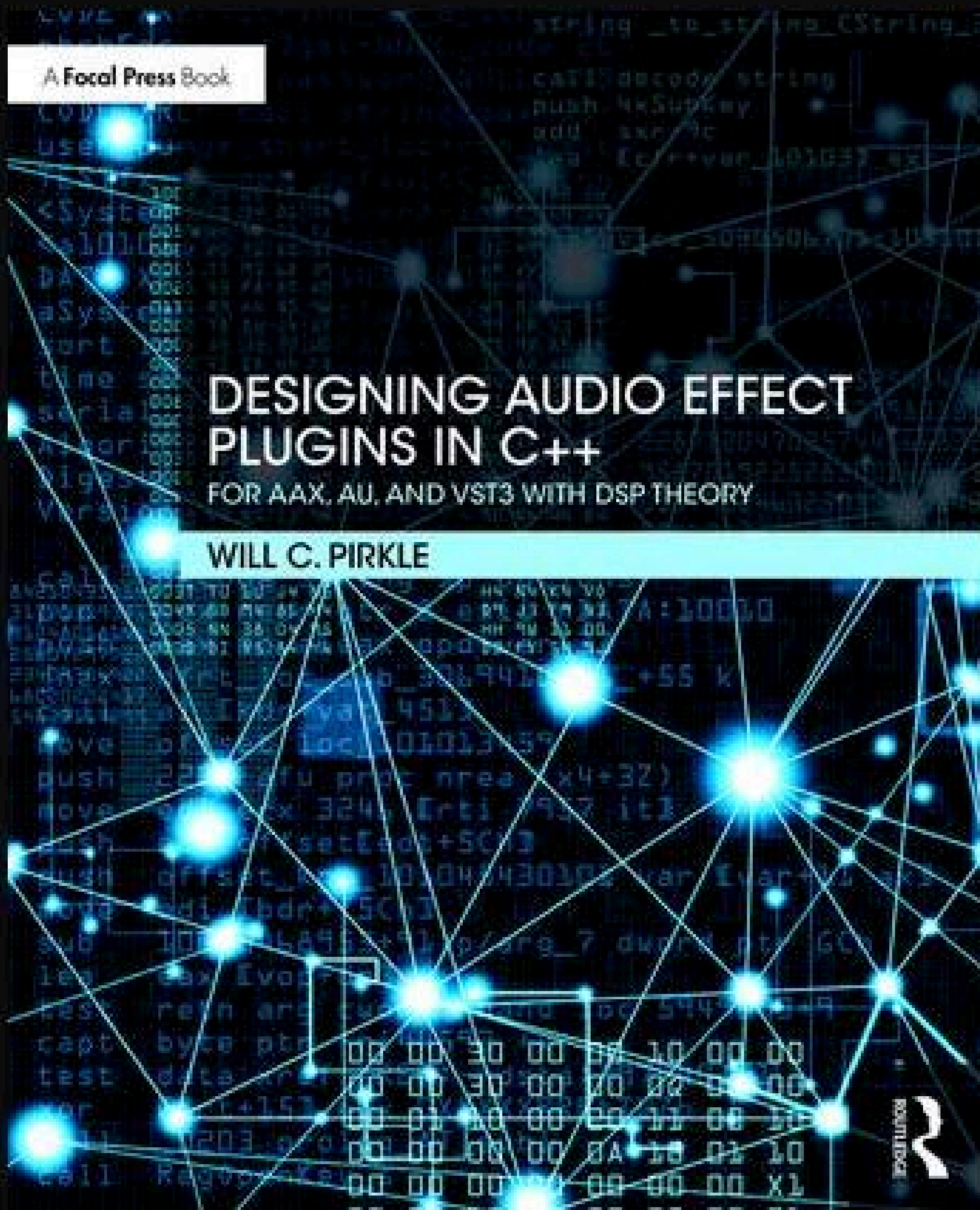
Addendum A19
Vacuum Tube & Distortion Emulation
Part 2



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May 14, 2019

Addendum Preface

This addendum accompanies Will Pirkle's *Designing Audio Effects Plugins in C++*, 2nd Edition. It specifically addresses material that was removed from Chapter 19 on nonlinear processing. Due to space limitations, the vacuum tube chapter had to be cut short and the projects slimmed down to simple wave shaping and filtering. I almost removed the entire chapter; however there has been demand for this kind of information.

The reason I removed the material has to do with a special situation that happens with vacuum tubes in high-gain modern guitar amplifiers. In these designs, the engineers purposefully overdrive the inputs to the tube devices in order to create the modern distortion that is currently popular. In contrast, non-guitar tube-amp circuits are usually designed to stay within the proper operating conditions so that their inputs are not purposefully overdriven.

When you overdrive the input to most amplifying devices like transistors or op-amps, once you've exceeded the operating limits, no more amplification will occur. The device simply won't amplify the signal outside of its input range, and the signal is hard-clipped at that point. For a vacuum tube, this is true for the ***negative*** part of an over driven input signal. But when the positive portion of the input goes outside the limit, everything changes and the tube's behavior becomes very dynamic. Cascading multiple overdriven stages together with differing amounts of gain between stages produces a very complex harmonic effect where the harmonic component amplitudes are directly linked to the signal amplitude. This situation is called ***grid-conduction***.

As I went further into the operational notes on grid-conduction it became more and more clear that in order for someone to understand how the objects emulate these tubes, they would need not only a background in analog electronics, but also some amount of tube circuit theory – at least enough to understand the fundamentals. *Without this information, my objects and code would appear to be almost like magic to you and there would be no way for you to personalize them and make them your own.* Once I added the tube theory, the chapter ballooned out to a ridiculous size and I was forced to remove it. So, that is why the chapter is organized the way it is, and that is why I wanted to make sure you have this addendum to accompany the book.

Modeling: Simulate or Emulate?

Right up front we have to make a choice as to the fundamental approach to a vacuum tube algorithm. There are two main paths to choose: *simulate* and *emulate*. In tube circuit simulation you use SPICE techniques to model the tube as an electronic circuit. In some cases this may involve iterative solutions and methods. Peavey's Revalver ® uses this approach that allows the user to specify exact component values for resistors, capacitors, etc… with excellent results. A downside for teaching is that in order to understand it, you not only need to know audio electronics and tube circuit design, but also circuit component SPICE simulation techniques.

Modeling via *emulation* has an equivalency with the perceptual reverb modeling, where you make no attempt to solve a set of equations that describes the system and instead you analyze the reverb and model it with structures that emulate sound waves echoing of off surfaces and the like.

For an interesting read concerning this kind of modeling, check out James Gleick's book *Chaos Making a New Science*. There is a section about modeling the human eyeball as it moves back and forth in the eye socket. Since it is a bag full of fluid, we have fluid dynamics at work, and since there are muscles attached at various points that pull in various directions, we have a physics problem. One approach is to develop a set of differential equations that describes the system, then setup an initial condition and solve the equations simultaneously. That would be a direct or simulation approach. A second solution to the problem is to model the eyeball moving back and forth as a ball rolling down an inclined plane that you can tilt from left to right. The 'eye' ball will roll down and swerve from left to right as you tilt the inclined plane. The path of the ball represents the motion within the socket. This solution discards the notion of solving simultaneous differential equations and provides an elegant solution.

This addendum will use the emulation-model approach for vacuum tubes. This involves first understanding how tube circuits are designed and operate. Then, we analyze what happens to the tube with an overdriven input and try solutions that model that system. I have multiple patents attached to the appendix of the document. A few of them involve *analog* modeling of vacuum tubes with analog circuit components. These are invaluable, as they must teach the modeling techniques that specifically address key vacuum tube functionality under different conditions. I have patents from Peavey, Scholz R&D and Yamaha in addition to an independent patent application regarding Class-B operation attached here.

The emulation techniques that I am currently using are more aligned with the Yamaha techniques that we will discuss. But in my mind, I see a connection to the Peavey TransTube® patents as well, since these are also excellent emulation modelers that model the same phenomena using analog components.

I worked very hard on the 2nd edition of *Designing Audio Effects Plugins in C++* and I wanted to make sure that this information was also available to the book readers to provide a more complete work.

All the Best,

Will Pirkle
14-May-2019

Contents:

There are two chapter plugins based on multiple chapter C++ effect objects. In order to distinguish between these new vacuum tube C++ objects and the ones in the book, I am using the more European name “valve” in the object names. In this Addendum the two main tube objects are *ClassAValve* and *ClassBValvePair*. The two plugin-projects are:

1. ***SuperSaturator***: a distortion box emulation specifically for guitar

2. ***WickerAmpCombo***: a complete tube guitar amp simulation, from input to speaker, with reverb; includes a 4-triode Class-A preamp and a 2-pentode Class B power amp all done with emulation modeling – a guitar amp tone stack model is included

A19 Addendum: Vacuum Tubes and Distortion Modeling

Part I: Basic Tube Circuit Theory

In this first section, I'll go over just enough tube theory to get you up and running reading tube schematics and figuring out how the FX book code and projects fit within the realm of the basic low distortion tube amplifiers. The most important constraint for this first section, which was the same constraint placed on the book code and projects was:

> CONSTRAINT: The inputs to the tube devices are always kept within the legal range of values that the design dictates. In these circuits, we may generate significant harmonic distortion, but we do not overdrive the tube device inputs nor do we attempt to clip the signal waveforms.

If you think about it, there are many types of audio devices and FX systems that operate within these constraints:

- microphone preamp
- tube-based EQs
- audiophile tube power amp
- compressors and limiters that include tubes
- other FX
- early amplifier designs

The nonlinear chapter in the FX book was aimed at these systems specifically as opposed to modern high gain guitar amplifiers in which the designers purposefully overdrive and overload various tubes within the circuit. This is because high-gain guitar (and bass) amps are designed with a very different purpose than the rest of the devices that include tubes. This means that guitar and bass amps represent a niche *sub-se*t of the tube audio circuits.

The second section of this addendum will focus specifically on high-gain guitar amplifiers and pre-amplifiers. This can then be extended to bass amplifiers as well, through lowered distortion levels and changes to the frequency responses of the sub-circuits.

A19.1 Basic Vacuum Tube Theory

Although there seems to be an argument over who invented the light bulb, we do know that Edison tried to *improve* the invention. When you run an electrical current through fine wire, the wire will glow, or give off light, for a short time just before it catches on fire. If you put the wire in a vacuum, it could be put in the glowing state and held there. Without oxygen, the wire (called a filament) cannot catch on fire. In the early light bulbs, the bulb would eventually expire because the vacuum was not perfect, but not before the inside of the glass bulb was covered with black soot – the "ashes" of the filament. The black soot caused the bulb to dim, and be less and less useful. The dimmed bulb was still giving off light, but it was hidden behind the layer of dark powder. Edison wanted to extend the life of the bulbs by eliminating, or at least delaying, the appearance of the soot.

The ashes were being propelled away from the filament because heating metal causes it to give off electrons - this is now called the *Edison Effect* or *Thermionic Effect*. When the metal reaches a certain temperature, electrons migrate to the surface as if to "boil off of it" in a similar kind of manner that steam boils off of the surface of water. This produces an electron cloud called a space charge. The ashes Edison saw were catching a ride with the electrons in the cloud, and winding up on the inner surface of the bulb. Edison added a metal plate to the inside of the light bulb, and attached it to a second lead (wire) coming out of the bulb. He then applied a positive charge to the plate (+) and then the plate was electron-deficient. When the electrons began boiling off of the filament, they, along with the soot, were attracted to the plate instead of the inner surface of the bulb. This represented a controlled flow of electrons in a vacuum. In effect, the plate's positive potential sucked up the electron space charge allowing *even* more electrons to migrate up and boil off of the filament's surface. Figure A19.1(a) shows the conceptual schematic diagram for this vacuum tube.

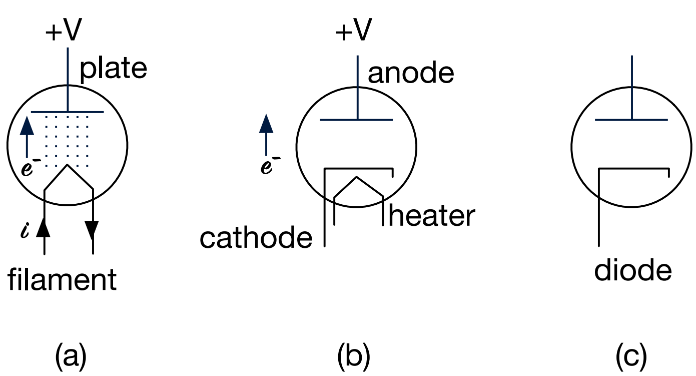


Figure A19.1: (a) a conceptual diagram of the tube; the filament is at the base, represented by the "v" shape, while the plate is at the top and (b) the modern tube uses indirect heating and a separate cathode; in (c) the heater is left out of the symbol for easier reading and clarity of schematics

When a potential is connected across the filament, and a positive charge applied to the plate, current flows from plate to filament. This arrangement is called *direct heating*. Note that the definition of current (i) and the direction of electron flow are reversed, so electrons flow from the filament to the plate, and current flows from the plate to the filament. This is simply the way they are defined.

Eventually, *indirect heating* was implemented. In this version, the filament gave way to a combination of a metal plate called a *cathode* and a heating-only filament called the *heater*. The heater in a tube normally runs off of a separate, low voltage AC or DC signal. The filament's job is simply to heat the cathode to a certain temperature. The heater heats up the cathode, which gives off electrons as a result of thermionic emission. In this case, if the plate voltage is held at 0V, no electrons flow from the heated cathode. As the plate voltage increases, more and more electrons are drawn off of the cathode. This makes the current flow predictable, and basically controllable (either current flows or it doesn't). The second plate is called the **cathode**. In an electrical system, a cathode emits electrons. The first plate is sometimes called the **anode**, but most people call it the plate. The cathode and anode are called electrodes, and a tube that has only these two connections (plus, the filament, of course) is called a **diode** – two electrodes.

A19.2 The Diode

The diode symbol is shown in Figure A19.1 (c) as it usually appears today, using an indirectly heated cathode. The plate must be positively charged with respect to the cathode for current to flow. This means that if you applied a bipolar AC voltage to the plate, current would only flow during the positive (+) portions of the waveform. Current will not flow during the negative (-) portions of the waveform since the plate would not attract the electrons off the cathode. In this manner, the diode behaves like a one-way valve, allowing only positive current to flow. Fleming invented the diode valve in 1905 – many people use the term "valve" for the vacuum tube, especially in Europe while Americans generally refer to them as "tubes."

A19.2.1 Half-Wave Rectification

The half-wave rectifier is just about the simplest tube circuit there is. Its job is to only allow positive voltages, or positive portions of an AC waveform, to pass through. It is the quintessential valve circuit because it behaves as a one-way (positive current only) valve. Figure A19.2 shows how the diode creates a half-wave rectifier from a bipolar sinusoidal input signal. The resistor load is used to complete the circuit, and is attached at the effective output of the diode.

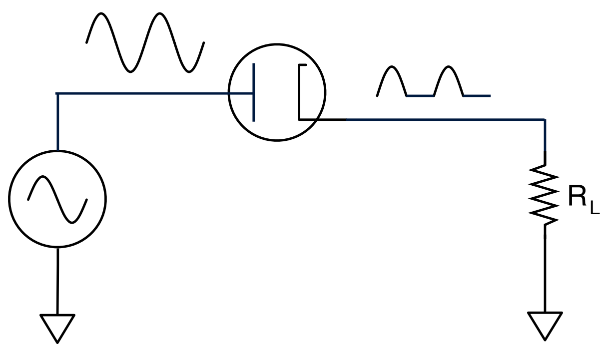


Figure A19.2: the diode used as a half-wave rectifier; since only positive current can flow, only the top portions of the waveform are transmitted through the tube; notice that the details of the heater are omitted as is a common practice

A19.2.2 Full-Wave Rectification

The full-wave rectifier is designed to invert the negative portions of the input signal, causing them to be positive portions, then combine them with the other positive portions, and pass these through the output. A full wave rectifier can be designed with either four diodes, or two diodes and a transformer. The latter is the most common way to implement a full wave tube rectifier. The two diodes are arranged so that their cathodes are connected together. This is such a common arrangement that a tube was designed specifically for this purpose. It is a dual-diode tube, but most people call it a *rectifier tube*. To create the inversion, a transformer can be used. The AC signal across the primary (non-tapped) coil appears as two signals across the secondary (tapped) coil.

Using the full wave rectifier, you can create a complete power supply that converts the 60Hz $120V_{AC}$ signal coming from the wall outlets to a constant 250V DC supply – it looks like a 250V battery for the components it powers. Figure A19.3 shows a complete full wave tube rectifier and power supply. The rectified signal is filtered through R_1, C_1 and C_2. Note the multi-tapped transformer that is used; two of its taps are for filaments/heaters. The output of the filter is applied to the tube plates in the rest of the circuit. Sometimes simple voltage dividers are used to create multiple DC supplies from the original DC output signal.

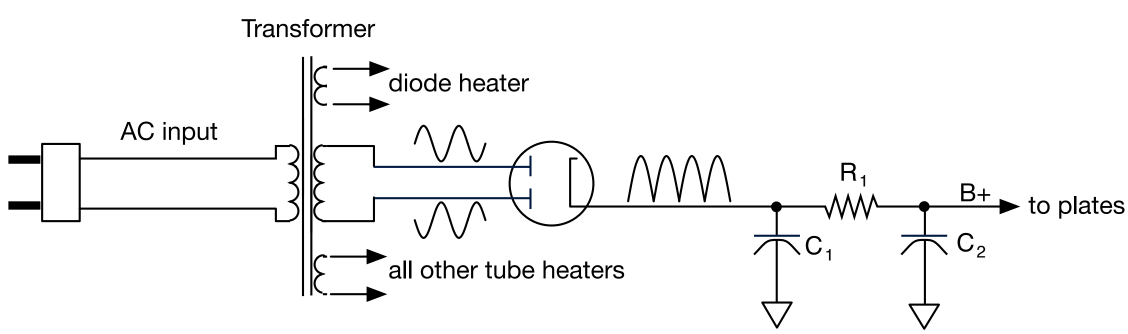


Figure A19.3: a simple tube power-supply; note the rectifier tube with double-anodes (plates) that drive a shared cathode

The final power supply output is used as the *plate voltage* supply for the rest of the amplifier. The name of this value differs from region to region. In some places it is named **HT** for High Tension. In others it is called **B+** while others call it $\mathbf{V_{PLATE}}$, $\mathbf{V_P}$, or just **V+**. Here I will use B+ when describing this value and V_P will specifically indicate the exact voltage at the plate itself, and not at the other end of the plate resistor (the B+ end).

A19.3 The Triode

The problem with the diode valve is that the current is either on or off. The control of electron flow lies purely in the plate. As the plate voltage becomes more positive, more electrons flow from the cathode. What if you could fix the plate voltage at a constant level, to create a constant pull on the electrons, then somehow control the amount of electrons that went from the cathode to the plate? This would be more like a real valve, which has a handle you turn to make more or less water flow through it. In this way, the current flow could be accurately controlled. The idea is to add a fine mesh of wires between the cathode and anode, kind of like a metal screen. Then, connect this screen to a third electrode. This device is called a **triode**. The screen is called the **control grid,** or just grid, and is the current controlling portion of the device. If a negative voltage (with respect to the cathode) is applied to the grid, it will set up a negative electric field between the cathode and plate, so electrons will not be as attracted to the plate. You should think of this as the cathode having less of a pull on the electrons from the cathode.

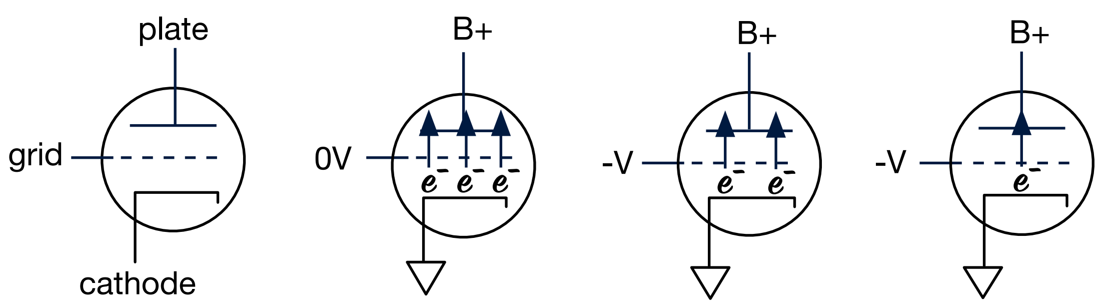


Figure A19.4: the triode consists of an added grid electrode; with 0V at the grid, electrons flow freely from cathode to anode – this is called *saturation*; as the grid to cathode voltage is made more negative, electrons are repelled and held down on the cathode; when the grid to cathode voltage is low enough, no electrons will move and current will stop – this is called *cut-off*

The plate has a fixed positive charge applied, while the cathode, with its abundance of electrons is negatively charged. When a negative voltage is applied to the grid, fewer electrons flow to the plate. This is shown schematically in Figure A19.4 with the vertical arrows representing moving electrons. Because the grid is negatively charged, some electrons do not make it through. One way to think about it is that the grid repels the electrons, so fewer of them flow. The electrons that do make it through flow through the holes in the screen mesh.

At some negative voltage, no electrons will flow to the plate. This condition is called cut-off. When the grid voltage is 0V, the maximum amount of current flows since there is no hindrance from the grid. The maximum current flow condition is called saturation. For a given plate voltage, each tube will exhibit a cut off voltage (the grid voltage required for zero current flow) and a saturation voltage (0V).

This is the fundamental principle behind audio amplification: use a low voltage, low current AC audio signal to control a much larger DC voltage or current. This larger DC voltage/current changes in proportion to the audio control signal, creating an AC voltage/current that is an amplified version of the original. The large DC voltage/current is supplied from the DC power supply. As you just saw with the tube rectifier circuit, the original power source is the AC 120V-60Hz wall outlet voltage/current applied to a rectifier and filter. So, the audio you hear coming from a power amp starts out as a high power 60Hz sine-wave. A vacuum tube can be designed to contain more than one tube component, as you saw with the rectifier tube, which contained two diodes. Many triodes are packaged similarly – two independent triodes inside of one tube. Probably the most common triode in the world is the 12AX7, which is actually a twin triode. On schematics, half a twin triode is shown as a triode with only half (or part) of the circle that represents the glass tube. Any time you deal with a 12AX7 schematic, you will see these half-triodes. Remember that this is only a packaging trick. The two triodes are completely independent.

A19.4 The Triode Class-A Preamp

A full discussion of triode preamp design is outside the scope of even this addendum, but fortunately you can find tons of information out there. From a plugin standpoint, we are interested in mimicking the behavior of the tube. One way to do this is with the wave digital filter approach. Another is by using a SPICE simulation approach and trying to calculate the instantaneous current or voltage in the devices or in the signal connections between the devices. No one approach is likely to be the perfect version, so you should try to learn as much as possible about the final result, and then apply engineering problem solving to make your own models and versions.

The basic idea behind the triode preamp is:

- Attach a resistor between the plate and the B+ supply; this is called the plate resistor or R_P
- Use the audio signal as a control voltage on the grid so that it modulates the flow of current in the tube
- The resulting modulated current must run through the plate resistor, dropping its signal across the resistor as a voltage
- We then tap the output at the junction between the plate and plate resistor to extract the amplified signal

In the following sections, I'll take you through a triode preamp design, using one of several approaches. Figure A19.5 shows a basic triode preamp module and we'll discuss the components and how the device works.

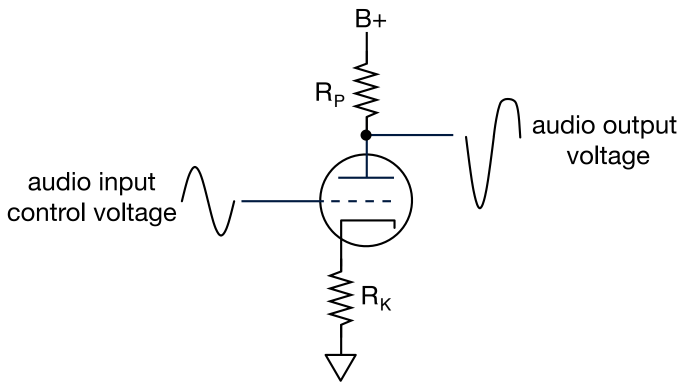


Figure A19.5: a simple triode preamplifier showing only the necessary components to understand how the output signal is generated; the output signal is purposefully drawn in an exaggerated manner to show that the amplification is nonlinear

In Figure A19.5, you can see two resistors. The plate resistor R_P is the load resistance for the output signal to generate across. The cathode resistor R_K is used to set the bias point for the audio input signal. Remember that the grid-cathode voltage must be either 0.0 (full on, saturation) or a negative value –V (to attenuate current flow all the way down to cut-off). The cathode resistor R_K is used to self-bias the triode so that a bipolar input at the grid will appear to swing between 0.0 and some maximum negative cutoff value –V. This is one way to negatively bias the input signal so that it acts as a proper negative control voltage.

In the book projects as well as the first part of this document, we will assume that the input signal into any of the tubes is constrained to fit within the correct range for the triode design, that is, on the range [0.0, $-V_C$] where $-V_C$ is the voltage that results in the cut-off of current flow.

Then, in the second part of the addendum, we will discuss the effects of overdriving the tube circuits so that the input grid voltage is positive (+) with respect to the cathode. When this happens, the nature of the amplified signal changes significantly and also varies with the amplifier design classification. This additional information is crucial if you want to implement your own plugin versions and is provided here because it was cut from the original book manuscript to save pages.

In addition, notice the output signal at the tube plate. It is inverted from the original, an electrical consequence of the fact that the tube current increases as the audio input voltage increases. More importantly, notice how the amplification is clearly non-linear. Although the figure was drawn to exaggerate this effect, it is important to understand that the amplification is asymmetrical with respect to the input signal polarity. The positive halves of the audio input signal are not amplified in the same manner as the negative portions. This asymmetrical distortion is the one type of tube distortion we will want to mimic. In the FX book, I showed you how to do that with waveshapers.

A19.4.1 Class-A Triode Preamp Design

We've established that the plan is to let the audio signal be the grid voltage control. You must now calculate the operating limits for this grid voltage and see how the audio signal will fit in. In the rest of this section, we'll be designing a Class-A Triode Preamp. A **Class-A amplifier** contains components that conduct the signal voltage and current 100% of the time. A **Class-B amplifier** contains components that share the work of conducting the signal – half the components conduct 50% of the time, the other half of the components conduct the other 50% of the time. All such designs begin with the triode plate voltage vs. plate current plot. This graph is provided by the manufacturer, and can be kind of tricky to use. Here is the graph for a 12AX7 (a.k.a. ECC83) triode:

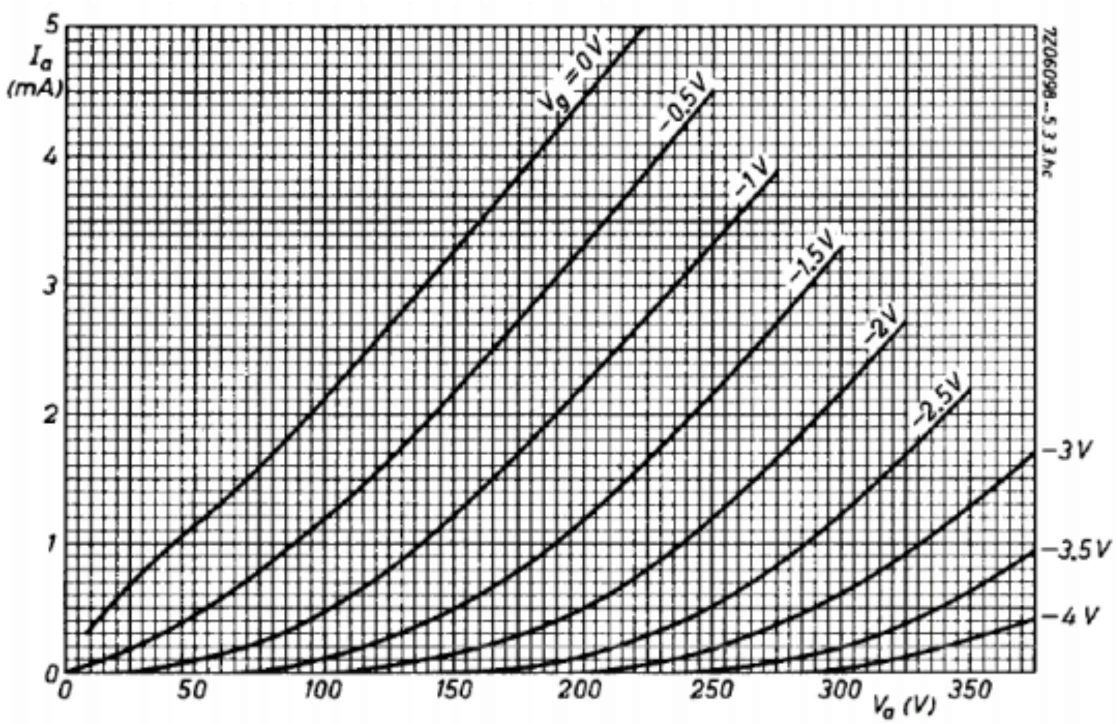


Figure A19.6: the plate voltage vs. plate current plot for several different grid voltages; from the Phillips ECC83/12AX7 datasheet, 1971

Notice the grid line curves that are labeled V_g and that start with the V_g = 0V at the upper center. Then, as you move down and to the right, you encounter the other grid voltage lines as the signal becomes more negative. The last line at the bottom right is for V_g = -4.0V. It is important to understand that our audio signal will be continuously changing, and will never lie on one grid voltage curve for very long. The useable plate voltages range from 0V to about 300V – notice how the last grid curve marked -4V just hits the x-axis around 300V. The plate current produced lies somewhere between 0 and 4mA and is plotted on the *y*-axis. It might not sound like a lot of current, but you can use it to create a large amount of voltage gain (we'll get to this later). In the next sections, we will use one of several different strategies to design the triode preamp.

A19.4.1.1 Decide on a Plate Voltage

One approach to the triode design is to start with a plate voltage and work from there. Other options include starting from a desired bias current (or standby current) or a desired plate resistance. Remember, this is just one of several approaches. Looking at the graph, you can see that the plate voltages can range from 0 to about 300V. You are going to use the grid voltage to control a current. You will apply this current to a load resistor, connected to the plate, to generate a voltage. The output voltage will be the difference between the plate and supply voltages. You'd like to be able to use a wide range of grid voltages, so you should pick a supply voltage at least halfway across the *x*-axis. Choose 200V since it is right in the middle of the plate voltage range.

- B+ = 200V

A19.4.1.2 Decide on a Load (Plate) Resistor (R_p).

You next need to choose a load resistor. We'll see how the plate resistor affects voltage gain later. For now, lets just pick a value. Choose R_p = 50K. This will affect the load line and bias calculations below. You might want to think about how you would work from the bias current backwards to finding this plate resistance as an exercise.

- R_p = 50K

A19.4.1.3 Draw the Load Line

You can make use of the fact that Ohm's law linearly relates voltage, current, and resistance by creating a load line. The load line will help us find the range of input (grid) voltages that will create a range of output voltages. You can then calculate the amplification factor and see what kind of voltage gain you get. You can also check the linearity of the amplifier to see if and when harmonic distortion will occur. To draw the load line, you need two pieces of information:

1. the maximum current that would flow through the plate resistor if the tube were replaced by a short to ground
2. the maximum voltage that would appear at the plate if the tube were replaced by an open circuit

Both of these parameters are simple to calculate. The maximum current that would flow is:

$$I_{MAX} = \frac{V}{R} = \frac{200}{50{,}000} = 0.004A = 4mA$$

The maximum voltage that would appear if the tube were an open circuit is simply the same as the supply voltage. If the tube were an open circuit, no current would flow through the resistor. This is only possible if the voltage on each side of the resistor is the same. So, V_{MAX} = 200V.

To find the load line, find the I_{MAX} point on the Plate Current (y) axis, and V_{MAX} on the Plate Voltage (x) axis. These are the two endpoints for the load line, so you can just draw a line between them. You can then find the output plate voltage for any input grid voltage by reading the plate voltage at the point where the grid voltage intersects the load line.

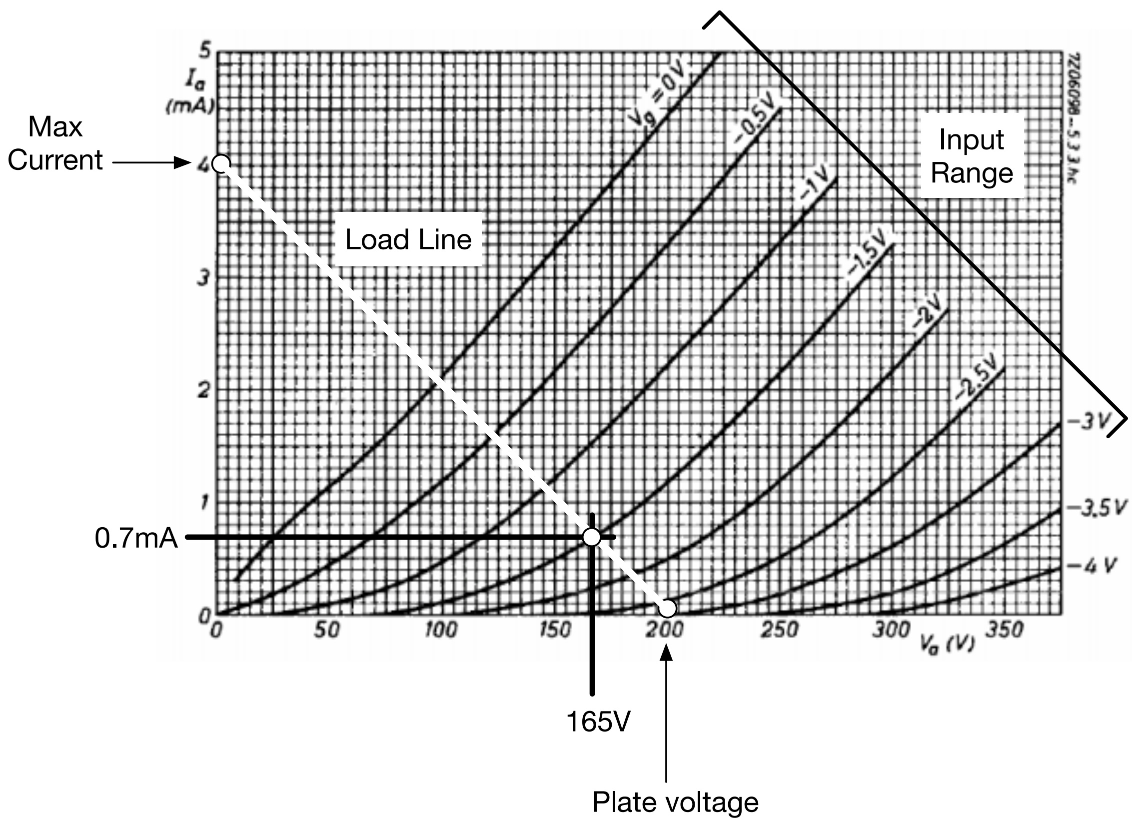


Figure A19.7: The load line ties together the relationship of grid voltage vs. plate voltage. For example, at a grid voltage of –1.5V, the plate voltage is 165V, and the plate current is 0.7mA; with a bias voltage of -1.5V, the input can swing $3V_{p-p}$

The load line is the key to testing different supply voltages and plate resistances. At this point, you need to decide on how to bias the amplifier to set the operating point (or center voltage and current).

Maximum Input Swing:
One strategy is to look at the load line and find the minimum and maximum values of V_g that intersect it. If you do that, you can see in the above load line that the range from V_g = 0.0 to V_g = -3.0 all intersect the load line. This means our largest audio input signal allowed would have a $3V_{p-p}$ amplitude and we would need to bias the amplifier using the V_g curve that is ½ of the overall input amplitude or -1.5V here.

Half Plate-Voltage Range:
Another strategy is to find the intersection points of the load line at the plate voltage (x) and the $V_g = 0.0$ curve as shown in Figure A19.9. Then you set the bias point so that it is half way between the plate voltage values. Here, that would correspond to a plate voltage of about 150V and an input bias voltage of -1V (Vg line that is intersected as a result), and a bias current of 1.1mA.

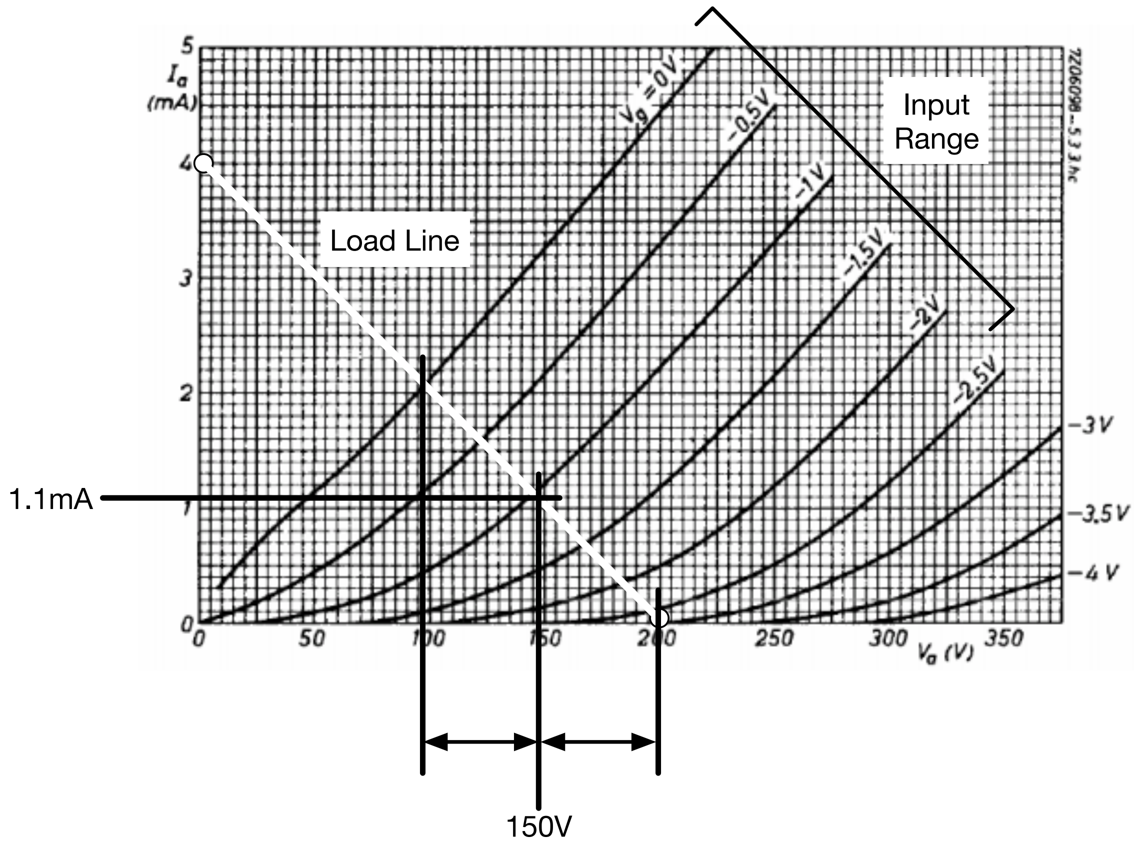


Figure A19.9: using the half-plate voltage method to set a bias point

You can also approach it from a current perspective, and set the bias current to be ½ the available range. Here, that would be about 1mA, which is very close to the 1.1 value above.

For our design we will use the maximum input swing method. You can now make the following predictions about the circuit operation:

- our input (grid) voltage can swing from 0V to –3V with respect to the cathode
- this range of grid voltages will result in a plate voltage that swings from 200V (V_{grid} = -3V) to 100V (V_{grid} = 0V)

Our voltage amplification is the ratio of input (grid) to output (plate) voltages:

$$A_V = \frac{V_{OUT}}{V_{IN}} = \frac{200V - 100V}{-3 - 0} = -33.3$$

The negative sign shows the phase inversion that has occurred. Think about it: when the grid voltage was at a maximum (0V), the output was at a minimum (100V). When the grid voltage reached its minimum (-3V) the output was at a maximum (200V). When the grid voltage is in the middle at -1.5V, the output sits at 165V – this is its DC offset and we'll remove that with a capacitor for preamps, or a transformer for power amps.

A19.4.1.4 Calculate the input bias voltage

You would like to apply our audio input signal to the grid. You've just seen that the grid voltage must swing between 0 and –3V. This means that our audio signal must also fit within these bounds. Our audio signal must not exceed 0V (upper limit) or –3V (lower limit), with respect to the cathode. The last part of that sentence is the most important – the grid must be negative with respect to the cathode, not with respect to ground, or any other absolute voltage. You know that the audio input signal needs to be biased down with respect to the cathode. What should the bias voltage be? In single supply op-amp designs that required biasing, you biased the 0V audio input to be halfway between the maximum and minimum supply voltages. In this case, you know that the audio input must stay between 0 and –3V. The audio signal's absolute amplitude value must not exceed this 3V range or distortion will occur – this is usually not a problem if you are amplifying microphones (V_{RMS} = 2 to 20mV) or musical instruments (V_{RMS} = about 0.5 to 2V). You would like to position the audio input signal so that if falls comfortable in the 0 to –3V range. Therefore, you choose a bias voltage that is halfway between these two values, or -1.5V.

V_{BIAS} = -1.5V.

Tabulate a few of the various grid voltages and corresponding plate currents for a supply voltage of 200V and R_p = 50K. You will keep the table handy and use it to continue the design process:

Supply Voltage = 200 V		
Grid Voltage (input)	Plate Current	Plate Voltage (output)
-3.00 V	0.0 mA (cutoff)	200 V
-2.50 V	0.1 mA	190 V
-2.00 V	0.4 mA	180 V
-1.50V	0.7 mA (bias)	165 V
-1.00 V	1.1 mA	145 V
-0.50 V	1.5 mA	120 V
0.00 V	2 mA (saturation)	100 V

Table A19.1 Some grid voltages and plate currents – the bias voltage and resulting output current and voltage are highlighted.

A19.5 Asymmetrical Amplification in the Triode

You should notice something a little strange about the way the plate current changes in response to the grid voltage; for the last four entries in the table, the grid voltage increases in 0.5 V steps (linearly increasing) yet the plate current does not increase linearly, especially near the saturation point. You can see this graphically by observing the distance between the points where the grid voltage lines intersect the load line - the distance between the points is non-constant even though the grid voltage lines increase at a constant interval (0.5V). This is called the *non-linear behavior* of a triode, and is important in some audio applications, where harmonic distortion is actually desirable. In these circuits, the designers plan on allowing the input signal to fluctuate over a wide range of possible grid voltages, producing a distorted output. In other audio applications, this non-linearity is not acceptable. In this case, the input signal is only allowed to fluctuate across a small range of grid voltages. Although the output is not amplified as much, the resulting waveform has low distortion.

- The bias voltage is at the center of the range of grid voltages.
- The output signal will have a DC offset of the plate voltage at $V_{BIAS.}$
- The designer may create a preamp that is very clean, operating in a small linear region of the overall load line, or a very distorted preamp that purposefully overdrives each tube stage.

Figure A19.10 shows the input/output plots for the maximum input swing and half-plate voltage range biasing strategies. Notice that the half-plate voltage range version can only accept an input of $2V_{p-p}$, it can output $80V_{p-p}$, giving it a gain of 40, which is actually higher than the maximum input swing case. Its output is also substantially less distorted than the maximum input swing version, however it is still asymmetrically distorted. It is clear from the DC offset that in both cases, the lower portion of the waveform receives more amplification than the upper portion.

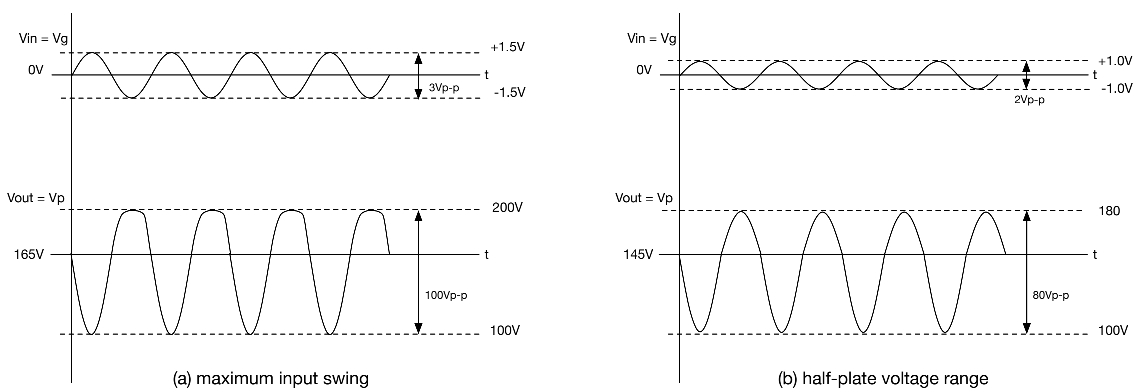


Figure A19.10: theoretical input/output relationship and plots for the (a) maximum input swing and (b) half-plate voltage bias schemes; the output waveforms are purposefully drawn in an exaggerated manner to denote the asymmetrical amplification

A19.5.1 Amplification Factor μ and Transconductance g_m

There are a couple of additional tube parameters that datasheets list and which pop up in the generalized tube equations. The first is the amplification factor (μ) and the second is the transconductance (g_m). The simplest way to understand these parameters is to examine Figure A19.11 that graphically shows their calculations. The amplification factor μ relates the output voltage at the plate labeled V_a to the input grid voltage V_g. Notice that this is done in an idealized manner with a horizontal "load line" and can be thought of as a single, ideal voltage amplification value. The transconductance relates the output current at the plate labeled I_a to the input grid voltage V_g demonstrating the change in output current to the change in input voltage.

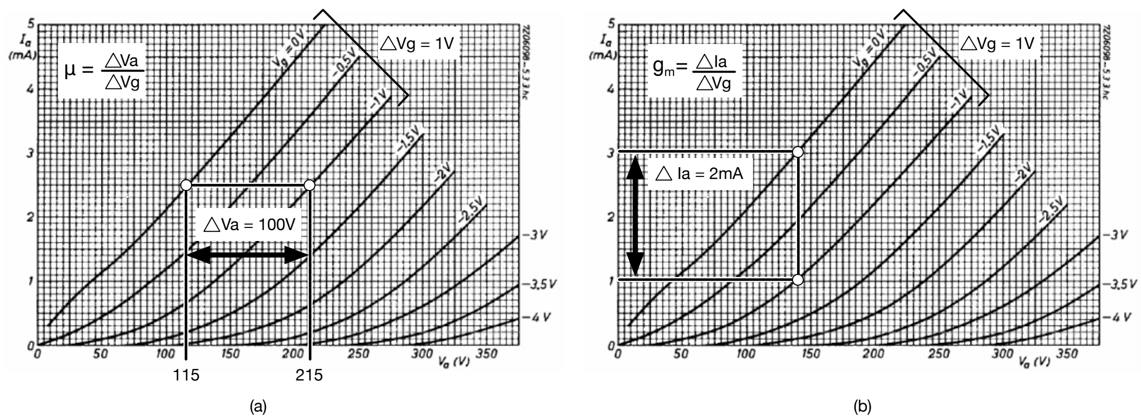


Figure A19.11: (a) the amplification factor (μ) is calculated from the output/input voltage relationship while (b) the transconductance g_m is calculated from the output-current/input-voltage relationship

A19.5.2 The Voltage Transfer Function

You can create a graph that shows the non-linear relationship of the input vs. output signals by plotting the input voltage (*x*-axis) against the output voltage (*y*-axis). This graphs the Voltage Transfer Function – a curve that represents this input/output relationship. In op amp circuits, the VTF is a straight line, until the output voltage gets to the rail voltage (rails out). Figure A19.12 shows the data from Table A19.1 plotted along with a linear connection of the upper points. The non-linearity can be seen in the curviness of the triode VTF as compared to a very linear device like an op-amp. The smoothly bent transfer function creates harmonic distortion that emphasizes the 2nd harmonic (one octave above actual input fundamental). Timbres with emphasized 2nd harmonics are often perceived as "strong" sounding.

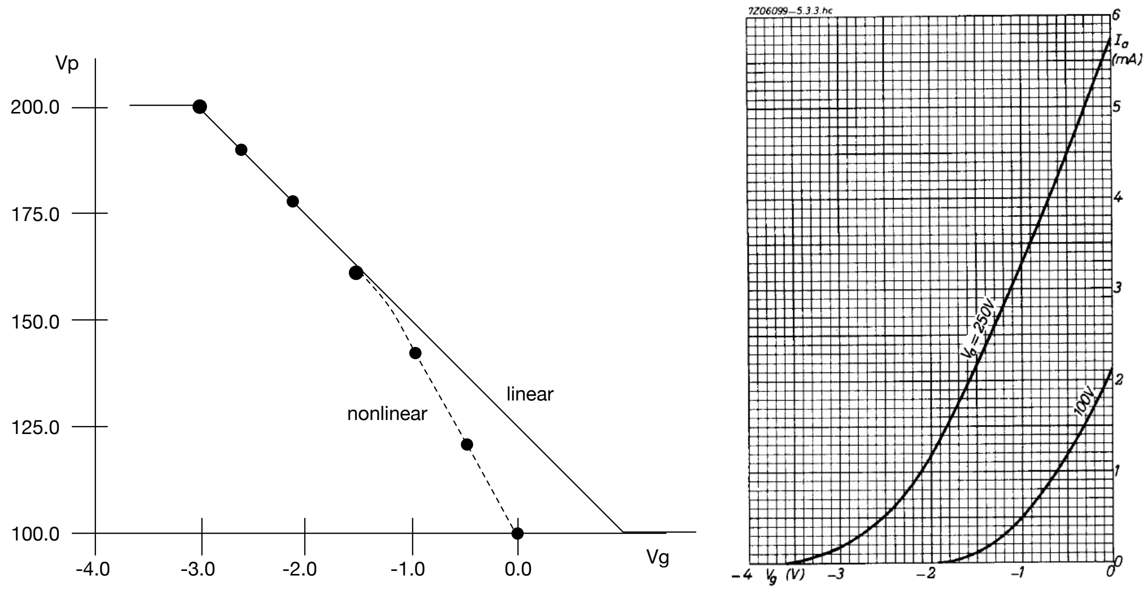


Figure A19.12: (a) the voltage transfer function for the basic triode amplifier reveals its nonlinear nature and (b) plot of V_g vs. I_p (labeled I_a for "anode" here) 12AX7/ECC83 datasheet demonstrates the 3/2 power law relationship between the grid voltage and plate current

A19.5.3 The 3/2 Power Law

It is easy to see where the asymmetrical distortion and resulting output waveform comes from when you review table A19.1 and look at the input grid voltage step sizes compared to the output voltage step sizes. For the first few rows, each change of 0.5V at the grid resulted in an output voltage change of 10.0V. However as you move down the table, the output step sizes become larger and larger for the same 0.5V input step size to the point that in the last two rows, the output has now changed by 20.0V or so. Another way to understand where the distortion and asymmetry comes from is to look at the intersection point spacing for the load line and the series of V_g lines it intersects. Since the V_g lines are drawn in -0.5V increments and the load line is straight, any deviation in spacing between adjacent V_g lines and the load line represents a nonlinear amplification step.

The data in Table A19.1 and the plot in Figure A19.11 certainly reveal the nonlinear nature of the triode amplifier. The relationship between the grid voltage V_g and the plate current I_p must also be nonlinear. This relationship is known as the 3/2 power law. It shows that the relationship between the grid voltage and plate current is actually quite complicated related by both a cube and a square-root as opposed to a simple linear relationship. It is this nonlinear relationship that gives rise to the nonlinear amplification that we observe.

$$I_p = k\left(V_p + \mu V_g\right)^{3/2} = \sqrt{k\left(V_p + \mu V_g\right)^3}$$

I_p = plate current [A19.1]

V_g = negative grid voltage

$\mu =$ tube amplification factor

$k =$ constant dependent on tube geometry

The amplification factor (μ) is a function of the tube and is 100 for the 12AX7/ECC83 however the variance in actual product is generally high. Remember that this is also an ideal voltage gain value. You can see the result of the 3/2-power law in the nonlinear plots shown in Figure A19.11 (b) measured at two values of the plate voltage taken from the 12AX7/ECC83 datasheet. If you plot some ideal curves directly from the equation, you will find the resulting curves are actually more linear in nature. It is interesting that what appears to be only a slight nonlinearity can be harnessed to produce massively harmonically distorted signals. One part of that is by cascading stages together but without purposefully overdriving them as we are doing in this first part of the document.

A19.6 Creating the Input Bias Voltage

It's time to deal with the grid voltage biasing. So far, we've just assumed that the audio input voltage has been adjusted to fit in the required grid voltage range. There are several ways to implement the biasing. One way would be to directly shift the audio signal similar to that in Figure A19.13 (a), a voltage divider is used to create the bias voltage which is applied via a pull-down resistor R_B. A coupling capacitor C_C prevents this DC offset from escaping the circuit. This is called **direct biasing**.

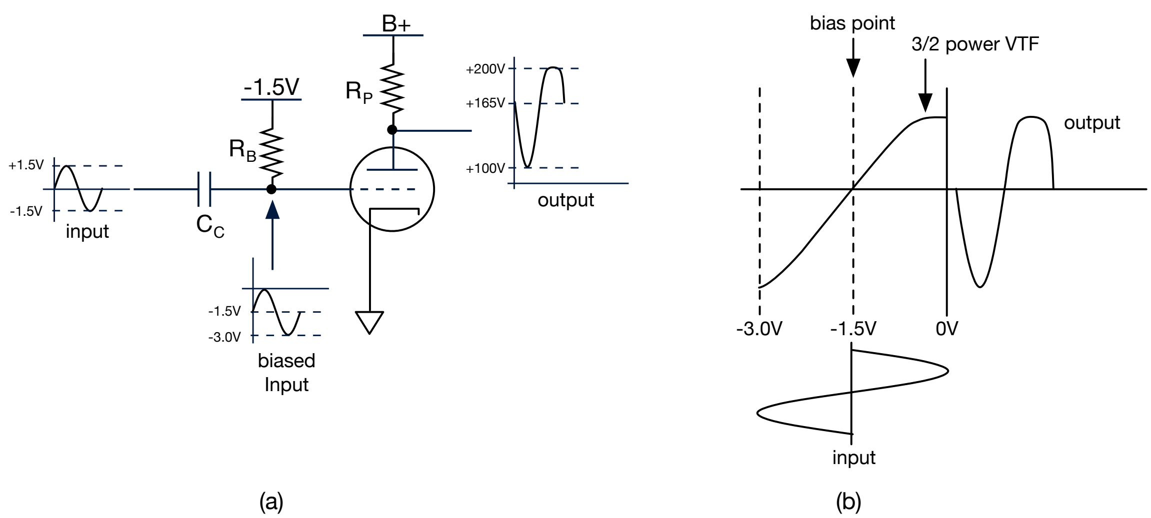


Figure A19.13: (a) direct-biased triode tube with input pulled down by -1.5V and (b) the input/output relationship showing the bias point, also called the "operating point" of the amplifier

It is important that you understand what is shown in Figure A19.13 (b) and please note that the output's inversion is built-into the transfer function plot. Here a wave shaper emulates the 3/2 power law's resulting voltage transfer function (VTF). It is important that you note and understand a couple of simple concepts in Figure A19.13 (b) that will apply to future sections of the document. In this figure:

- the input signal's amplitude never exceeds the allowed range of $3V_{p\text{-}p}$
- the bias point sets the center of operation around the VTF
- the output has a substantial DC offset voltage (the output voltages are omitted for clarity)

The notion of the "bias point" or "DC operating point" or just "operating point" of the amplifier is extremely important for future discussions. Here we have a single -1.5V DC offset that is constant and unchanging. This is because the input signal is constricted to its allowed range of $3V_{p\text{-}p}$.

A19.6.1 Cathode Self-Biasing

As it turns out, there is a much easier, more elegant way to bias a Class-A Triode amplifier that results in the same -1.5V DC offset for the grid-cathode voltage. This method is called **cathode self-biasing** and it takes advantage of the fact that the grid voltage has to be negative with respect to the cathode, not ground. It also takes advantage of the fact that there is a finite amount of current flowing at the bias point: 0.7 mA in this example (see Table A19.1). The cathode self-bias is achieved by the addition of a resistor R_k from the cathode to ground. Assume the bias current is flowing and calculate the cathode voltage in Figure A19.14: (a). A resistor R_K has been added to the cathode. If you assume that all the current flows from the plate to the cathode ($I_k = I_P$ is a valid assumption) then you can easily calculate the voltage at node V_k as:

$$V_k = I_k R_k = I_P R_k$$

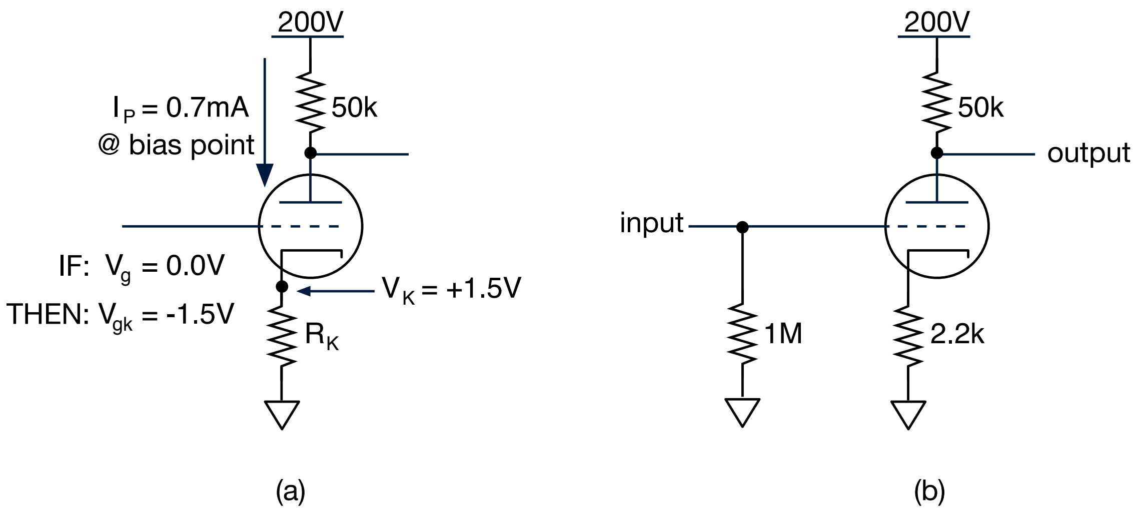


Figure A19.14: (a) addition of a cathode resistor; if we calculate the proper value, we can force a +1.5V offset at node V_k and (b) the resulting triode preamp with 1M input pull-down resistor added

This means that the grid voltage applied as an input will be measured with respect to V_K. Suppose an audio signal of 0V were applied to the grid. If the cathode voltage were +1.5V, then the grid would be negative with respect to the cathode – in this case, the grid would be –1.5V with respect to the cathode. In other words, you can make the grid voltage negative with respect to the cathode by placing a positive voltage on the cathode! This is the essence of the self-bias method. All you need to do now is calculate the value of R_K that causes V_K to be +1.5V at the bias point. You can just rearrange Ohm's Law as:

$$R_k = \frac{V_k}{I_C} = \frac{1.5V}{0.7mA} = 2141\Omega \approx 2.2K$$

You now need to actually bias the grid. You've actually selected to bias the grid at 0V. Think about it: with the grid at 0V, half the available current flows (0.7mA) and the output is sitting at its quiescent level of 165V. So, you need to ensure that the grid is biased to 0V. You can add a resistor to ground to bias the grid at 0V called a pull-down resistor; another way of thinking about this is what happens if the user disconnects the input to the amp? This resistor keeps the grid biased at 0.0V when there is no input applied at the amp – exactly what we want.

A19.6.2 Degenerative Feedback

There is a potential caveat in the cathode self-biasing scheme. The grid voltage is compared against the cathode voltage, and current flows accordingly. Suppose the cathode is sitting at 1.5V, and the input is at 0V. The grid appears to be –1.5V with respect to the cathode, 0.7mA of current is flowing, and the plate is sitting at 165V. Now, suppose the grid voltage increases to +1.0V. The grid is now less negative with respect to the cathode, so more current flows. The plate voltage will decrease (you expect this – the input is rising, and output is falling). The problem is that the cathode voltage will increase (V=IR) due to the increase in plate current. This throws off our whole biasing plan. The increase in the cathode voltage makes our grid voltage (still at +1.0V) appear **more** negative with respect to it, and the plate current then drops. The cathode self-biasing resistor is considered a kind of negative feedback, because *increases* in voltage across it create net *decrease* in voltage gain. In a sense, the cathode resistor is working against us. The grid voltage never makes it through its full range of values because the cathode voltage keeps changing it in the opposite direction, partially canceling the grid voltage.

On the other hand, the feedback is not enough to totally shut down the amplification factor, just give it a major decrease. The good news is that triode amps, like all audio amps, have the same Gain vs. Bandwidth relationship. As the gain drops, the bandwidth becomes wider, and the amp becomes more and more stable. So, if you can live with decreased gain, then the cathode self-biasing scheme works fine. You will find this kind of amplifier circuit as the preamp for many older guitar amps, which produced less distorted sounds. You might also find it on a high-fidelity stereo power amp, in the preamp section. Remember, lower gain equals less distortion and higher bandwidth and early amplifiers were designed for linear operation.

A19.6.3 Cathode Bypass Increases Gain

You may want to really take advantage in the high-gain non-linearity and use a larger extent of the grid voltage range. The solution is to add another component to the circuit; a parallel capacitor connected to the cathode. If you select the proper value, the cathode capacitor can be made to have a low impedance to audio frequency signals. This means that the AC component of the signal will travel from the plate to the cathode through the capacitor. For DC signals, the capacitor has an infinite impedance so it appears as an open-circuit to the DC component. By separating the AC and DC components this way, you can maintain a constant bias voltage and current, and produce an AC signal at the plate that doesn't interfere with the biasing mechanism.

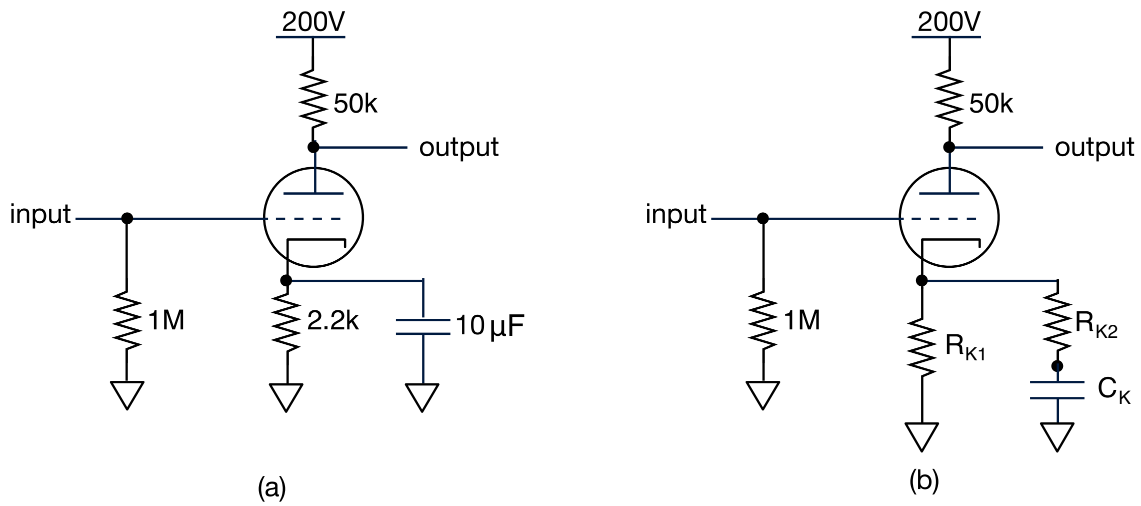


Figure A19.14: (a) addition of a cathode bypass cap for added gain and (b) common example of a partially bypassed cathode resistor

This circuit is starting to take shape as you add the cathode capacitor C_k that separates the combined AC/DC signal into a separate AC and DC path. The DC bias voltage at the cathode is preserved because the cathode capacitor effectively removes the AC component. The rule of thumb for calculating the value of C_k is to choose a value that results in a capacitive reactance that is 10% of the value of the cathode self-biasing resistor, at the lowest frequency of operation. For mic preamps, you might choose the lowest frequency to be 20 Hz, while bass amps would require 41 Hz (low E string on bass) and guitar amps would only need to go down to 82 Hz (low E string on guitar). Suppose we're designing a guitar amp. Then the value of C_k is found as:

$$X_k = 10\%(R_k) = 220\Omega$$
$$C_k = \frac{1}{(\omega)(X_k)} = \frac{1}{(2\pi f)(X_k)} = \frac{1}{2\pi(81)(220)} = 8.9\mu F \approx 10\mu F$$

Since 8.9µF is not a standard value, you would choose the next larger standard value of 10µF. You should note that in some cases, especially high-distortion guitar amps, you can wind up with too much bass response in the preamp stage, and often the value of the capacitor is decreased to remove more bass frequencies. Figure A19.14 (a) shows the addition of this capacitor.

> The cathode capacitor sets the low frequency response of the amp circuit. At very low frequencies, the capacitor impedance is high, so most AC current flows through the cathode resistor. This produces negative feedback, and brings the gain down. As the input frequency increases, the capacitor becomes more of a short circuit, increasing the gain.

In modern tube preamps, there is usually a lot of massaging of the individual bass response of each triode stage. In many cases, every preamp triode module is voiced in a slightly different manner. Figure A19.14 (b) shows a partially bypassed cathode capacitor. Here, there are two AC paths, one through each branch from the cathode. This will also modify the filtering aspects as well.

You should also note that overdriving any signal will add upper harmonics to it, and our ears will synthesize a bass response to those harmonics (this is exploited in the Virtual Bass algorithms from the same FX book chapter). Therefore, the added distortion can make the signal sound overly bass heavy and for guitar preamps, this generally results in a "muddy" sound. Removing the bass, or otherwise band-pass filtering the signal between stages helps improve the quality of the distortion. And, you can always add the bass back later via filtering, and after all distortion has been applied.

A19.6.4 Filtering Effect of the Bypass Capacitor

The FX book does go into some detail about the filtering effect of the bypass capacitor and this is something you should refer to as needed. If you are good with circuit simulation, this is a great way to experiment with different values of the bypass capacitor or multiple resistor/capacitor combinations for partially bypassed amplifiers. Figure A19.15 shows both the time and frequency domain plots for our amplifier with no bypass, and full bypass with the 10µF bypass cap we designed. A partial bypass version is also plotted with R_{K1} = 1.8k and R_{K2} = 1k and C_K = 10µF for comparison. Notice how the fully bypassed version exhibits clipping on the upper waveform. This represents the tube in a fully cutoff state with no plate current flowing – the output voltage at the plate will equal the B+ voltage of 200V. Since the output is inverted, this clipped waveform portion represents what was originally the *bottom* portion of the input signal. In this case, the input signal's bottom portion reached the cut-off point before V_{GK} = -3.0V

The fully bypassed cathode resistor output shows the first type of waveform clipping we have seen so far. In this case, the tube has stopped amplifying in this polarity (direction) so the output simply stops increasing. This is called *cut-off clipping*. Increases in the input signal's negative amplitude do not get amplified, but they also do not alter the input signal in any way. In other words, the clipping that is occurring is happening inside the tube itself – this will be important for the second part of the document.

This kind of clipping may be accomplished with relatively simple wave shaping and is the part that is covered in the FX book where you have numerous asymmetrical wave shapers to choose from. The way the signal clips is considered to be "soft" because it leaves quite rounded corners as compared to a solid-state hard-clipped waveform with very sharp edges.

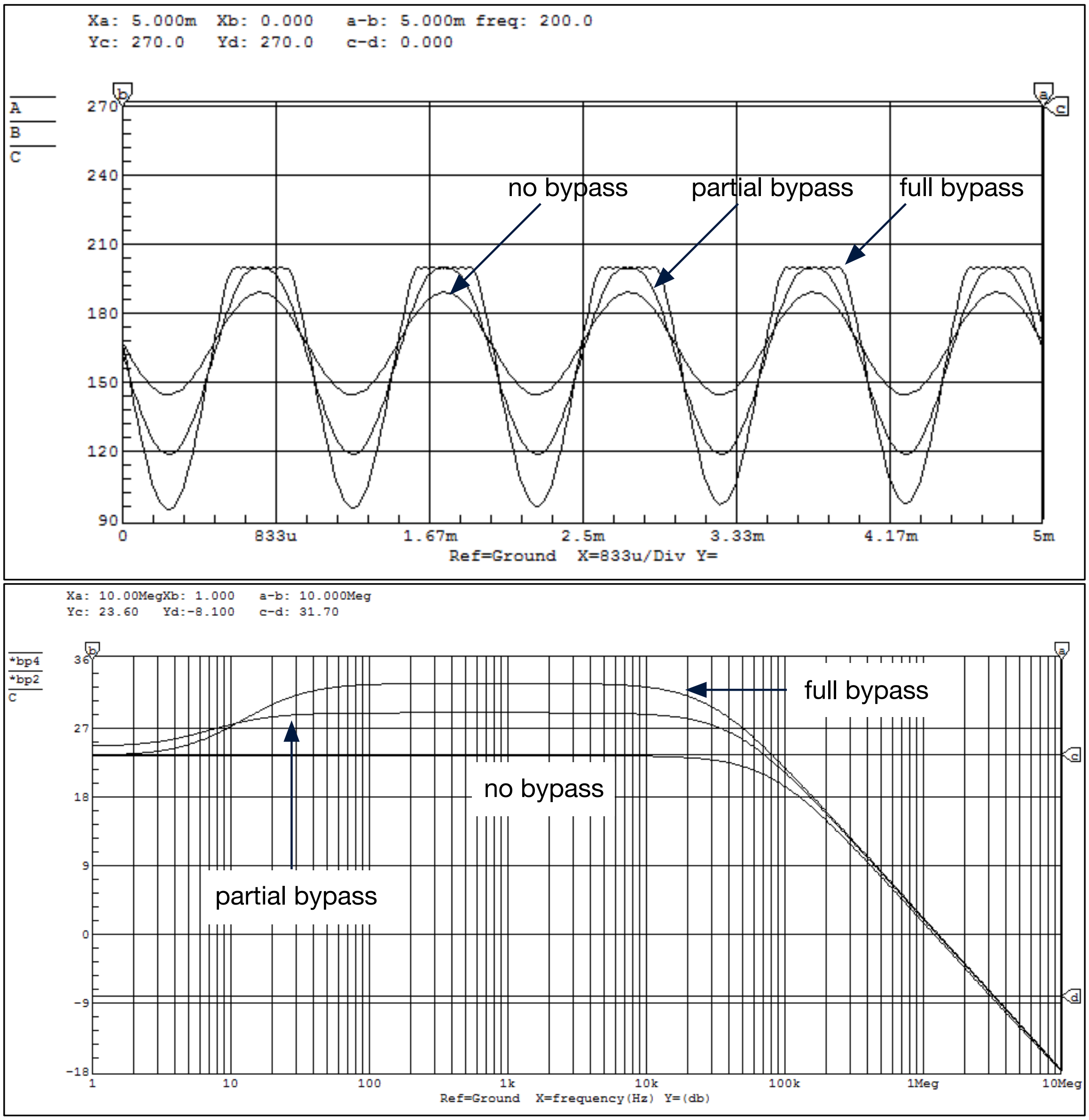


Figure A19.15: (top) the time domain output of a single triode class A amplifier from this chapter with various types of cathode bypassing and (b) the frequency response that results

In Figure A19.15 (b) you can see the low frequency shelving effect that results from the cathode bypassing operation. This shelving filter is part of the *TriodeClassA* object that the book discusses. The addition of the filter allows you to custom voice each of the triode modules you emulate, allowing you to mimic the way multi-stage preamps are voiced. Filtering in combination with wave shaping was the essence of the book project's design approach and is a useful, if not overly simplified way of generating the distorted output we desire, without having to wade too deeply in triode circuit design. This document will serve to augment the book projects and code as we have ample room to wade around in the circuit design all we like!

A19.6.5 Preventing "Unwanted" Distortion: the Grid Stopper

This is the beginning of a new discussion that is not included in the FX book. What happens if/when the grid-cathode voltage is higher than 0V? Once the grid takes on a value above the cutoff point, it begins to appear as another plate (anode) for the cathode, and electrons will begin flowing from the cathode and out of the grid. This causes what is known as *blocking distortion* in the output. Blocking distortion occurs when electrons flow out of the grid, so if you could limit the amount of grid current, you can limit the blocking distortion that occurs. This is pretty simple to accomplish – resistors are the component of choice when it comes to limiting current – that's what they do. You will add a resistor R_G in series with the grid to limit the current flowing through it. This resistor is known as a *grid-stopper* resistor. For the 12AX7, 68k is the usual value. With this last resistor in place, the Class-A triode preamp module is nearly completed and shown in Figure A19.16 (a). Note that we are going to come back to this issue of overdriving the grid above 0.0V – in fact, we will assume that the input is massively overdriven above 0.0V to see what happens with this blocking distortion.

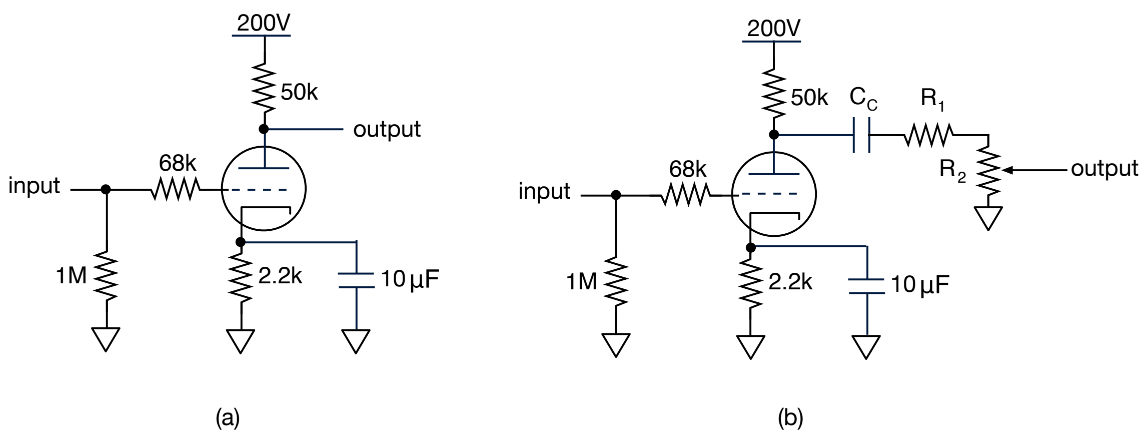


Figure A19.16: (a) our class-A preamp is nearly complete with the addition of the 68K grid-stopper resistor and (b) completed with the addition of an output coupling capacitor and voltage attenuation/divider circuit

A19.6.6 Output Coupling and Amplitude Control

The last thing we need to do is deal with the 165V DC offset voltage that is riding on the audio output at the plate. We can add a DC blocking capacitor to remove the DC component. In multi-tube preamps, the amplitude adjustment between each tube stage may be accomplished with a simple resistor divider. In some cases, both a potentiometer and a fixed resistor are incorporated to accomplish a set amount of voltage division with the ability to adjust the attenuation as needed. A generalized version is shown in Figure A19.16 (b) that includes a coupling cap C_C along with a variable attenuator formed with fixed resistor R_1 and potentiometer R_2. This creates the second filtering operation in the triode module. A high pass filter is formed between the output coupling capacitor and whatever resistance (or combination of resistances) follows it downstream. This high pass filter is also included in the FX book's *TriodeClassA* object. The effect of this high pass filter may augment the shelf that the cathode bypass cap causes, or it may obliterate it altogether if its corner frequency is much higher than the shelving frequency.

A19.6.7 Cascading Triode Stages: No Overloaded Grids

As pointed out in the FX book, tube preamps usually consist of series cascading gain stages, with the output of each stage being inverted and asymmetrically amplified, then attenuated (a little or a lot, depending on the design), and re-amplified again through the next stage. Each time, the signal is inverted and amplified asymmetrically.

If you create a series of triode tube stage models, and then attenuate the signal between stages so that the next stage's input range does **not** fall outside the allowed V_g range, then you can emulate this subtle soft clipped processing via simple wave shaping and filtering.

> The FX book's limited space only allowed me to discuss and show a tube simulation in which the input signal size was never allowed to go beyond the maximum input value that the triode could accept. For something like a microphone tube-preamp, which may only include one or two triode stages, and which is clearly not designed for maximum distortion, this is fine. In addition, some classic guitar and bass amps were never really designed for high gain overloaded conditions in the preamp because the music of the time did not require (or want) it. These are also good candidates for soft-clipping wave shaping and other less severe manipulation. In the 2nd part of this document, we will discuss overdriving tube stages with V_g values that are positive with respect to the cathode, creating a completely different distortion sound and waveform. This is the primary reason for the addendum.

Let's take a simple example of a two-triode preamp with cascading gain stages.

A19.6.7.1 Gibson GA-5 Preamp

The Gibson GA-5 guitar amplifier's preamp consists of two triode preamp circuits using a dual triode 12AX7/ECC83 tube. The input Figure A19.17 shows this simple preamp circuit. As an exercise, you might want to try to plot the load line and reverse engineer its design. The input preamp uses a fully bypassed cathode self-biasing resistor. Its output is DC coupled through the 0.02uF cap and then attenuated through the 1M potentiometer that acts as a variable voltage divider and labeled "Volume" on the front panel. This feeds a nearly identical stage, except that it includes a negative feedback input from the output stage. The negative feedback consists of a signal that is in phase with the input signal to the tube. This acts as forced degenerative feedback by following (in phase) and *increasing* the cathode voltage, bringing down the overall gain. This increases stability and bandwidth, and allows for some filtered feedback in the form of the "Presence" control in some amplifiers.

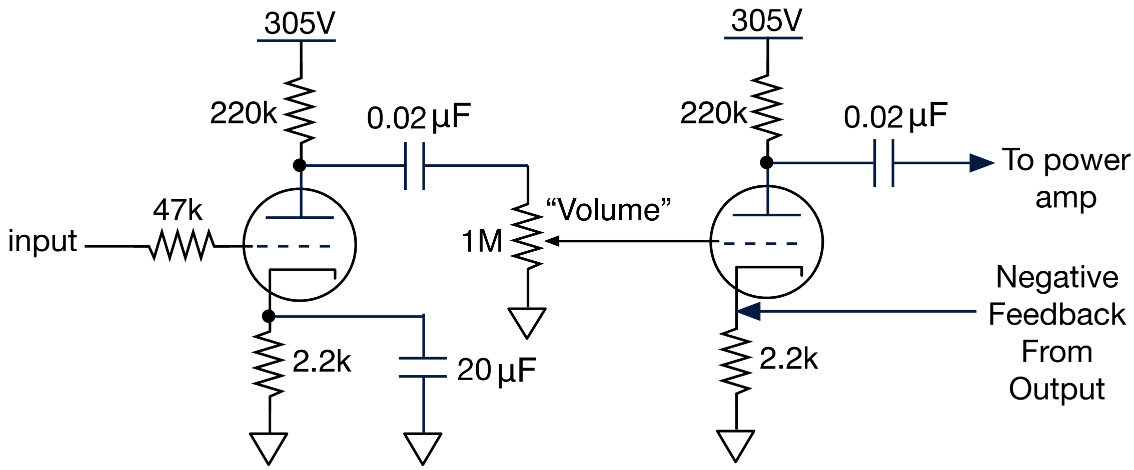


Figure A19.17: the Gibson GA-5 preamp

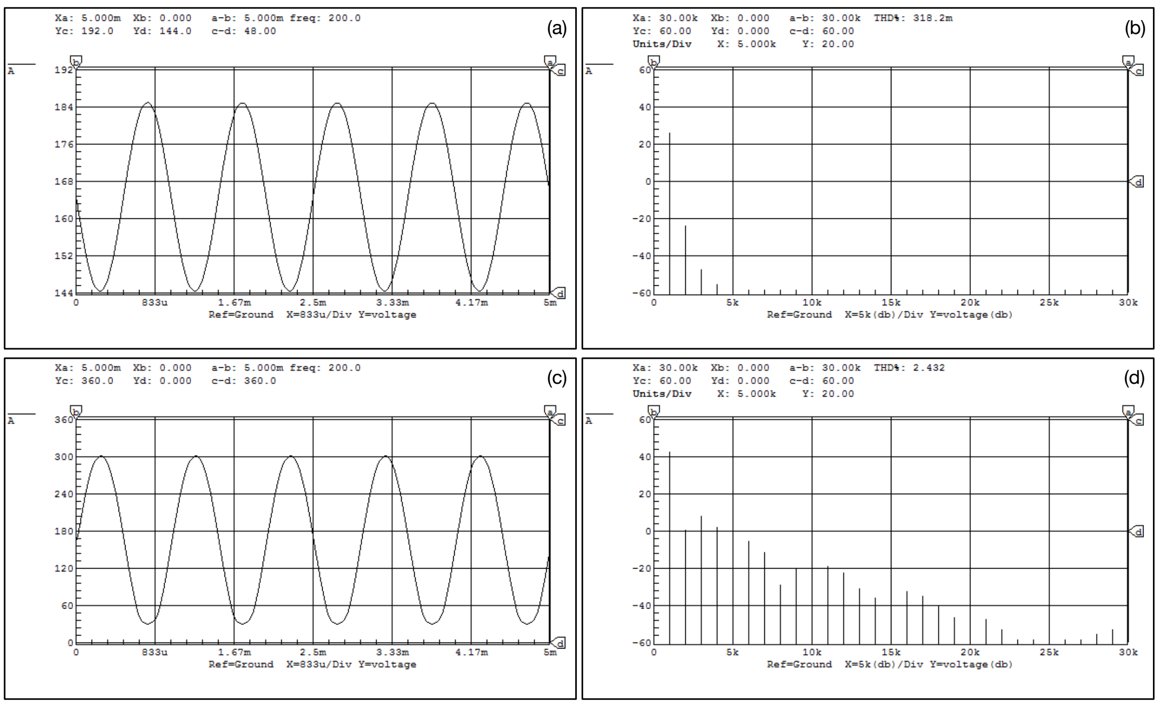


Figure A19.18: the time domain and frequency spectra (FFTs) for SPICE simulations of the Gibson GA-5 preamp; the input is $1V_{p\text{-}p}$ and the volume control is at 20%, and we assume no negative feedback (worst case); the plots show (a) the time and (b) spectral plots of the output of the first tube while (c) and (d) show time and spectral plots of the output of the second tube

Figure A19.18 is interesting as it shows the build up of harmonics through the simple preamp with a medium sized input of $1V_{p\text{-}p}$ and the volume control just turned up to 20%. Notice that the output of the second tube reveals asymmetric amplification (the lower halves are a little more tubby) <u>but there is no hard clipping occurring that is visually evident</u>. And, we see quite a build up of the harmonic content of the signal. Note that there is no negative feedback into the second tube in the simulation, so in the actual amplifier we would expect to see a cleaner-still signal in both time and spectral content.

Figure A19.19 shows the frequency response of the preamp that includes indications of the gain increase through the pair of tubes, but also in the frequency shaping that occurs. You can see that the output of the 2nd tube has more gain, but also has a narrower bandwidth. This is a result of multiple filtering operations through each stage. On the right side is the output of a simulation with four identical triode stages and three potentiometers, one between each. You can see that as the gain increases, the bandwidth of the cascade of them shrinks. We can make an improvement to the *TriodeClassA* tube by including an output low pass filter that simulates the inherent limited bandwidth of the amplifier. If we cascade the stages together, the bandwidth should shrink accordingly as the gain is increased producing the narrowing band-pass response that we observe here.

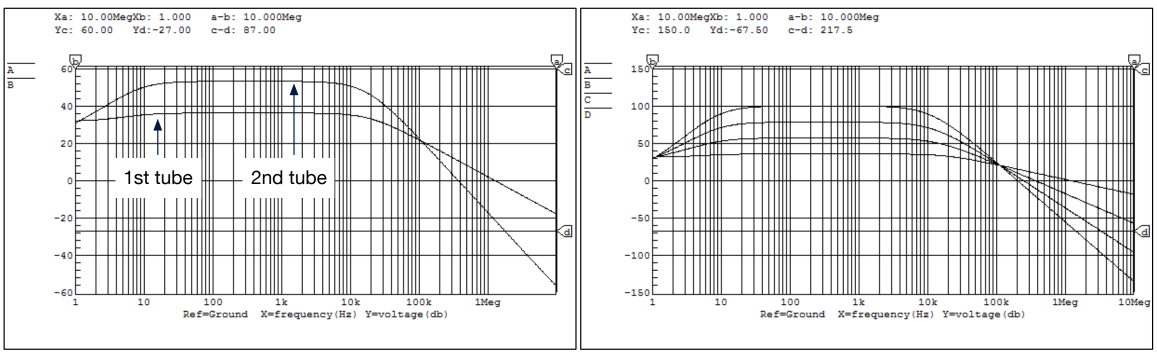


Figure A19.19: (left) the frequency response of the outputs of the two tubes; the first tube gives about 38dB of mid-band gain, with a small shelf from the bypassed cathode resistor while the second tube output adds more reduction in bass (via the coupling capacitor and the already reduced component from the first tube), and a faster roll-off of high frequencies as well (right) output of an experimental version with four identical preamp tubes

A19.7 Tetrodes & Pentodes

The tetrode and pentode, containing four and five electrodes respectively, were advancements in the triode design. First, they allowed the tube amp to be designed in a more linear manner because their plate voltage and current vs. grid voltage curves more closely approximate an ideal case. Secondly, they allowed much higher voltages and currents to provide ample power amplification to drive a loudspeaker. The tetrode helped to reduce a parasitic capacitance inherent between the grid and plate by inserting another grid between the pair. This "screen grid" is tied to a voltage that is close to, but slightly less than the main B+ voltage. The insertion of the grid prevents a capacitive potential from forming across the (normal) grid and plate. The pentode solved a problem called *secondary emission* in which electrons were knocked form their orbits in the plate material as the beam of electrons smashed into them at high velocity (the electrons coming from the cathode are moving at thousands of kilometers per second). The pentode added another grid called the "suppressor grid" between the screen grid and the plate. The new grid is tied to the cathode so it represents a negative charge that repels the secondary emission electrons back to their homes. The pentode and tetrode may both be run in “triode” mode by simply tying the screen grid to the normal B+ voltage. This effectively removes it from the circuit.

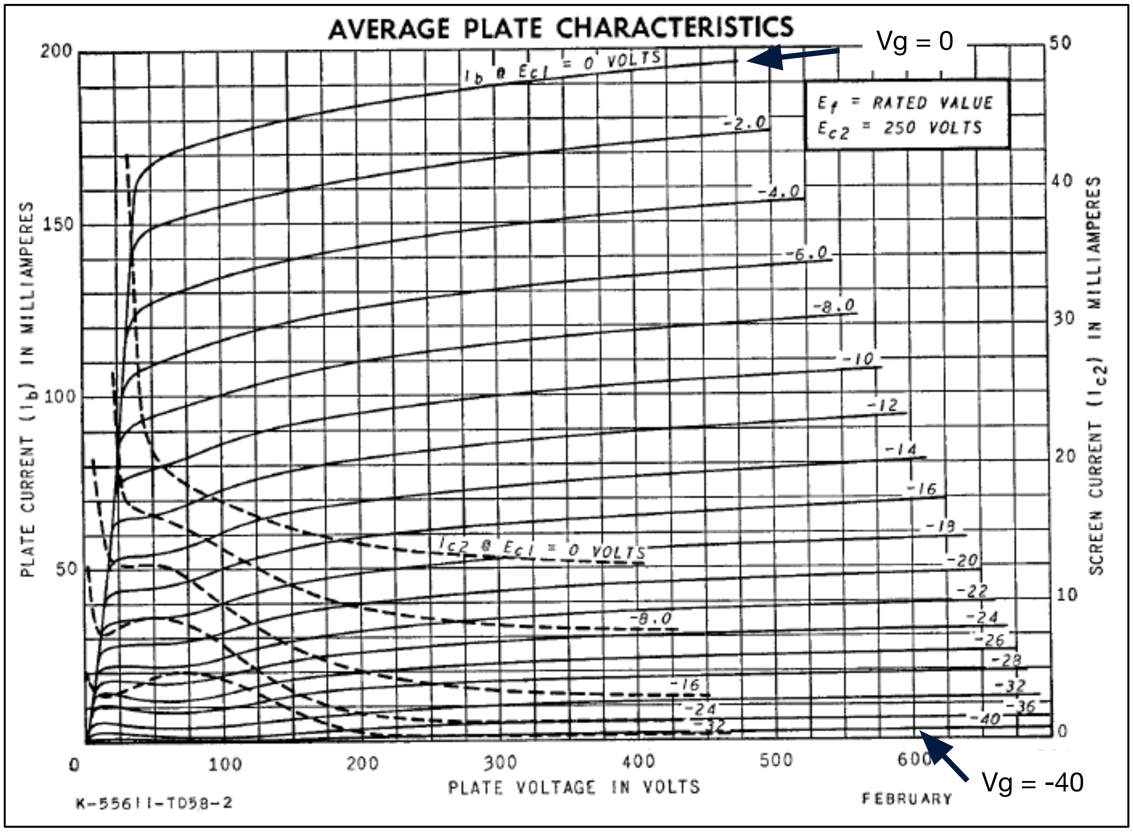


Figure A19.20: the plate current and plate voltage vs. grid voltage for the 6L6-GC pentode tube

Figure A19.20 shows the plate current and plate voltage vs. grid voltage for the 6L6-GC pentode tube. The dotted lines represent a third plot showing the screen grid current and you can ignore these for our discussion here. We can make some observations about Figure A19.20:

- the permissible grid voltage range is drastically increased over the 12AX7/ECC83 triode and we can see the last curve at V_g = -40V
- the grid curves are now laid over on their sides and do not look like the triode curves (at all)
- scanning down the “rows” formed by the grid voltage curves, we can see that the spacing between the curves begins to shrink, and at the bottom end, the curves are 4V apart rather than 2V apart, and continue to shrink closer together

The first bullet shows that this tube has an input range that could potentially span $40V_{p\text{-}p}$. This means that we could make a simple tube amp whose preamp tubes could directly drive the power amp tube with enough voltage to cover its input range (and more). We’ll show an example of that shortly.

To understand the ramifications of the second bullet point, think about what we would ideally like to see in the tube's plate voltage and current vs. grid voltage plots. Figure A19.21 shows the same pentode plots with an ideal version to the right. Imagine a load line plotted across the ideal version as shown in the Figure. Notice that all of the grid lines are perfectly evenly spaced and perfectly horizontal. Then, the intersection points with the load line would always be evenly spaced in both directions so that we would achieve perfectly linear amplification at any grid voltage, and any plate voltage/plate resistor combination. So the pentode gets us closer to our ideal case, at least in part.

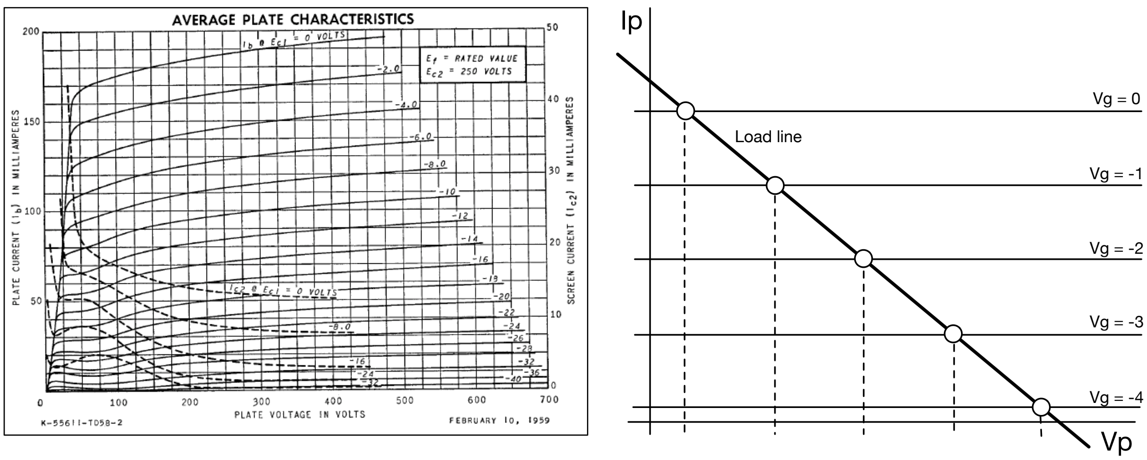


Figure A19.21: the "laid over" grid voltage curves of the pentode along with an idealized version

Finally, the last bullet point is very important. It means that even though the "laid over" grid voltage curves are looking more like the ideal response, their uneven spacing is still going to introduce harmonic distortion because the load line intersection points will not be evenly spaced apart. Because these curves are sort of lying sideways, we can get a lot of visual information from observing the way that the spacing in between the lines contracts as we move down them. For signals that are close to Vg = 0, the amplification will be linear. As the input signal at the grid increases in the negative direction, the output amplification step size will decrease, squashing the waveform a bit and introducing harmonic distortion.

Once again, we can model this nonlinear amplification easily and cheaply with simple wave shaping. We would need to change the wave shaping equation so that it generated a harmonic signature that was closer to the pentode under examination rather than the 12AX7 we've been working with so far.

A19.8 Class-A Pentode

The Class-A pentode circuit is shown in Figure A19.22 (a). You can see the addition of the screen grid and its associated voltage, as well as the suppressor grid. Note that the suppressor grid electrode is internal and we do not connect anything directly to it. The design follows exactly as with the triode including the load line, cathode self-biasing, and cathode bypass capacitor. The type of nonlinear amplification is different from the triode. By connecting the screen-grid to the plate, the pentode may be run in triode operation – this is a simple feature to add to a Class-A pentode amplifier with a switch and a few resistors.

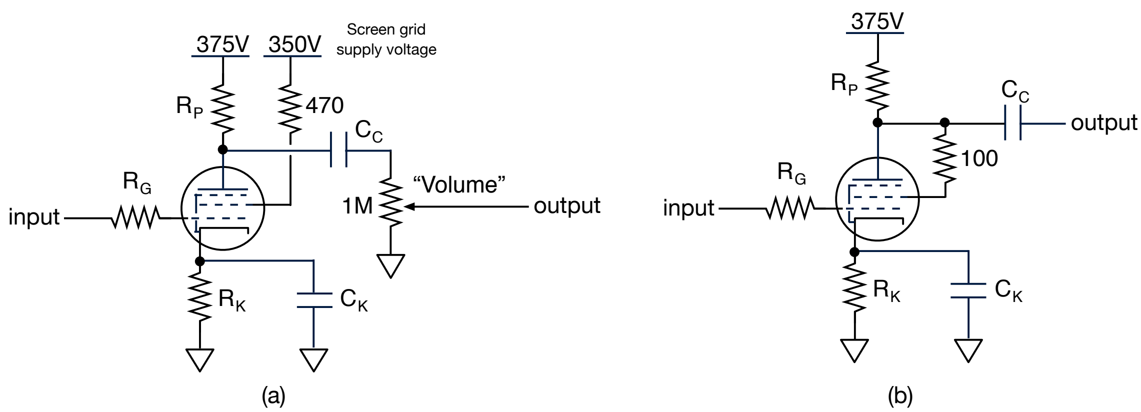


Figure A19.22: (a) a pentode tube connected as a Class-A amplifier and (b) running the pentode in triode mode by connecting the screen grid to the plate via a low-valued grid-stopper resistor that prevents parasitic oscillations

A19.8.1 Output Transformer Coupled Loudspeaker

The 12AX7/ECC83 triode preamplifier we designed can output a $100V_{p\text{-}p}$ waveform ($70.7V_{RMS}$), but only with a maximum current of 2mA. The RMS power it delivers into the plate resistor R_p would only be 0.14W. However, the pentode is a different story; from the plot in Figure A19.20, you can see that it outputs *hundreds* of milliamps of current and is able to deliver a substantial amount of power into the plate resistor, which acts as the load for the amplifier. A simple pentode power amp can deliver around 5W of power – much louder than you might think. The problem is that the plate resistor is what receives the output power. We would like to couple the amp to a loudspeaker of around 4Ω - 32Ω but the circuit design will require a plate resistor in the tens or hundreds of kilo-ohms, ranging from about 5k to 500k.

In order to get the power delivered into a loudspeaker, we need to harness the power dropped across the plate resistor and transfer it to the loudspeaker. This is accomplished with a transformer known as the *output transformer*. The transformer accomplishes three feats at once:

- It converts the loudspeaker's low impedance into the value the amp requires on the order of hundreds of kilo-ohms via load-reflection, a consequence of the transformer itself

- It theoretically transfers 100% of the power from the primary coil, which acts as the plate load, into the loudspeaker that is attached to the secondary coil
- It prevents the DC offset present at the plate from being transferred into the loudspeaker because transformers can not transmit DC signals

If you already know how transformers work, you can skip the rest of this section. It is included here for completeness and those who do not have any transformer experience.

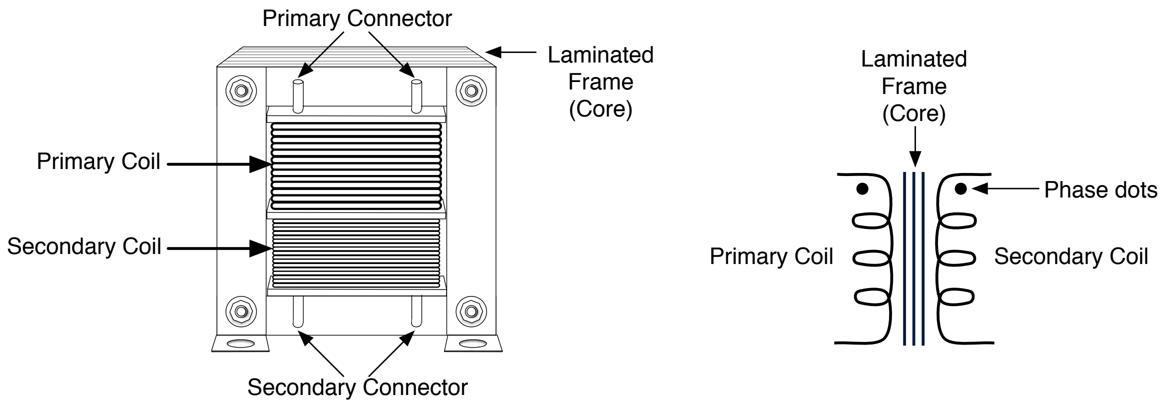


Figure A19.23: (left) a typical transformer with stacked coils separated with insulators and (right) the schematic symbol

Transformers are components consisting of two or more inductors placed physically close to one another, usually in a back-to-back or side-by-side arrangement. Most often, the inductor coils are wrapped around an iron core in the shape of a ring or 'H' however this is not a steadfast requirement. The simplest type of transformer consists of two coils called the *primary* and *secondary* coils. More complex transformers may have multiple, independent coils in the primary and/or secondary.

Transformers take advantage of Faraday's Law in two directions. Recall from your physics class that when an AC current is applied to a coil, it creates a varying (expanding and collapsing) magnetic field around the coil. Likewise, when a coil of wire is immersed in a varying magnetic field, it induces a current in the coil. Because the transformer's coils are placed physically close together, a magnetic field generated in one coil in response to a current will envelop the other coil, inducing a current in it. You can transfer an AC signal between the two coils without the coils actually touching. The transfer mechanism is an expanding and contracting magnetic field. The addition of a core helps make this transfer more efficient. The coils themselves use wire that has a special insulating coating painted on it so that even though the wires are wrapped tightly around each other, they do not make electrical contact or conduction because of the coating.

Transformers have numerous uses in audio that we don't have time to completely discuss. However, there are two main functions we use for the tube output transformer: load reflection and maximum power transfer. There is one set of basic transformer equations that relates the voltage, current, power and load reflection all at once. The fundamental basis of operation is in the number of turns of wire in each coil. The coils in Figure

A19.23 were purposefully drawn in a manner that shows a different number of turns of wire in each coil. The turns ratio ***a*** (also called the transformation ratio) is defined as:

$$a = turns\ ratio = \frac{\text{number of turns in primary}}{\text{number of turns in secondary}} = \frac{n_p}{n_s} \qquad [A19.2]$$

Please note that in some books and papers, the turns ratio is defined as the number of turns in the secondary divided by the number of turns in the primary, or a flipped version of the equation. Ultimately, it does not matter because of the way the ratio is used.

The transfer of voltage and current across the two coils is related to the turns ratio as:

$$a = \frac{n_p}{n_s} = \frac{V_p}{V_s} = \frac{I_s}{I_p} \qquad [A19.3]$$

Note that the ratio of voltages across the primary and secondary is identical to the turns ratio, while the ratio of currents through each coil is inversely related. In other words, as the voltage is increased, the current is decreased, and vice versa.

Example: A transformer has a turns ratio of 1:10. An AC generator is attached to the primary coil. A very high load resistance is connected to the secondary (to force current to flow). The generator is capable of producing 5V at 1A. What is are the resulting voltage across and current through the secondary? Is power increased or decreased as a result?

Answer: Using equation [A19.3], you can easily find:

$$V_s = \frac{V_p n_s}{n_p} = \frac{(5)(10)}{1} = 50V$$

$$I_s = \frac{I_p n_p}{n_s} = \frac{(1)(1)}{10} = 0.1A$$

Therefore, you increased the voltage (amplification) and decreased the current (attenuation). The power transferred from input to output is:

$$P_{IN} = I_{IN} V_{IN} = (1)(5) = 5W$$

$$P_{OUT} = I_{OUT} V_{OUT} = (0.1)(50) = 5W$$

You see that the power is neither increased nor decreased and this makes sense for the conservation of energy. But, it also shows that at least theoretically, all of the power delivered into the primary coil will be transferred into the secondary coil. In reality this does not happen due to physical losses, heat, etc…, but the transformer does transfer most of the power across the coil.

If you work through the Ohm's law relationships between the coils and their associated impedances, you can flush out the last part of the transformer equation that relates these impedances to the turns ratio.

$$a = \frac{n_p}{n_s} = \frac{V_p}{V_s} = \frac{I_s}{I_p} = \sqrt{\frac{Z_p}{Z_L}}$$

Z_p is the impedance of the primary coil [A19.4]

Z_L is the impedance of the load connected to the secondary coil

This last part of the equation is called *load-reflection*. Using this property, we can make the loudspeaker's low impedance to appear as a much higher impedance to the tube – we can try to make it match the desired plate resistance our design calls for.

Example: Calculate the turns ratio of the required transformer that reflects an 8Ω load as a 2000Ω load for the tube output.

Answer: Rearrange [A19.4] for the turns ratio as:

$$a = \sqrt{\frac{Z_P}{Z_L}} = \sqrt{\frac{2000}{8}} = 15.81$$

In order to couple the output of a Class-A tube amp into a loudspeaker, you replace the plate resistor with the primary coil of the output transformer. The secondary coil is attached to the loudspeaker, whose impedance is reflected back appearing to be the plate resistance. This is shown in Figure A19.24.

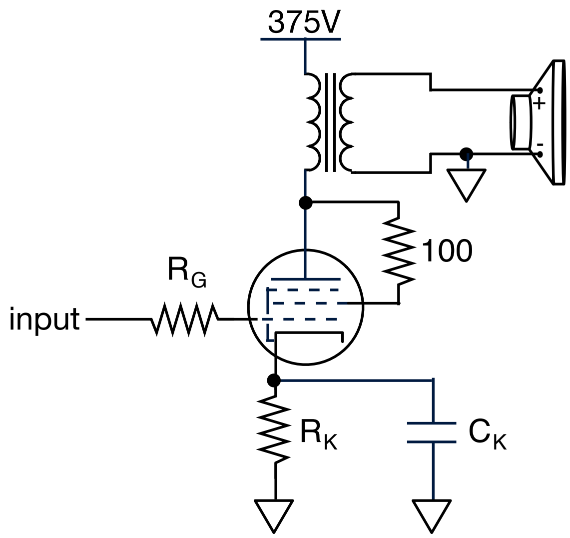


Figure A19.24: a pentode amp, running in triode mode is coupled to a loudspeaker through the output transformer whose primary coil acts as the amplifier's plate load

It should be noted that the design methodology for selecting a proper output transformer is quite involved and requires many other parameters and calculations to ensure safe operation for the tube. Please do not think it is trivial. Here we are only discussing it so far as we need for plugin emulation.

A19.8.1.1 Effects of the Output Transformer

Modeling the output transformer can be done in several ways, from SPICE simulation to simple filtering. The output transformer will add a band pass filter response to the signal; its coils and construction create parasitic capacitances and mutual inductances and the wire itself has a DC resistance, so the transformer turns into a complex filtering device. If you do some googling, you can find typical band-pass filter frequency responses of commonly used transformers. When the transformer is being driven with a moderate signal and is operating in its linear region, its main signal processing will be this band-pass filtering operation.

When the transformer is overloaded due to a high-gain amplifier such as those common today, it goes into magnetic saturation. The primary coil's magnetic field strength reaches a limit and it can't be made any larger, therefore no added signal can be transferred to the secondary coil.

True Story about Transformer Emulation

A very well known plugin manufacturer, whose main products are advanced emulations of classic gear, has an emulation/model for a classic, early compressor that is designed with tubes and multiple transformers, used for internal circuit coupling as well as output coupling. This particular piece of hardware is highly prized for its compression and of course 100% analog circuitry.

During the hardware analysis phase, this company found that all of the transformers had been over-engineered for the product. The transformers were much larger than needed, and were operating so far into their linear regions that they contributed no substantial nonlinearities. Their bandwidths exceeded the product's own requirements.

In this case, the manufacturer chose to simply not model the transformers because there just wasn't enough of a signal processing contribution to warrant wasting the CPU cycles and memory.

A19.9 Gibson GA-5 Pure Class-A Tube Amp

The combination of a single 12AX7 preamp tube and a single pentode (6L6-GC or 6V6-GC) as the power amp is a classic design for some of the earliest low power guitar amps. These ran in pure Class-A fashion and could generate around 5W of power. The triode preamp could easily produce a 40V signal required to drive the pentode directly. The output of the power amp tube was coupled to an output transformer to deliver power to the loudspeaker. Negative feedback applied as a forced degenerative feedback into one of the preamp tubes stabilizes the amplifier and reduces distortion and noise. Figure A19.25 shows the Gibson GA-5 amplifier from input to output; this circuit is similar to the early Fender Princeton 5F2 circuit.

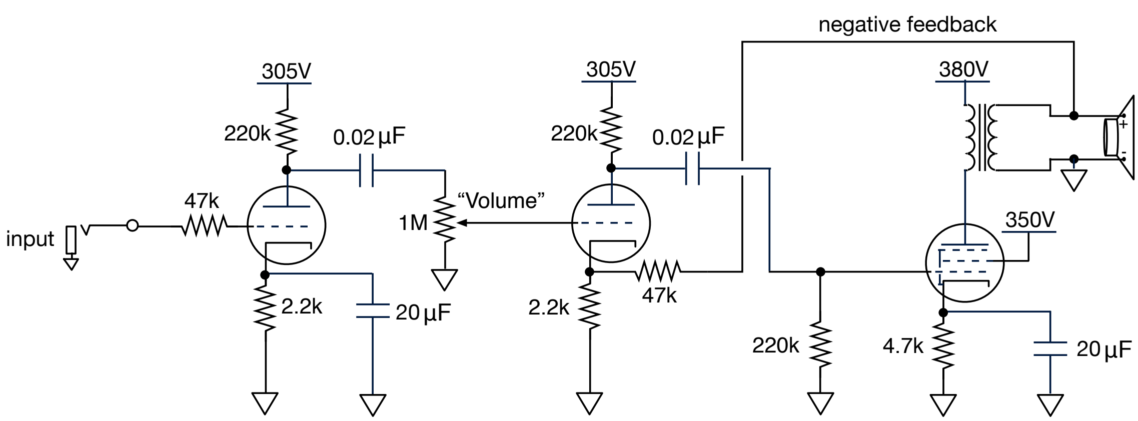


Figure A19.25: the complete Gibson GA-5 amplifier (second identical input jack and 47k grid stopper not shown); the triodes are packaged as a dual 12AX7 and the pentode is a 6V6

With moderate input levels and volume settings, this amplifier will produce very soft clipped waveforms and an overall "clean" sound. The FX books wave shapers and filtering would work nicely here because the amount of nonlinear processing is small. The output transformer could be modeled as a band pass filter, but at low levels it won't be driven into magnetic saturation. Figure A19.26 shows the SPICE simulation of the GA-5 amplifier at three low level settings for the Volume potentiometer. Notice how the bottoms of the waveforms are just starting to flatten out with the Volume control at 30%. This particular type of flattening (clipping) is not the same as the cut-off clipping that we saw before and will be the topic for the 2nd part of the document. We can see that it is just starting to happen as we open up the Volume control. The spectrum on the right shows the harmonic distortion with the Volume control at the 10% position. The other two settings also produce relatively mild harmonic distortion. Again, the wave shaping and filtering methods from the FX book should be able to handle the mild distortion here.

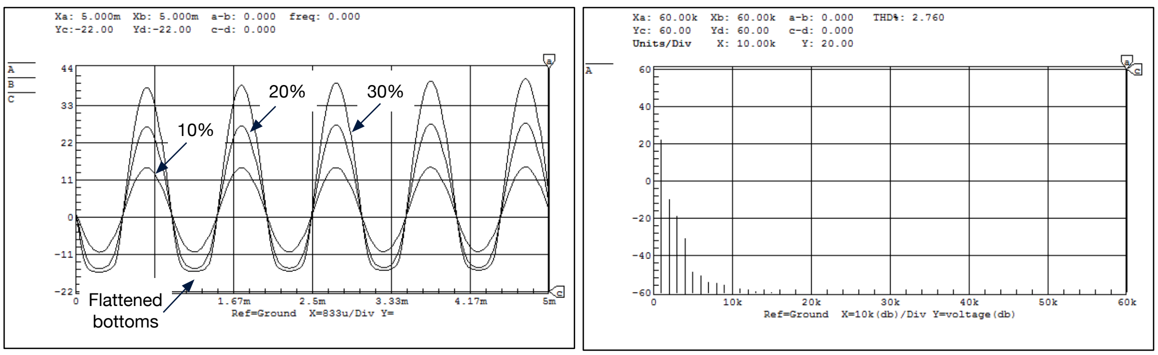


Figure A19.26: the SPICE simulation for the GA-5 amplifier (left) time domain outputs for three settings of the Volume potentiometer and (right) the spectrum with the 10% Volume setting

A19.10 Push-Pull Amplification

Push-pull is a term used to describe an amplifier design topology that intends to increase the power gain above the traditional amplifiers we've studied so far. Some people associate push-pull with Class-B amplification, which we'll cover next. However, you can also design a push-pull Class-A amplifier and for those Pure Class-A amplifiers such as the Mesa-Boogie Lonestar Special ® the push-pull method will allow for higher power outputs.

All push-pull amplifiers are designed to work in a "complementary" fashion. This is slightly different for a tube amplifier compared with a transistor (solid state) amplifier so for this discussion we'll just refer to tube amplifiers. For tubes, push-pull circuits start with a phase-splitter. This circuit produces nearly identical copies of the normal signal and its -180 degree out of phase "opposite." We sometimes call these two the hot and cold signals. The idea is identical to balanced audio connections that carry the normal and phase-inverted versions of the same signal.

A19.10.1 Class-A Push-Pull Amplifier

For the Class-A case, a push-pull amplifier can also be referred to as a differential drive system. In a standard *single ended* system like we've studied so far, the loudspeaker load has one side connected to ground. The amplifier drives a single signal into one side (end) of the load. In a differential drive or Class-A push-pull system, the amplifier produces two outputs and drives each side of the load with an identical but out of phase signal. In this case, one side of the load is NOT grounded – both sides are driven as shown in Figure A19.27.

The differential or balanced output has an interesting effect on the power output. Because we can calculate power in terms of the voltage ***V*** delivered to the load, and the fact that driving the load differentially effectively doubles the voltage across it, then the power will quadruple as a result. So push-pull Class-A amplifiers can switch from having a 5W single ended configuration to a 20W push-pull configuration if the manufacturer designs in that ability. You must be careful about the load, however, as one side of it must not be grounded.

$$P_{SE} = \frac{V^2}{R_L} \qquad P_{DIFF} = \frac{(2V)^2}{R_L} = 4\frac{V^2}{R_L} \qquad \text{[A19.5]}$$

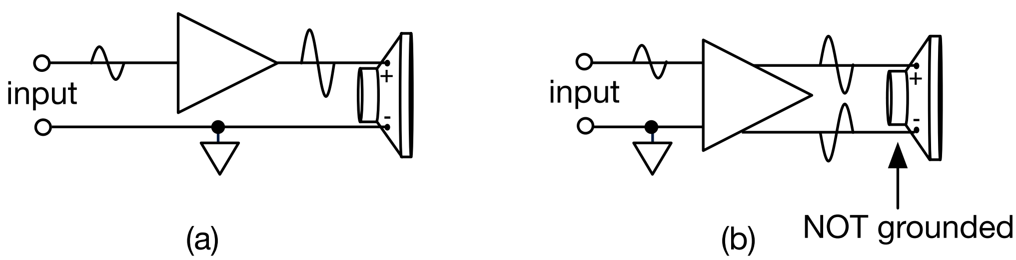


Figure A19.27: (a) single ended amplifier grounds one side of the load while (b) a differential drive amplifier drives both sides of the load with inverted versions of the same waveform

For a push-pull Class-A amplifier, you use a phase splitter to create the balanced or differential signal. Then you amplify each through its own output tube. The two tubes share a common plate load in the form of a center tapped transformer with the B+ voltage applied at the center tap. The loudspeaker is connected to the secondary of the transformer and is not grounded.

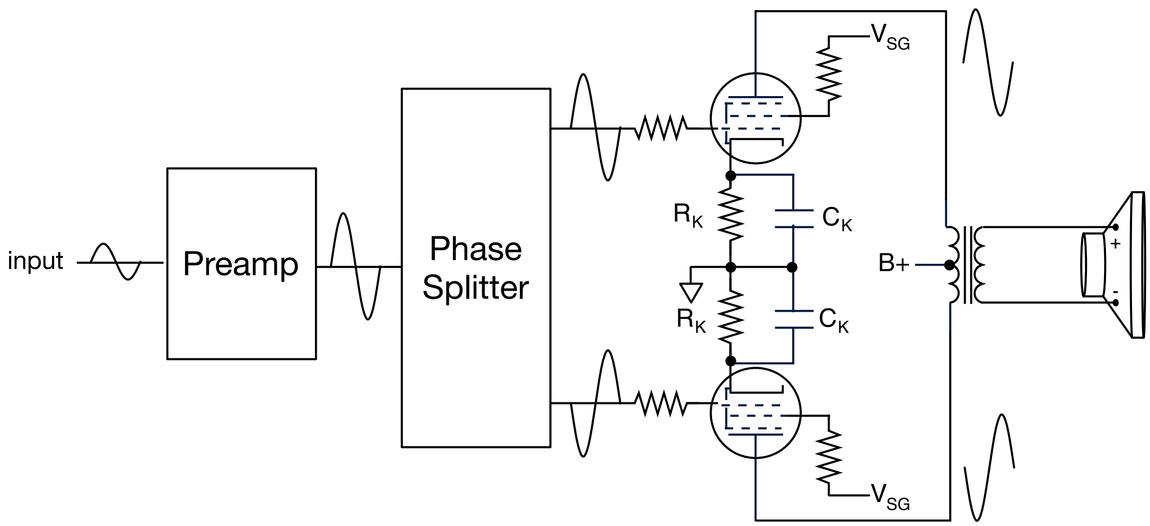


Figure A19.28: the block diagram for a push-pull Class-A amplifier; notice how the lower tube is drawn flipped, a common device to make the schematics look nicer

Other than the output connection, this amplifier configuration is essentially two identical Class-A amps connected to a common load. The previous design approach with load line, biasing resistor, cathode bypass, etc. is all the same, just done twice and affording a higher output power.

There are two commonly used phase splitter circuits: the one-triode *cathodyne* and the two-triode *paraphase splitter*. My 1974 silver-faced Princeton Reverb amp uses the cathodyne, while my 1996 Mesa Boogie DC-2 uses the paraphase splitter. Both phase splitters are imperfect and do not put out true, perfectly inverted outputs. The cathodyne can't create voltage gain whereas the paraphase splitter can. In addition the paraphase splitter has a negative feedback entry point, allowing a negative feedback path into it, rather than one of the preamp tubes. While the phase splitter is important, there doesn't seem to be much interest in modeling it. And, though the paraphase splitter can be designed to add gain, it is generally not thought of as a source of harmonic distortion the way the cascading triode preamp is. That said, both phase splitters are Class-A triode circuits and therefore they do contribute some nonlinear amplification.

A19.11 Class-B Amplification

Class-B amplification is always done in a push-pull manner. In this case the phase splitter is employed and the phase-inverted signals are delivered to a pair of output tubes. The difference is in how the tubes are biased. For Class-A operation, we bias the tube so that the input is fully amplified across a range of grid voltages, V_G. We look at the grid lines, and choose our bias point to be somewhere in the center of either the input range of values (maximum input swing) or the output range of voltages (half-plate voltage). In either case, current is always flowing in the tube unless it is in the cutoff state, and with no input applied, somewhere around ½ of the maximum current will be flowing. This is the essence of Class-A: the tube device is always conducting for 100% of the signal's voltage swing (with the allowed exception of being in the cutoff state).

For Class-B operation, we direct-bias the amplifier right at the cutoff point so that when no input is applied, no current is flowing. This means that the Class-B tube can only amplify the positive portions of the input waveform. It will only conduct current 50% of the time for a sinusoidal input. There is no cathode self-biasing resistor as the cathodes are held at 0V (ground) and the input is biased down to the cutoff point.

This means we need two tubes and the phase splitter. Each tube will amplify ½ of the signal. As with the Class-A approach, the outputs of the two tubes share a common plate load, which is likewise a center tapped transformer with the B+ voltage applied at the center tap. In this case the loudspeaker is attached in a single-ended configuration with one side grounded. This forces the transformer's differential input to invert and sew together the two halves of the waveform.

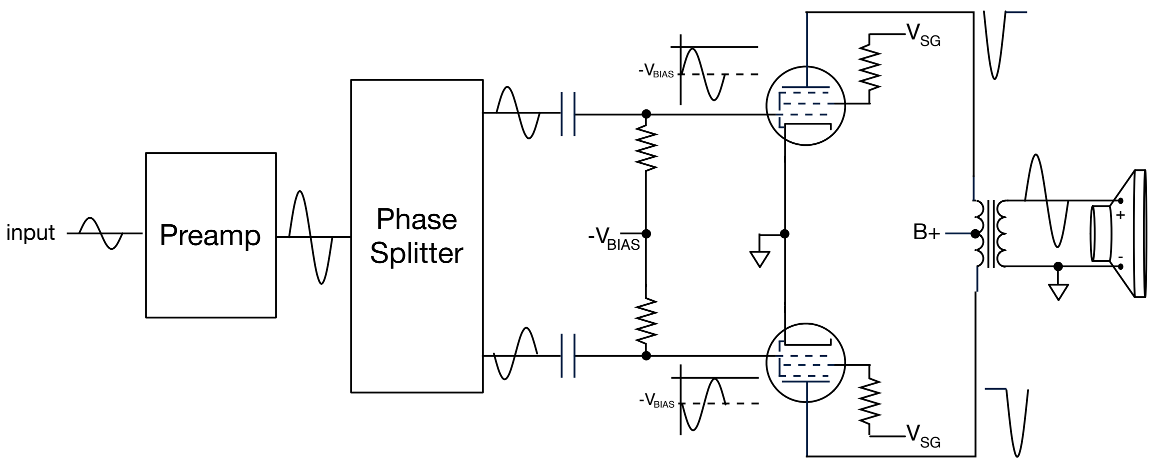


Figure A19.29: block diagram for the Class-B push-pull amplifier with the output tubes biased at cutoff, only amplifying the top portions of the input, which are flipped a the plate (output); the transformer configuration flips one of the inputs and then sews them together to create the output

A19.11.1 Crossover Distortion: Part 1

In Class-B systems, the tubes are biased just at the cutoff point and each will amplify (conduct) as required when the grid-cathode voltage changes. For input signals that lie on the proper range of grid voltage values and never exceeding these bounds, the output is nonlinearly amplified, but only over half its swing. The second tube amplifies the opposite polarity portion of the waveform, but it does so with the same nonlinearity as the first tube. This means that Class-B amplification, while still nonlinear, is generally *symmetrical* and affects the positive halves of the waveform in the same nonlinear manner as the negative halves. For emulation with wave shaping, this means using symmetrical wave shaper functions that mimic the same nonlinearity you observe in the grid voltage spacing on the plate voltage and current vs. grid voltage plot.

This is in contrast to the solid state Class-B amplifier with BJT devices. In this case, the devices are biased all the way off as well, however it takes at least one diode drop of voltage (+/- 0.7V) to get the transistor to conduct. So this small center-range of the output signal gets chopped out producing what is called *crossover distortion*. Regardless of the output signal's size, this small band of voltage will always be missing. This manifests itself as notches in the center part of the waveform.

A19.11.1.1 Just-Biased Class-B Amplifier

Figure A19.30 shows a SPICE simulation of a transformer coupled Class-B tube amp with 6L6-GC pentode power tubes and a 12AX7 triode paraphase splitter. The amplifier's input voltages are within the tolerable grid range so the tubes are not overdriven. When the bias voltage is set to -40V, the last V_G curve in Figure A19.20, the output (a) is smooth and continuous and has symmetrical nonlinear amplification – both top and bottom halves are distorted in the same manner which shows up as harmonic distortion of about 5% in (b). Notice that for this just-biased case, the signal does not show any visible crossover distortion notches at the center point and the waveform is smooth. This is in contrast to the solid-state Class-B crossover distortion that is omnipresent.

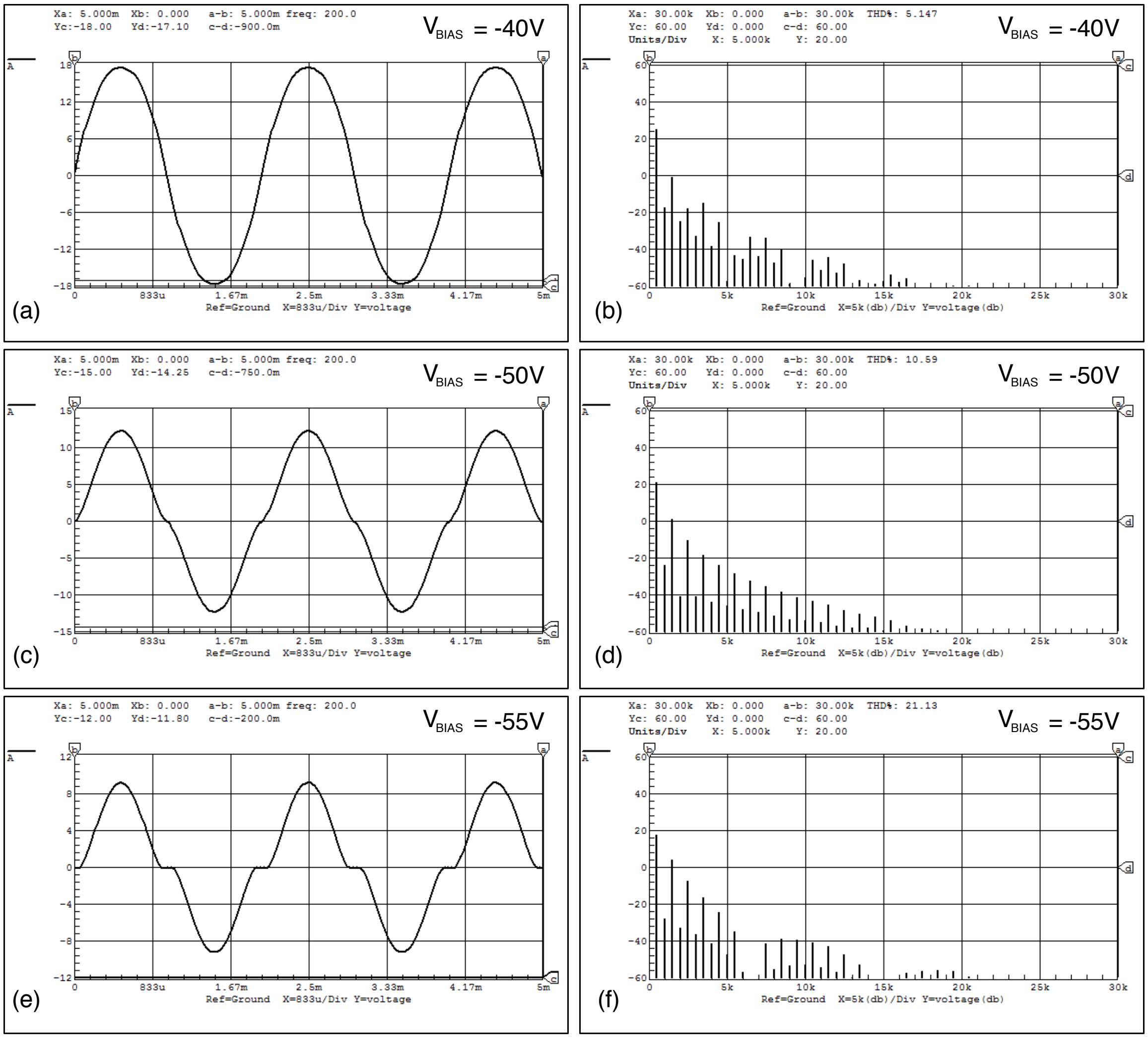


Figure A19.30: SPICE simulation of a transformer coupled Class-B tube amp with 6L6-GC pentode power tubes and a 12AX7 triode paraphase splitter, the simulations are made for three different bias voltages: -40V (just-biased), -50V (over-biased), -55V (very over-biased) for time (a), (c) and (e) and FFT spectrum showing harmonic distortion in (b), (d), and (f)

A19.11.1.2 Over-Biased Class-B Amplifier

If the bias voltage in this design is lowered to -50V, the amplifier is said to be *over-biased*. This means that the portion of the input grid voltage between -40V and -50V won't get amplified and you can see the results of this in both time A19.30 (c) and harmonic distortion of about 10% in (d). Notice how the spectrum now contains two sets of harmonic envelopes, one for the odd harmonics and the other for the even ones. It is clear that the 3rd, 5th, 7th, etc. harmonics are amplified more than the 2nd, 4th, and so on.

When the system is heavily over-biased at -55V, the crossover distortion is now visually evident in the time domain plot A19.30 (e). The harmonic distortion is now 21% (f) for this bias scheme. This is one kind of crossover distortion that can occur, but only when the Class-B tubes are over-biased significantly. A feature to notice here is that the tops and bottoms of the waveform are not altered, only the center portion where the distortion notches occur. We will soon study a second kind of crossover distortion that is accompanied by waveform clipping in the next part of the document.

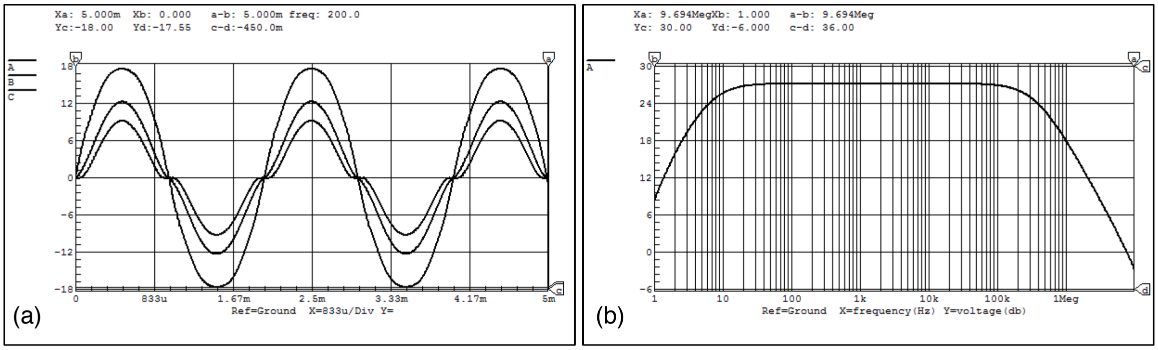


Figure A19.31: (a) shows the three time domain outputs at the three bias voltages used in Figure A19.30 overlaid, demonstrating how much of the signal is unamplified in over-biased systems and (b) shows the simulated frequency response of the just-biased Class-B amplifier

> Unless the output tubes are heavily over-biased, a Class-B amplifier driven with an input signal whose amplitude lies within the proper grid voltage range will produce an output that is fundamentally free of crossover distortion and, though nonlinearly amplified, will be symmetrically distorted in a soft-clip manner. This will produce an emphasized odd order distortion signal whose even-ordered components have lower amplitudes. The third harmonic is an octave plus a fifth above the fundamental and this type of distortion is associated with the Class-B amplifier.

A19.12 Part 1: Review

In this first part of the addendum, we've discussed basic tube amplifier circuit design. In the simulations and discussions, I've been careful to constantly remind you that we are dealing with properly biased circuits operating under proper conditions in which the tube input voltages are always within the design limits.

The FX book code and projects use simple waveshaping and filtering to emulate these systems. You should also note that just because a signal is soft-clipped and has no visible "chopped tops or bottoms" doesn't mean that its distortion isn't severe. On the contrary, cascading multiple soft-clipping amplifier modules can result in significant harmonic distortion without waveform clipping.

A19.12.1 Single Ended Class-A Triode and Pentode

Both the triode and pentode may be designed as singled ended Class-A amplifiers and this might be the most common format for triodes. Singled ended Class-A amplifiers deliver power to a load with one side grounded. We saw that both triode and pentode Class-A amplifiers produced harmonic distortion because they amplified the signal in a nonlinear manner. In addition to being nonlinear, this distortion is also asymmetrical in that the positive and negative portions of the waveform are amplified differently. The distortion is a result of the 3/2 power law that relates the output plate current to the input grid voltage. The slightly nonlinear curve that this equation generates is responsible for the distortion that occurs when the input signal is constrained to a proper range of values.

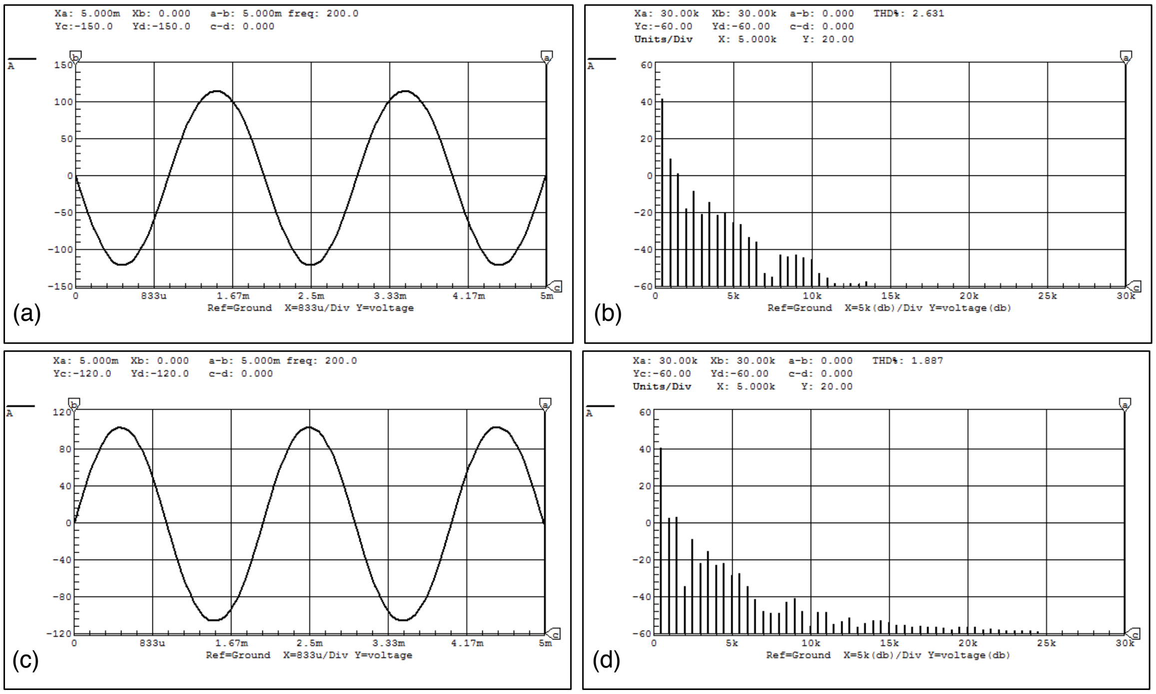


Figure A19.32: the outputs of the two tubes in the Gibson GA-5 preamplifier (a) shows the output of the first tube with $3V_{p-p}$ applied at the input and (b) is its spectrum while (c) shows the output of the 2^{nd} tube with the volume control at about 10% and (d) is its spectrum

Figure A19.32 shows examples of cascaded triodes where the input signal range has been kept right at the maximum input size the design allows. This is the very softly distorted but also “fat” sound that the tube in some devices delivers where the intention is not over the top distortion but a milder more transparent kind of signal enhancement.

A19.12.2 Push-Pull Class-A Pentode

The Class-A pentode may also be run in a push-pull configuration. Its design is not much different from the standard Class-B output stage, except we lower the bias voltage to be in the center of our allowed input grid voltages. Figure A19.33 shows the SPICE simulation for a 6L6-GC based push-pull Class-A design coupled into a loudspeaker via a transformer. Notice the symmetrical distortion here – the output is driving the load differentially with two opposing signals so the resulting combination will be symmetrical. Notice the interesting effect in the spectrum in (d) when we increased the input size into the Class-B tubes. You can see that the even harmonics virtually vanish as we hit the maximum legal input size. This reveals that the nonlinear amplification is almost exactly symmetrical for this input size. To emulate this with a waveshaper, you would need to find the appropriate symmetrical function that gives you the harmonic envelope you desire. Note the lack of even harmonics in Figure A19.33 (b) – this is because the waveform is undergoing almost perfectly symmetrical distortion.

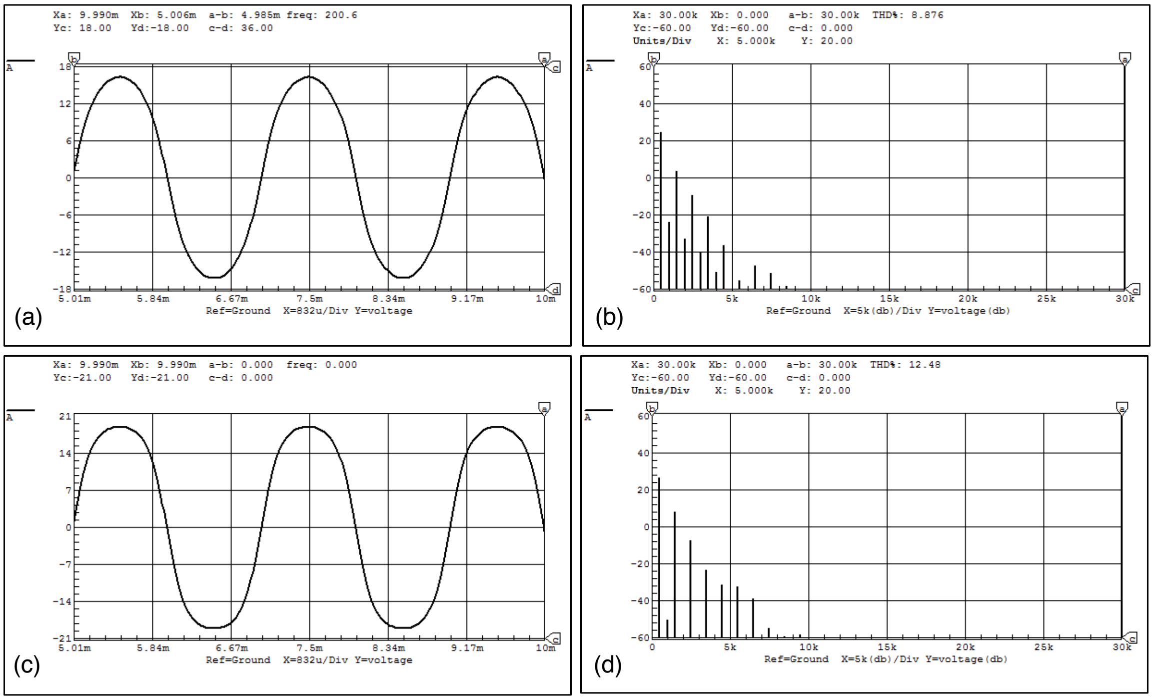


Figure A19.33: SPICE simulation of the output in time and frequency spectrum for a push-pull Class-A amplifier with an input size of (a), (b) $40V_{p\text{-}p}$ and (c), (d) $50V_{p\text{-}p}$

A19.12.3 Push-Pull Class-B Pentode

The more common arrangement for push-pull amplifiers is the Class-B power stage. In this case, the two tubes are identically biased so that they are right at the cutoff point and the each amplify ½ of the input waveform. We saw in Figure A19.30 that as long as the amplifier is not over-biased, an input size on the correct range will produce an output that is smooth and does not include crossover distortion notches. However, over-biasing the amplifier will produce this crossover distortion, as the amplifier is shut-off for longer periods of time during which there is no amplification.

Part II: Fundamentals of Overdriving the Grid

With all of the legal notices about maintaining proper input size to the tube stages, you must be wondering what happens when we violate those conditions and purposefully overdrive the tubes, sometimes massively. The result of overdriving the inputs of the tubes this way depends on the topology. For cascading triode preamps, overdriving the tube stages in succession produces massive harmonic distortion and clipped waveforms with a preponderance of 2nd harmonic (even) distortion. This is the fundamental rule for high-gain guitar preamps some of which can contain up to six or eight triodes cascaded together. For the Class-B output stage, overdriving the power tubes creates what is sometimes called "tube compression" or "tube limiting" and guitarists can actually sense and feel this compression happening as they play. The resulting distortion is symmetrical and includes more of the 3rd harmonic distortion than the triode Class-A circuit. This only happens in the Class-B output; the push-pull Class-A output has a similar but not identical sonic feature that we can emulate along with the Class-B version. We'll discuss that at the same time as the Class-B emulation.

A19.13 Overdriving the Grid

In both Class-A and Class-B cases, overdriving the tubes means overdriving the grid voltage and pushing it outside of its design limits. We can overdrive the input to the tube in two ways, by either violating the upper input voltage limit value, or the lower one. It turns out that these two situations have very different results in the tube and in our emulation of it as well.

A19.13.1 Overdrive when V_{GK} is Too Negative

When V_{GK} hits is lowest negative limit (this was -3V for the 12AX7 triode circuit we studied and -40V for the Class-B pentode design), the negative voltage repels the space charge at the cathode, effectively holding the electrons down and preventing current from flowing in the tube. When the input voltage causes V_{GK} to become even more negative than this cutoff point, the triode cannot be "even more cut off than 100%" and it simply stops amplifying since no more current is flowing. This produces a clipped output signal. Since the triode inverts the signal, the clipping appears on the tops of the output waveform. We saw this happen in Figure A19.15 when we added the cathode bypass capacitor and realized that $3V_{p-p}$ was actually slightly too large as our maximum input signal size. This kind of clipping is called *cut-off clipping* because the tube is in the cut-off state and no current flows during this time. This clipping is less important for us as far as tube circuit emulation goes because we can easily implement it as a moderately soft clip at the waveshaper. This means that the waveshaper's lower limit value just stays flat (or ever so slightly angled upwards) as our input signal's level falls below its clip point of -1.0.

A19.13.2 Overdrive when V_{GK} is Positive (greater than 0)

When we overdrive the input of the tube by allowing the grid-cathode voltage to become positive or greater than 0.0, everything changes in our amplifier's operation. This is the condition that is not tolerated in high-performance, low noise preamplifiers and other circuits where distortion is the main enemy. We are going to be spending the much of the document dealing with this condition and what it creates within the tube circuit. Figure A19.34 (a) shows an overdriven input for our 12AX7 design from Part I where the bottom portions cause cut-off clipping and the top portions cause *grid-conduction*.

A19.13.2.1 Grid-Conduction

When we allow the V_{GK} to go above 0V by overdriving the tube in the positive direction, it now appears to the cathode as another anode, in addition to the existing plate (anode). When you connect a cathode and anode together in which current can only flow in one direction, you create a diode. So, when $V_{GK} > 0V$, a diode appears that connects the input (grid) to the cathode. When this happens, the tops of the input waveform that cause $V_{GK} > 0V$ will begin to be clipped off because of the diode, which acts like a one-way diode clipper, removes only the tops of the waveform. This is called grid-conduction and it creates the other type of waveform clipping in the tube. However, the big difference between this clipping and cut-off clipping is that this clipping occurs at the ***input*** to the tube, not inside of it. In Figure A19.34 (b) you can see a massively overdriven input to the tube. In Figure A19.34 (c) the diode appears for positive signals only that clips it to the diode voltage, which is about 3V here (the maximum V_{GK} value allowed by design).

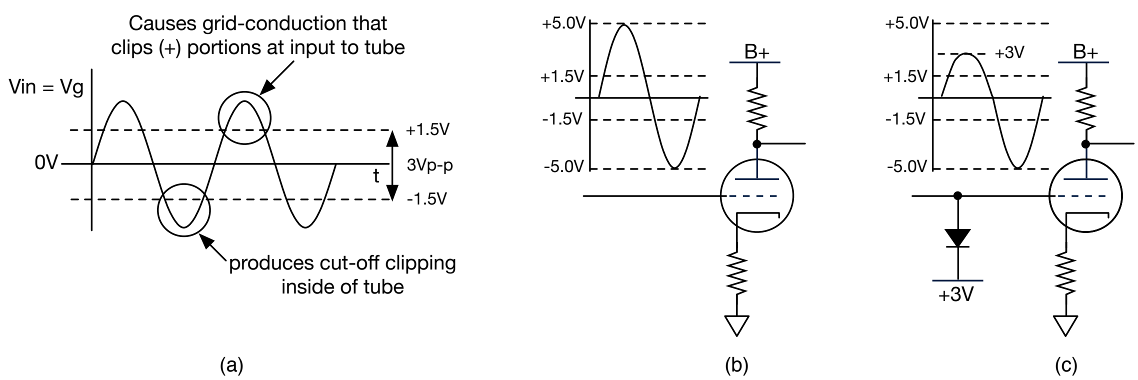


Figure A19.34: (a) when the input to the grid exceeds its design limits, the bottom portions cause cut-off clipping in the tube while the top portions cause grid conduction clipping at the input (b) heavy overdrive of input with $10V_{p-p}$ sinusoid with signal shown before grid conduction and (c) the diode has appeared at the input that clips it to about +3V in this example

Figure A19.35 shows a SPICE simulation for the 12AX7 tube circuit from Figure A19.16 but with the input overdriven with a $10V_{p-p}$ sinusoid. You can see something odd about the clipping. It is not clipping the waveform right off, and not even in a soft-clip either. The tops of the waveform look almost like the tops of houses.

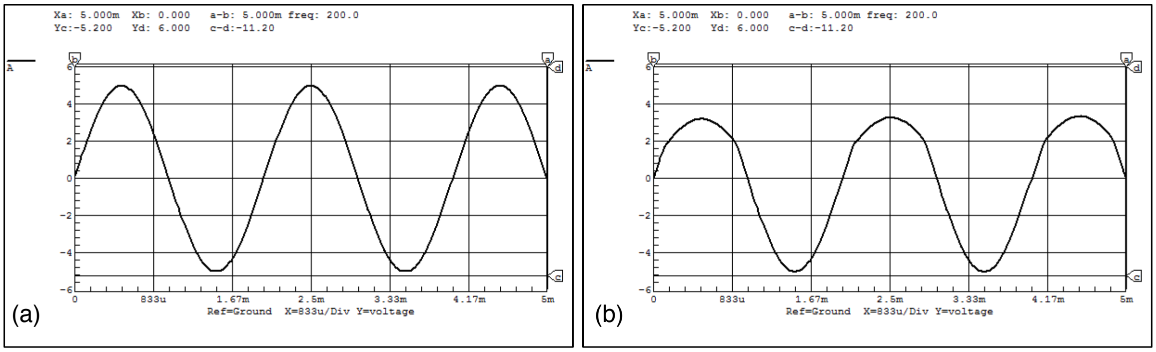


Figure A19.35: (a) the 10Vp-p input to the 12AX7 tube on the input side to the 68k grid-stopper resistor and (b) at the grid directly at tube input

Now examine Figure A19.36 (a) that shows the output of this tube stage. The top portions have been clipped due to cut-off clipping – due to inversion, these clipped tops were originally the bottoms of the input signal. The output's lower side looks a bit stranger. The peaks have been elongated in the negative directions; the "house-tops" have become more exaggerated; the output now occupies the complete plate voltage range, swinging from 0V all the way up to 200V. Figure A19.36 (b) shows the FFT spectrum with 48% total harmonic distortion. The harmonic envelope is complex but does show the pattern of a signal whose duty cycle is not 50%. If you measure the halves at the DC offset of 165V, the bottom portions have a slightly shorter duration than the tops.

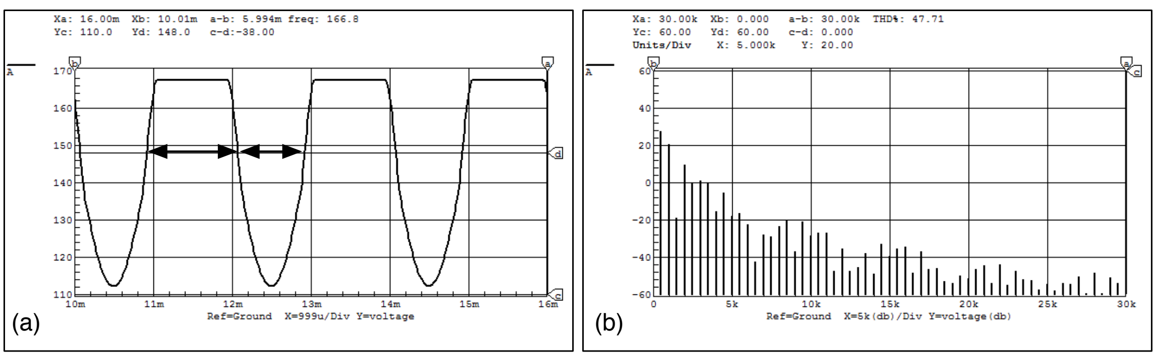


Figure A19.36: (a) the output of the heavily overdriven 12AX7 tube has a very strange shape with cut-off clipping on the top and grid-conduction clipping on the bottom (b) the FFT spectrum is complex and harmonically dense

There are some things that don't seem to add-up correctly here. On the input, why are the tops not fully clipped, or at least clipped with soft corners? On the output, why are the bottoms of the waveform so stretched out? Notice also that at the DC operating point of 148V, the top portion of the waveform is wider than the bottom portion. The duty cycle of the input waveform has changed.

Before moving on, consider what happens when we increase the input to $20V_{p\text{-}p}$ and $40V_{p\text{-}p}$ and shown in Figure A19.37. The spectrum is harmonically rich for the $20V_{p\text{-}p}$ case in (b). Notice how large the 2^{nd} harmonic is in relationship with the fundamental – they are almost the same amplitude. Increasing the input further creates hard clipping on both halves of the waveform as the tube rails out at its limits. The more interesting thing happening is put into context in (d) which shows the three outputs overlaid with the DC operating points of each labeled on the right. You can see that as the input increases, the output spends more and more time in the positive swing and that the duty cycle of the input becomes shifted. Also note that in our standard 12AX7 circuit, the output can swing all the way to about $200V_{p\text{-}p}$ when we include the positive grid-conduction.

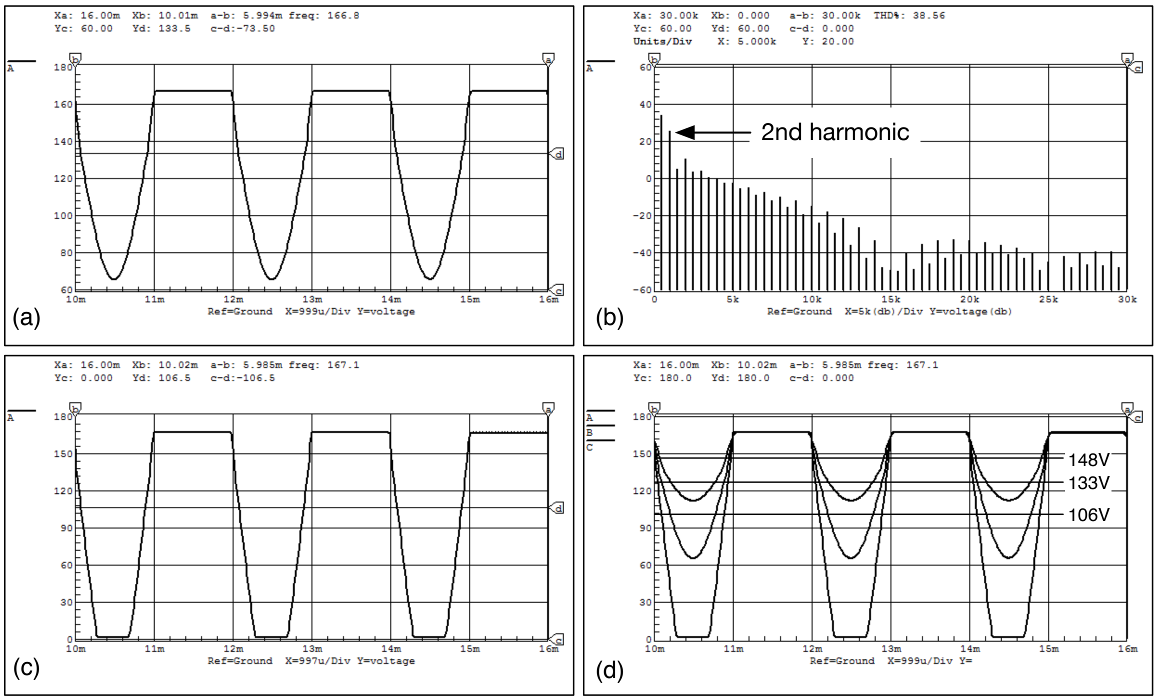


Figure A19.37: when we heavily overdrive the 12AX7 even further, we observe that the duty cycle of the resulting square wave is not 50%; for a 20Vp-p input, the time domain output is shown in (a) while its spectrum is shown in (b). Increasing to 40Vp-p yields the output in (c) and (d) shows the three outputs overlaid – note that at the maximum input the output swings almost the full 200V

A19.14 Class-A Dynamic Nonlinear Amplification

The effects that we observe in the SPICE experiments of the last section are due to dynamic nonlinear amplification in the tube, which happens if we allow V_{GK} to become positive. The order of operations is as follows:

1. V_{GK} goes above 0V and a diode forms that clips the input tops
2. Clipping the tops of the input waveform before they enter the tube gives them a net negative (-) DC offset, due to the loss of the top portions
3. The input signal now contains its own inherent negative DC bias which is combined with the existing negative DC bias at the tube input
4. The tube is now over-biased and the DC operating point has now shifted down, due to the added negative DC offset that the clipping caused

Figure A19.38 shows what is happening here for the Class-A triode case. In the Figure, (a) and (b) show the results of our 12AX7 with a proper input range of 3Vp-p. Note that in (b) the 3/2 power law VTF has been inverted to show that behavior of the tube as well. That 3/2 power law VTF is a wave shaper in code. In (c), overdriving the grid causes the input tops to become clipped off producing an over-biased signal to the tube input due to the added negative bias voltage that is added. The bias shift is shown in (d) where the operating point has dropped to -2.5V. Notice that this causes the negative portions of the waveform to undergo severe cutoff clipping the waveform is shifted to the left relative to the transfer function curve. The top portion, clipped at the input in a soft manner, then has room to be stretched out nearly linearly by the rest of the VTF.

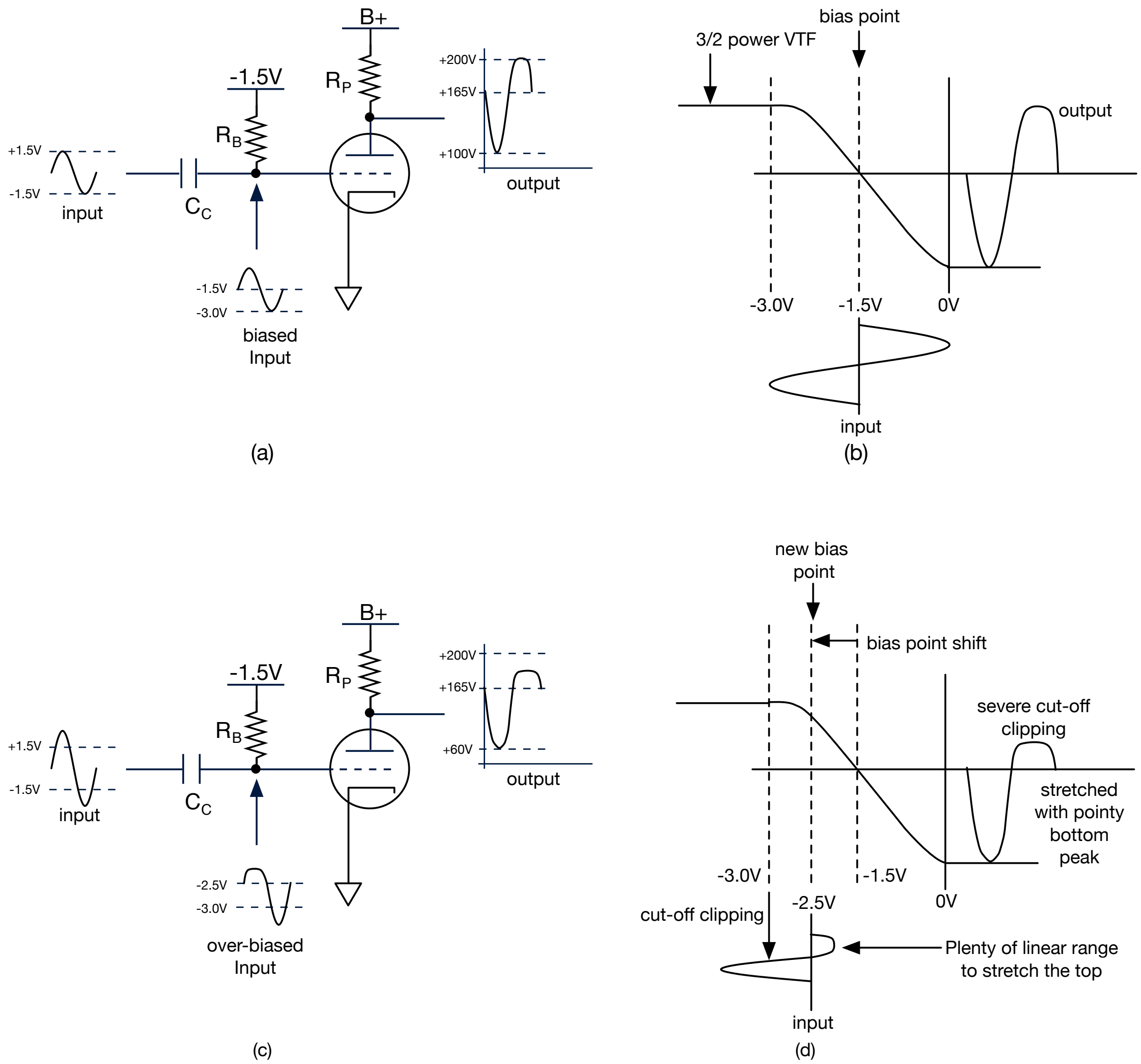


Figure A19.38: this is a graphical representation of the DC bias shift that occurs when we overload the triode (a) normal input range and (b) normal output (c) overdriven input and (d) output after DC bias shifting has occurred

As the input signal is increased, the stretching of the positive portion we observe in Figure A19.38 (d) becomes more intense. Remember that the inversion will cause this positive portion to become the negative portion at the output. Because of the negative DC bias shift, the input signal will spend more of its time in the cut-off zone below -3V.

When this portion of the signal is inverted at the output, it means that the positive portion will now be wider than the negative portion when we examine the waveform taking into account its DC offset. This is where the change in duty-cycle comes from – it is the DC level shift that causes it to happen. As the input is increased up to its maximum of about $100V_{p\text{-}p}$, the DC level shift becomes worse, and the duty cycle of the output waveform becomes very low. This is a part of the tube sound that guitarists like.

> The thing that makes this far more complex than the simple cases in the first part of the document is that this DC level shifting is dynamic and constantly changing based on the amplitude of the input waveform. The more positive portion that gets clipped, the more of a negative DC bias the signal takes on, shifting the operating point. It is important to understand that this is dynamic. As soon as the tops of the input begin to be clipped, the DC offset of the signal changes, causing it to slide down. This means that if we want to emulate this behavior in code, we will need a way to track either the input amplitude or the DC offset of the input as the tops are being clipped. That information can then be used to shift the DC operation point of our wave shaper in code.

A19.15 Class-B Dynamic Nonlinear Amplification

When we overload the inputs to a Class-B amplifier, the same thing happens as far as the diode and the positive peak clipping. However the resulting signal and waveform are entirely different from the Class-A case. This effect is sometimes called tube compression but it has nothing to do with compressors in the normal sense. Figure A19.39 shows a pentode operating as ½ of a Class-B circuit. It has an input signal that just fits within its input range and it is biased to the cut-off voltage of -47V. In (b) you can see the bias point or DC operating point and how the input signal lines up with it; only the top portions of this waveform may be amplified. The composite output in (c) shows some nonlinear amplification but no visible crossover distortion.

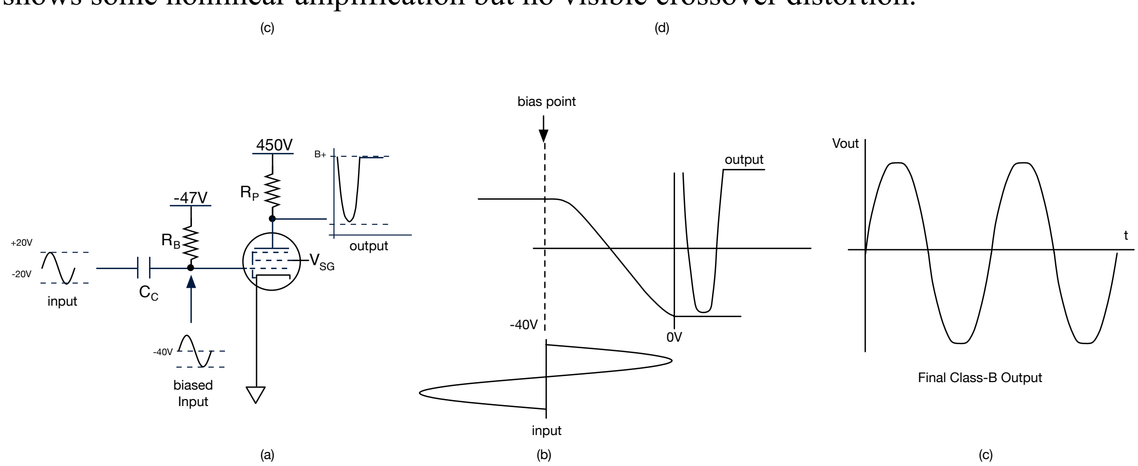


Figure A19.39: (a) ½ of a Class-B circuit is shown with an input that is just on the proper range of values (b) the bias point is just at cutoff, so only the top portion of the waveform is amplified (and inverted) (c) the output of the complete Class-B amplifier shows no visible crossover distortion

Figure A19.40 shows what happens when the input is increased to a level outside this range. The sequence of operations that happen when we overload the input to a Class-B tube is as follows:

1. V_{GK} goes above 0V and a diode forms that clips the input tops
2. Clipping the tops of the input waveform before they enter the tube gives them a net negative (-) DC offset, due to the loss of the top portions
3. The input signal now contains its own inherent negative DC bias which is combined with the existing negative DC bias at the tube input
4. The tube is now over-biased and the DC operating point has now shifted down, due to the added negative DC offset that the clipping caused
5. Since this is a Class-B amplifier already biased at cut-off, over-biasing the stage will shift the input down further so that a portion of the input signal will not be amplified
6. The un-amplified part will appear as notches between the waveform cycles and the tops and bottoms of the waveform will be soft-clipped from the action at the input

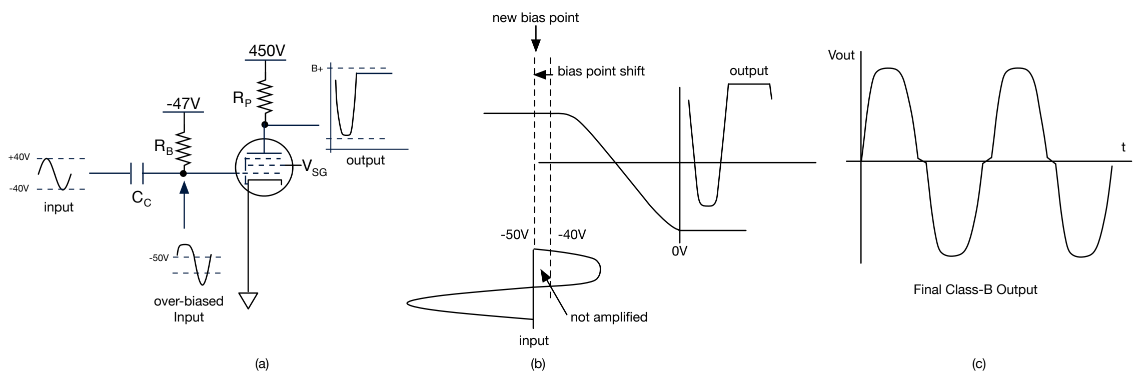


Figure A19.40: (a) ½ of a Class-B circuit is shown with an input that far exceeds the input limits (b) this produces a bias shift downward, but the bias point is just at cutoff, so only a part of the top portion of the waveform is amplified (and inverted) while a section of the top does not receive any amplification at all (c) the output of the complete Class-B amplifier shows visible crossover distortion and a flattened waveform

A19.15.1 Crossover Distortion: Part 2

You can see from the bias point shift in Figure A19.40 (b) that a part of the input will not be amplified. As the input increases, the peak-to-peak output level will eventually max out and the tops will begin becoming more squared off. All of this time, the crossover distortion has been getting worse and worse. The inside portion of the signal gets compressed into the tiny region around the crossover distortion notches. This creates the effect of signal compression and limiting, in a completely different manner. You can think of it as compressing from the inside because of how the center part of the waveform is removed, more and more as the input increases. This is where the term “tube compression” comes from. The signal really is compressed and can’t grow beyond a certain amplitude. Notice that this is a different operation from the over-biased amplifier, where the waveform did not show any clipping because we held the input to a reasonable size. When a Class-B amplifier is already over-biased to begin with, the grid-conduction condition is only exacerbated that much more and the DC bias shift produces more compression and crossover distortion.

Figure A19.41 shows the SPICE simulation of a Class-B amplifier driven hard into crossover distortion with squaring off of the waveforms. These are continuations of Figure A19.40, which shows the process at the beginning of the operation. You can see that the amplifier goes into a signal limiting condition, as the output can never grow past a certain point. Notice also the interesting shape of the distorted wave halves, which have a soft trapezoidal contour. The crossover distortion's notch angle also changes and becomes more severe and flattened as the input increases. Another interesting thing to note is how the even harmonics begin to disappear as the input level increases and the two halves become more perfect square(ish) waves. Notice how the even harmonics disappear as we increase gain and increase the symmetrical nonlinear distortion.

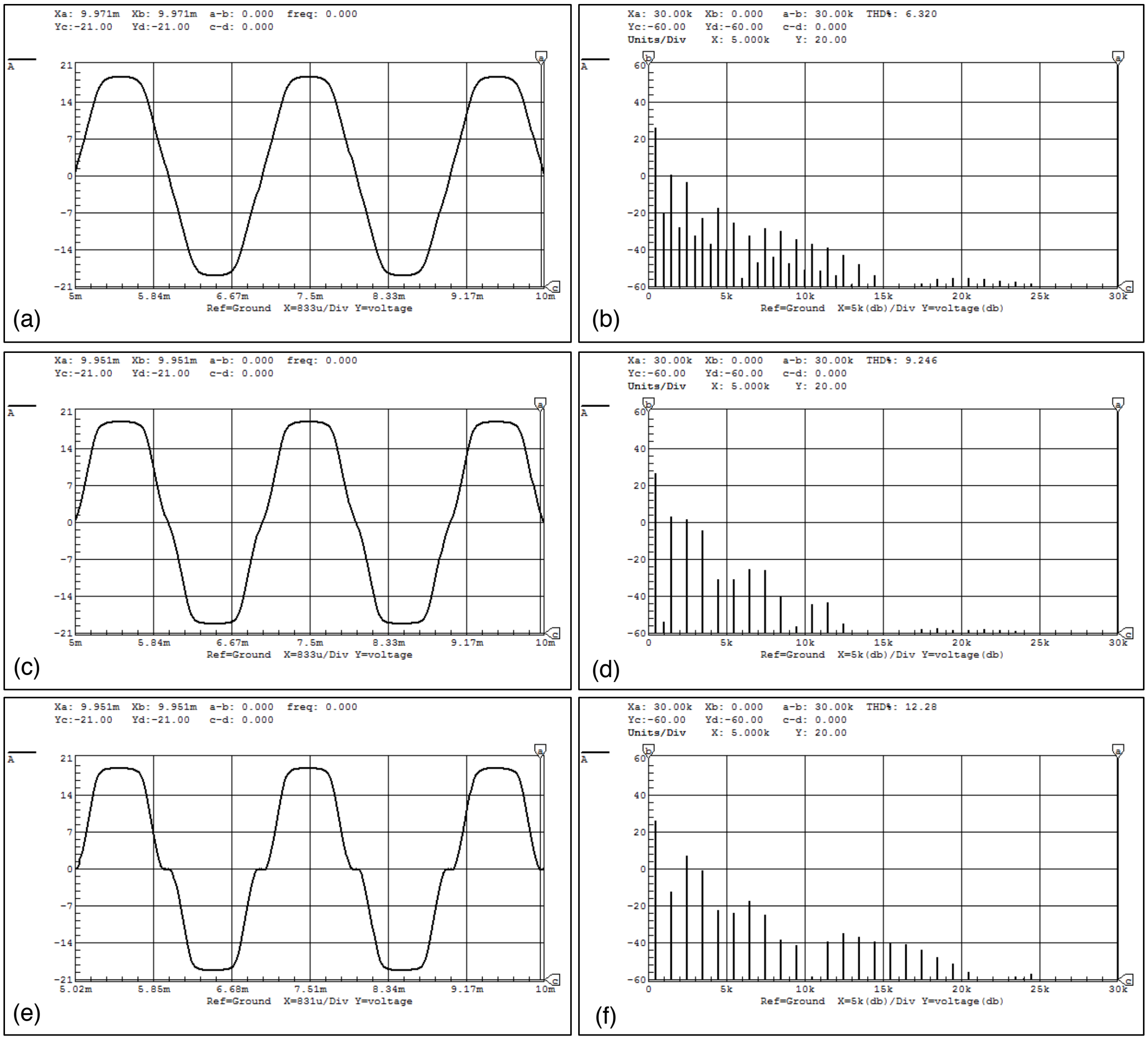


Figure A19.41: SPICE simulation of a Class-B 6L6-CG amplifier overdriven with input signal of $80V_{p\text{-}p}$ (a) and spectrum (b), $100V_{p\text{-}p}$ (c) and spectrum (d) and $120V_{p\text{-}p}$ (e) and spectrum (f)

A19.16 Push-Pull Class-A Dynamic Nonlinear Amplification

Before ending the section on grid-conduction overload, you might also think about the push-pull Class-A amplifier. In this case each amplifier is biased to Class-A and each drives one side of a differential load. When this amplifier is overdriven, we will get something that looks like Figure A19.41 (a) up until the point that the input size grows larger than the bias range allows. At this point, crossover distortion will occur as the clipped input signal effectively over-biases the amp into Class-B territory. Figure A19.42 show a succession of push-pull Class-A outputs with the input being further overdriven. Note that the spectra are very different up until the point that crossover distortion occurs. In effect, all we are doing is staving off the inevitable but we do produce a different output waveform and spectrum from a normal Class-B amplifier.

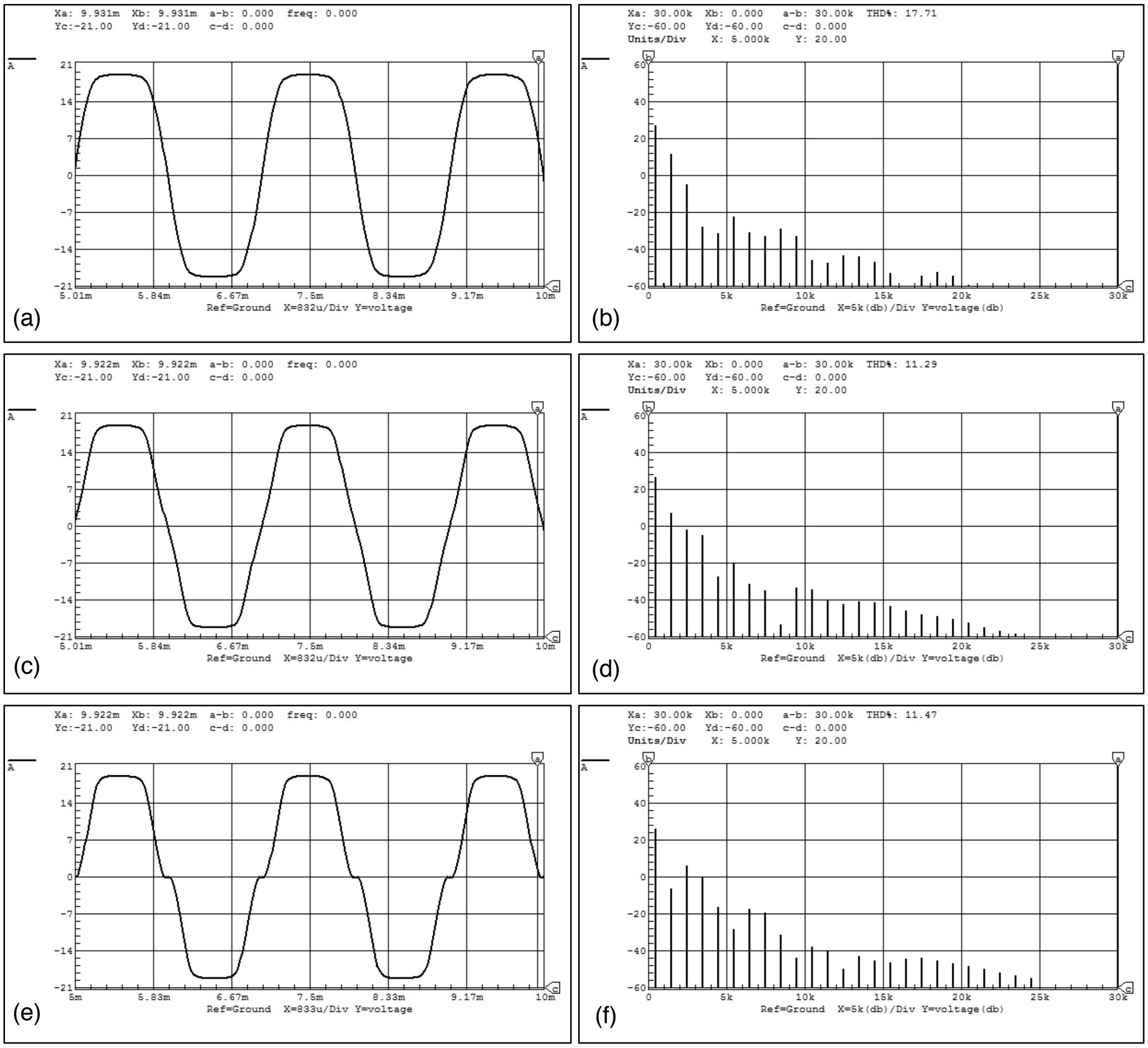


Figure A19. 42: SPICE simulation of a push-pull Class-A 6L6-CG amplifier overdriven with input signal of $80V_{p-p}$ (a) and spectrum (b), $100V_{p-p}$ (c) and spectrum (d) and $120V_{p-p}$ (e) and spectrum (f)

Part III: Digital Implementation Techniques

The first two parts of this document deal with the background theory on tube circuit design and analysis, and overdriving the tube inputs and the dynamic nonlinear amplification that results. If you've followed up to here, you might understand why I decided to limit the book to static nonlinear amplification. To fully grasp what you need to do for emulation, you really do need to understand what the circuits are doing and what dynamic processes are at play when the inputs are overdriven. And for anyone without a background in some kind of analog circuit design, it is understandably difficult to have this stuff thrown at you. But you can't become a better researcher and developer of your own tube emulation schemes if you don't understand the basic principles at work and I felt like unless that material was available in the way I wanted to present it, inclusion of dynamic nonlinear amplification would be daunting if not impossible for some readers. With this addendum, I can stretch out and try to present each of the pieces we need to deal with.

A19.17 Patent Study

You might notice that the last part of this document is full of patents. There are several reasons for this and if you haven't acquainted yourself with *googlepatents.com*, now might be a good time to start. It is difficult to find information on commercially available products for good reason – the manufacturer has no reason to publish their company secrets for the rest of the world to see. And anyone who has worked on these projects has likely signed an NDA prohibiting them from discussing anything, including the algorithms involved (myself included). If you go to *amazon.com* you won't find a book by Yamaha or Line-6 describing how their modeling software operates.

However, if they want to patent these ideas, then they have to disclose the essence of their discovery along with at least one preferred embodiment (example) written in plain language. Patents are designed to "teach" an embodiment of an invention and in doing so, they must also describe the processes, and even give reasoning on why their invention should be considered unique enough to be patent-able. In many cases this means referring to other patents or documents (called "prior art") that further their case in point. With googlepatents.com, you can easily search the prior art of a patent, and find out what patents their authors cited, and what newer patents cite them. You can drill up or drill down in time and the results can be astounding and very rewarding. There are three sets of patents included here.

A19.17.1 Peavey

Peavey invested heavily in designing solid-state amplifiers that emulated tubes and in the process they generated numerous patents. Studying how analog designers try to emulate tubes can only help you in your own approach. Their three TransTube patents cover the two topics in Part II involving overdriving cascading triodes (preamp) and overdriving a Class-B pentode (power amp). Their background discussions on these topics are fantastic and well worth reading. If you are an analog circuit designer, you will likely find them to be brilliant in their execution and the resemblance of the silicon based circuit architectures to the tube based circuits is evident and useful for digital implementations. The patents included are:

Pat: 5,524,055 *Solid State Circuit for Emulating Tube Compression Effect* [Peavey, 055]
Pat: 5,619,578 *Multistage Solid State Amplifier That Emulates Tube Distortion* [Peavey, 578]

A19.17.2 Yamaha

The Yamaha patent included here is quite interesting. Some of the DSP block diagrams are uncannily like the Peavey analog block diagrams and circuits. And, my own versions are also based on the Yamaha patent. This patent is fairly easy to understand however it is very light on tube details. There is a lot of information about processing the audio, but without explanation of why you would want to process it that way. This addendum serves as the bridge that will allow you to understand why they are undertaking the DC level shifting and other operations. In addition, it is clear that table lookups are used in place of wave shaping equations and that they do not disclose the contents of these tables, only that they are carefully chosen to produce the desired output/result.

Pat: 5,841,875 *Digital Audio Signal Processor with Harmonics Modification* [Yamaha]

A19.17.3 Poletti

The included patent application from 2008 by Mark Poletti appears to have been abandoned for pursuit as a patent (thus it is in application format, which is not quite the same), however it has some useful information especially regarding the waveshaper equations used for the Class-B emulation.

Pat App: US 2008/0049950 *Nonlinear Processor for Audio Signals* [Poletti]

A19.18 Improved Triode Class-A Model

We can improve on the triode object in the FX book by including the ability to model the tube under overloaded conditions. This involves modeling the input waveform clipping that occurs when $V_{GK} > 0$.

A19.18.1 Peavey Class-A TransTube

One Peavey solution is shown in Figure A19.43 where n-channel FETs replace the triodes. N-channel FETs require a negative DC bias, as with triodes. The addition of the diodes to ground on the FET gates causes positive waveform clipping, introducing a negative DC bias, which over-biases the FET and shifts the DC operating point down, as with the triode. The process is repeated through several stages with voltage dividers between each stage to set the attenuation level for clipping. Part of the patent involves using the diodes to create the DC bias shift for the n-channel FET device and then cascading a set of these structures as with the triodes they emulate. The fact that cascading modules are used, each generating its own harmonics due to varying DC bias shifts, makes this an attractive solution in both analog and digital forms.

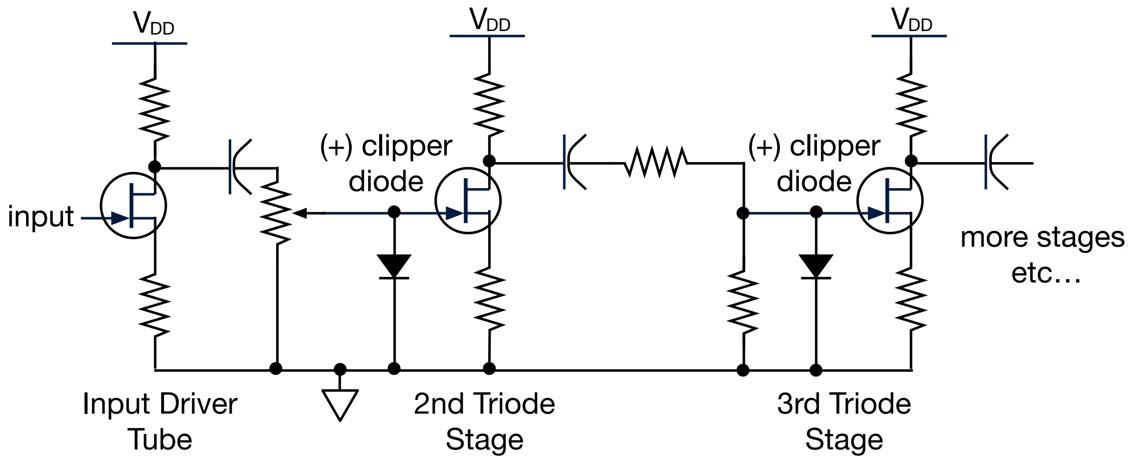


Figure A19.43: Peavey TransTube solution: use n-channel FETs to emulate triodes with negative bias and use diode clippers to clip positive portions of waveforms thus over-biasing the input signal

A more important part of the patent involves the FET audio signal output amplitude as compared to the input diode's clipping voltage, which the patent lists as about 0.5V. The patent states:

"It should be noted that this invention is not simply a diode clipping means, and an operating point shifter. Several U.S. patents discuss operating point shift by means of a diode, with resulting second harmonic distortion. The invention shows the important discovery *that a diode with a mere 0.5V forward voltage clipping value, when driven from a typical solid state device with an output capability of about 20 to 30 volts will closely match the particular output/input ratio of the existing tube circuits and therefore closely emulate the tube sound*. The present invention also teaches that multiple operating point shifts produced by multiple stages generate a multiplicity of levels and amounts of second harmonic distortion over a wide range input signal levels."

This underscores the importance of having a large amplifier gain that overdrives each successive stage's input significantly. By carefully choosing the inter-stage attenuation level and cascading multiple stages together, the input signal will go through multiple DC bias shifts of differing intensities, generating a rich and complex set of harmonics. This is part of the high-gain tube preamp sound and cascading stages together is part of a good emulation. As an analog system, the Peavey circuit's clipper automatically shifts the bias point as the input is being clipped. It should also be noted that the Peavey patent also includes a Darlington BJT embodiment as well as the FET version.

A19.18.2 Yamaha Class-A Emulation

For a digital implementation that uses cascaded stages, we will need to mimic the analog version. The Yamaha patent provides three different approaches for emulating the overdriven cascaded triodes. All of them involve the same set of steps:

- Clip the input signal, or pretend to clip it
- Detect the DC offset in the signal
- Use the detected offset to shift the operating point of a lookup table waveshaper either directly or indirectly

The Yamaha approaches involve multiple stored tables. Some of them are for waveshapers. One of them stores an input-current (I_{IN}) to grid-current (I_G) calculation. In the first approach, the triode's input grid resistance R_G and capacitance C are used along with a reverse Euler integrator to find the DC offset shift required for a given positively clipped waveform. The Euler integration is:

$$V_c = \frac{\left\{I_G(n-1) + \left[V_{in}(n-1) - V_c(n-1)/R_G\right]\right\}}{Cf_s} \qquad \text{[A19.6]}$$

The result of the integration is the DC offset shift that the waveshaper (lookup table) is biased with. This is shown in Figure A19.44 (a). In Figure A19.44 (b) and (c), the same idea applies, however the DC level is subtracted directly from the input prior to waveshaping. In one version, a low-pass filter and coefficient multiplier extract the DC offset and subtract it from the input. In the other, an envelope detector is used. The amplitude of the input signal is compared against a table of amplitudes versus DC level shifts that *would have result*ed when the input was clipped. Then, this DC value is subtracted from the input, providing the over-biasing signal.

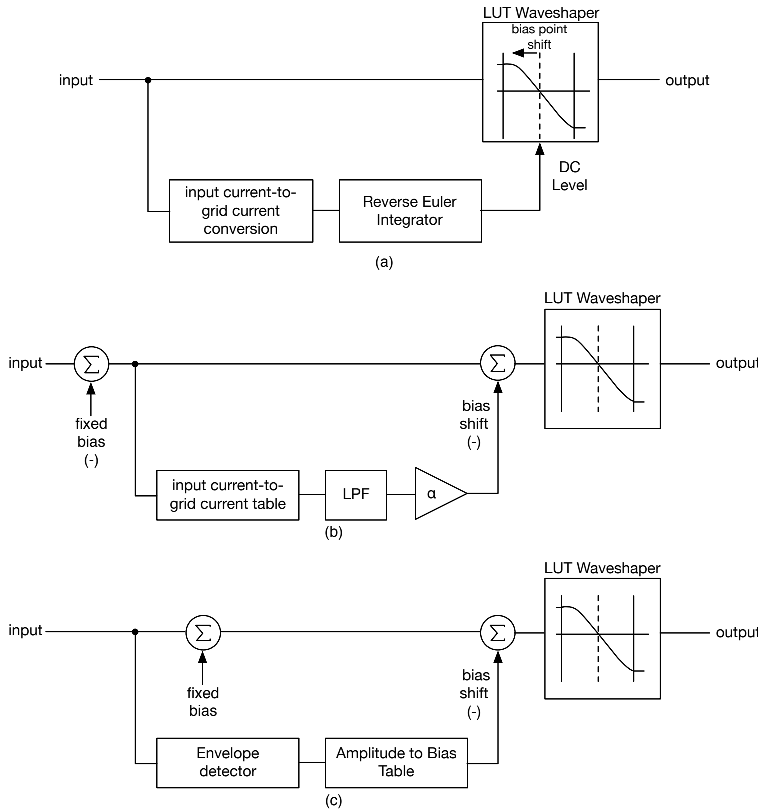


Figure A19.44: three Yamaha approaches to the overloaded triode problem using a lookup table (LUT) based waveshaper where (a) uses a reverse Euler integrator to extract a DC component from a table while (b) uses a low-pass filtered grid-current LUT as the DC shifter directly and (c) an envelope detector finds the signal amplitude and a table of amplitude-to-bias-shifts is applied directly

Neither the Peavey nor the Yamaha solutions attempt to model the 3/2 power law directly, though we can assume the Yamaha lookup waveshaper tables somehow reflect it in their values. The Peavey analog version uses the inherent DC level shifting that the clipping causes, whereas the Yamaha version must subtract the DC offset either directly or via modifying the waveshaper.

A19.18.3 Pirkle: Improved Class-A Triode Object

In Figure A19.45 you can see my solution, which is based very closely on the Yamaha block diagram but with some modifications. In effect, I am digitizing the Peavey TransTube emulation. The first observation is that when the overloaded grid conduction occurs, the resulting diode does pull electrons from the cathode, but as they approach the grid, the much higher plate voltage sucks some of them right up and conducts. The result of this process is that the input at the grid is not actually clipped per se, but is more *compressed* downward instead. The resulting input may well clip the positive side of the waveshaper and result in squaring of the signal, but this happens in the tube itself, not at the grid. The Yamaha versions use an input current to grid current conversion table in two of their approaches, whereas I am using empirical data.

I used measurements taken from my SPICE simulations to come up with a compression value based on how far the input signal level lies above the clipping point, from 1 to 100 volts. For the 12AX7 circuit we've been working with, that happens when $V_{in} > 1.5V$ which is ½ of our $3V_{p\text{-}p}$ range. This produces a special kind of waveshaper that only operates on the positive input values and is linear until the clip point, then produces soft compression above that. The equation for A, the attenuation coefficient is:

$$A = \begin{cases} 1.0 & V_{in} \le V_{clip} \\ 0.447 + 0.545 e^{-0.3241584(V_{in} - V_{clip})} & V_{in} > V_{clip} \end{cases}$$

After the input has been processed, a lossy integrator measures the DC offset now present in the signal. This DC offset is scaled with a coefficient (that may be 1.0) and then the new input is level shifted prior to wave shaping via a DC offset subtraction.

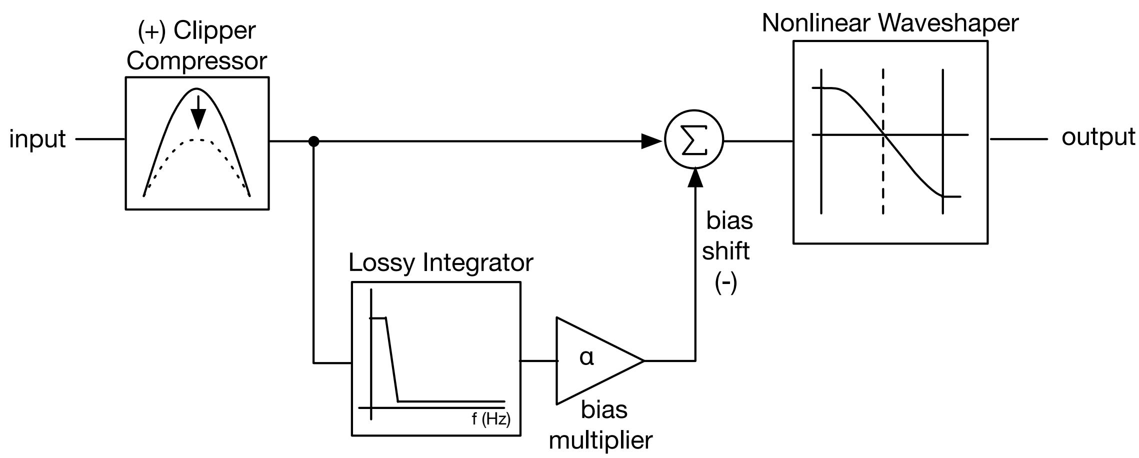


Figure A19.45: my overdriven triode emulation includes an input compressor/clipper and a lossy integrator to measure the DC offset of the processed input; this DC offset is used as the level shifting mechanism prior to wave shaping

In my adaptation, each triode is a closed system that implements the 12AX7 circuit we designed at the start of the addendum. In this system, the proper input signal size is $3V_{p\text{-}p}$ and we assume cathode self-biasing, so the input range will be -1.5V to +1.5V. The clip-point is taken to be 1.5(V). In this version, we use the fact that our floating point implementation gives us plenty of headroom, so we allow the signal to become much larger than the normal range of [-1,+1] with +1.5 as the positive clip threshold.

Each *ClassAValve* object contains a HPF for emulating the output coupling capacitor's effect. This filter will remove the DC bias from the output signal, which is caused when we apply the over-bias value. In addition, a lowpass filter is added to allow you to control the upper bandwidth of the device. A parasitic capacitance exists between the plate and grid that causes a negative feedback path, but only for very high frequencies. This capacitor is sometimes called the Miller Cap and it is named the same way in the object parameters.

We will discuss this object in much greater detail Part V where we implement our algorithms in C++ objects. We will look at the time and frequency spectrum plots in great detail then as well. Please also note that I am calling this model the "Pirkle" version for this document only. There very well may be a similar implementation that I don't know about, and my version is already very closely aligned with both the Yamaha and Peavey TransTube patents. Please do not fill my email inbox with comments about my naming convention here.

A19.19 Pentode Class-B Model

The root cause of the tube compression in Class-B power amps is the grid conduction of positive V_{GK} voltages as with the triode. However, the manifestation of the crossover distortion and softly squared clipping occurs differently because there are two tubes working together, each being over-biased during conduction of the upper and lower waveform halves. The over-biasing causes the crossover distortion, and cut-off clipping causes the limiting of the output amplitude. This amplitude limiting, combined with the compression of the waveform such that the center is smashed down into the crossover distortion region, creates the tube compression effect. Remember that push-pull Class-A amplifiers will eventually suffer from the same compression, but only after they have been so over-biased that they essentially turn into Class-B amplifiers.

A19.19.1 Peavey Class-B TransTube

The TransTube solution once again exactly emulates what is happening in the pair of tubes. As you can see in Figure A19.46 there are two signal paths that mimic the two tubes in the Class-B arrangement. If you look at the upper branch, you can see the use of a positive clipper diode as the grid-conduction emulation, and an ingenious over-biased diode as the negative clipper, which has the same effect as shifting the DC operating point of the Class-B amplifier. The lower branch performs the same function on the opposite waveform half by simply reversing the direction of all the diodes and the bias voltages. The outputs of the two branches are summed. As the input signal grows, driving the Class-B emulator further into the over-biased condition, the output waveforms add together with increasing crossover distortion and center waveform compression while the tops are softly clipped and squared off. The output waveform of the SPICE simulation for the circuit is also shown – notice the massive DC offsets of the top and bottom waveforms, which are the pair of output signals. The combined response demonstrates the tube-compression effect.

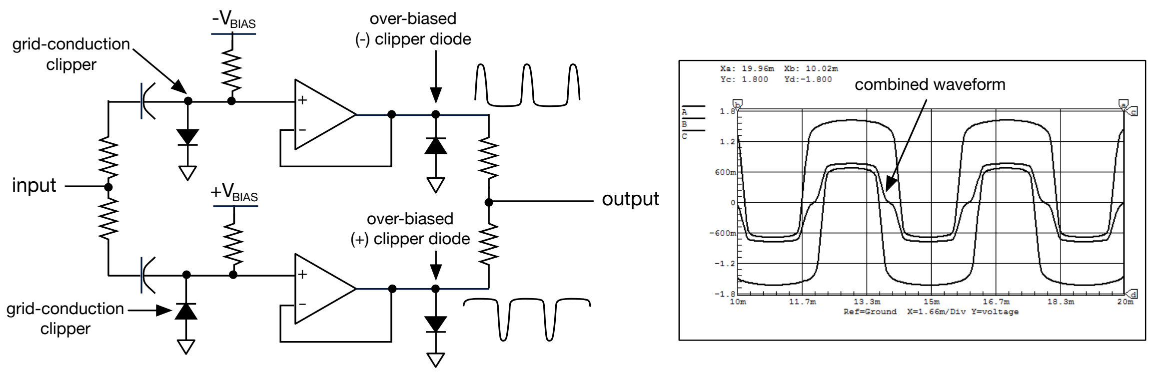


Figure A19.46: Peavey TransTube Class-B emulator uses over-biased clipping diodes to generate its two outputs; the two bias voltages hold the output clipper diodes in a just-biased-on state and when the input signal level swings above the grid conduction clipper voltages, they provide the over-biased signal; when summed, the result is an excellent match for Class-B tube compression as seen in the SPICE simulation; as the input level increases, more and more of the signal is compressed into the crossover distortion region

A19.19.2 Yamaha Class-B Emulation

If you understand how the TransTube version operates, then the Yamaha version will appear as a kind of digital translation using look-up tables for the wave shaper, and bias shifting via a DC offset to create the output signals. Yamaha offers three different embodiments for the overloaded triode in their patent and any of them may be used. Here, there are also two branches that operate on the upper and lower portions of the input waveform. The lookup table waveshaper is set to have a comparator-like amplified edge, which is used to shape the signal. Yamaha states that the table is populated with data that makes the output waveform have the proper shape, but as with the other parts of the patent, they do not elaborate or give many equations. When summed together, the two branches produce the same effect as TransTube. Inversion in one branch mimics the phase inversion from the Class-B topology and the LUT waveshaping and DC shifting processes are identical but opposite.

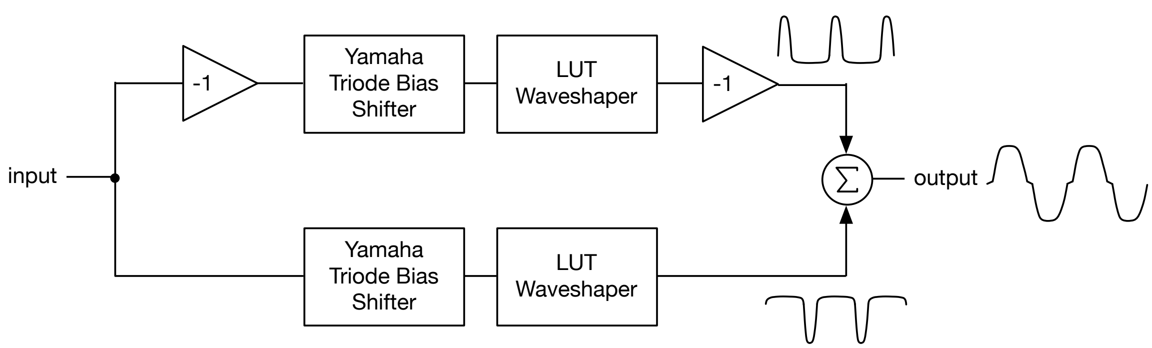


Figure A19.47: the fundamental idea behind the Yamaha digital Class-B tube emulator is remarkably similar to the Peavey TransTube solution but in digital form

A19.19.3 Poletti Class-B Emulation

The Poletti implementation of the overdriven Class-B amplifier is interesting and has many options for modifying the distortion properties. There are far more controls with this algorithm than the others and it can also generate distortions and waveforms that are quite bizarre and exotic, from an "asymmetrical Class-B" to complete and utter destruction of the waveform. The Poletti method relies on an equation for a waveshaper from the book *Simulation of Communication Systems* by Jeruchim, Balaban and Shanmugan. The waveshaper appears to be part of a family of memory-less limiter functions that are designed as soft-clippers, and it also works well as a standalone waveshaper for Class-B emulation without overload. The function is:

$$y(n) = \begin{cases} \dfrac{kx(n)}{1 - \dfrac{kx(n)}{Ln}} & x(n) \le 0 \\ \\ \dfrac{kx(n)}{1 + \dfrac{kx(n)}{Lp}} & x(n) > 0 \end{cases} \qquad [19.7]$$

There are three (3) parameters for this asymmetrical waveshaper. The gain value *k* is the same scalar you saw in the FX book that we called "saturation" while there are two limit values, L_n for the negative bound and L_p for the positive bound. Symmetrical waveshaping occurs when you let $L_n = L_p$. Figure A19.48 shows the outputs of these waveshapers for two sets of coefficients, one for shaping the positive half and the other for the negative portion using values prescribed in the patent application. Notice how the waveshaping is highly asymmetrical with these values, which will clip the positive and negative portions of the waveform down to +/- 0.6. Notice what happens when the DC offset is removed in Figure A19.48 (c) and (f).

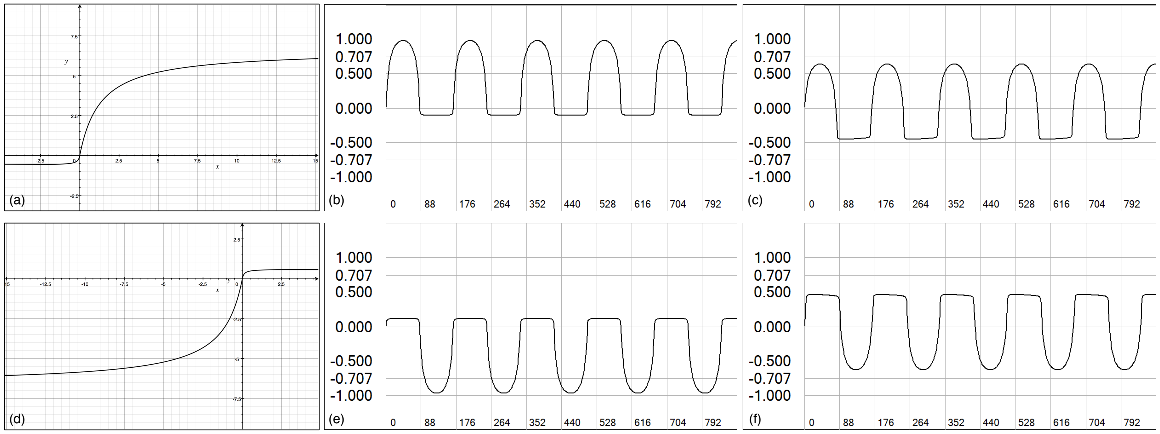


Figure A19.48: Poletti waveshapers with $k = 40$ and output scaling for (a) the positive input waveform ($L_p = 6.6$, $L_n = 0.6$) (b) shows the output of while (c) shows the output after high-pass filtering to remove the DC component and (d) waveshaper for negative input waveform ($L_p = 0.6$, $L_n = 6.6$) (e) shows the output of while (f) shows the output after high-pass filtering to remove the DC component

When the DC offset is removed and the signals recombined, they form the soft, squared off waveform of the Class-B amp prior to overdriving the grid input. To add the crossover distortion and tube compression, yet another waveshaper is applied, this one a symmetrical version of the same waveshaper function. When the gain (k) values between the two sets of waveshapers (asymmetrical and symmetrical) are mismatched, the crossover distortion appears. The values in the patent application were either grossly incorrect (ignore the negative L_x values) or did not produce the compression effect smoothly as the input amplitude increases. After some experimenting, I found values that worked well. The beauty of this algorithm is actually in all the *other* stuff you can do to manipulate the output, so it was worthwhile to spend some time fine-tuning it. Figure A19.49 shows the Poletti Class-B emulation algorithm. Note that all waveshapers share the same equations but use different coefficients; the symmetrical version has $L_n = L_p$.

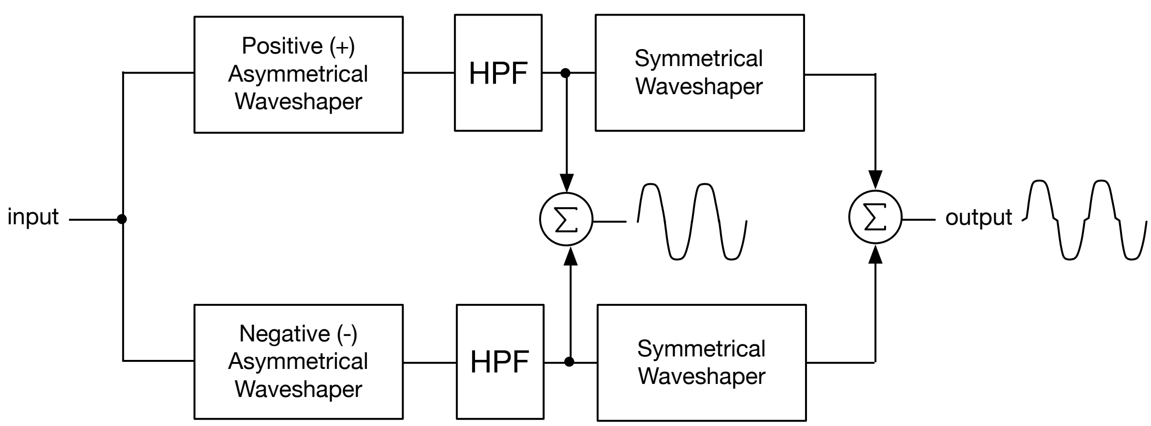


Figure A19.49: the Poletti Class-B Emulator uses another pair of signal paths to affect the positive and negative waveform halves independently; at the point just after the HPF, the summed waveform resembles the output of a push-pull Class-A amplifier, so you may utilize this output as well

Figure A19.50 shows the output of the Poletti algorithm with a variety of input gain settings – remember that the input gain drives this signal processing. Here, the input gain is allowed to go has high as 100 (+40dB) and multiple level adjustments are made to keep the output in a reasonable range for our plugins. Full details of the algorithm and settings are included in Part V when we implement these as C ++ objects. Compare these plots with Figure A19.41's tube simulation with 6L6-GCs (note that the input frequency is different so the spectral components are spread out differently, but the overall spectral envelopes match fairly well). Also note that you can obtain slight to drastically different results by altering some of the many parameters for this algorithm.

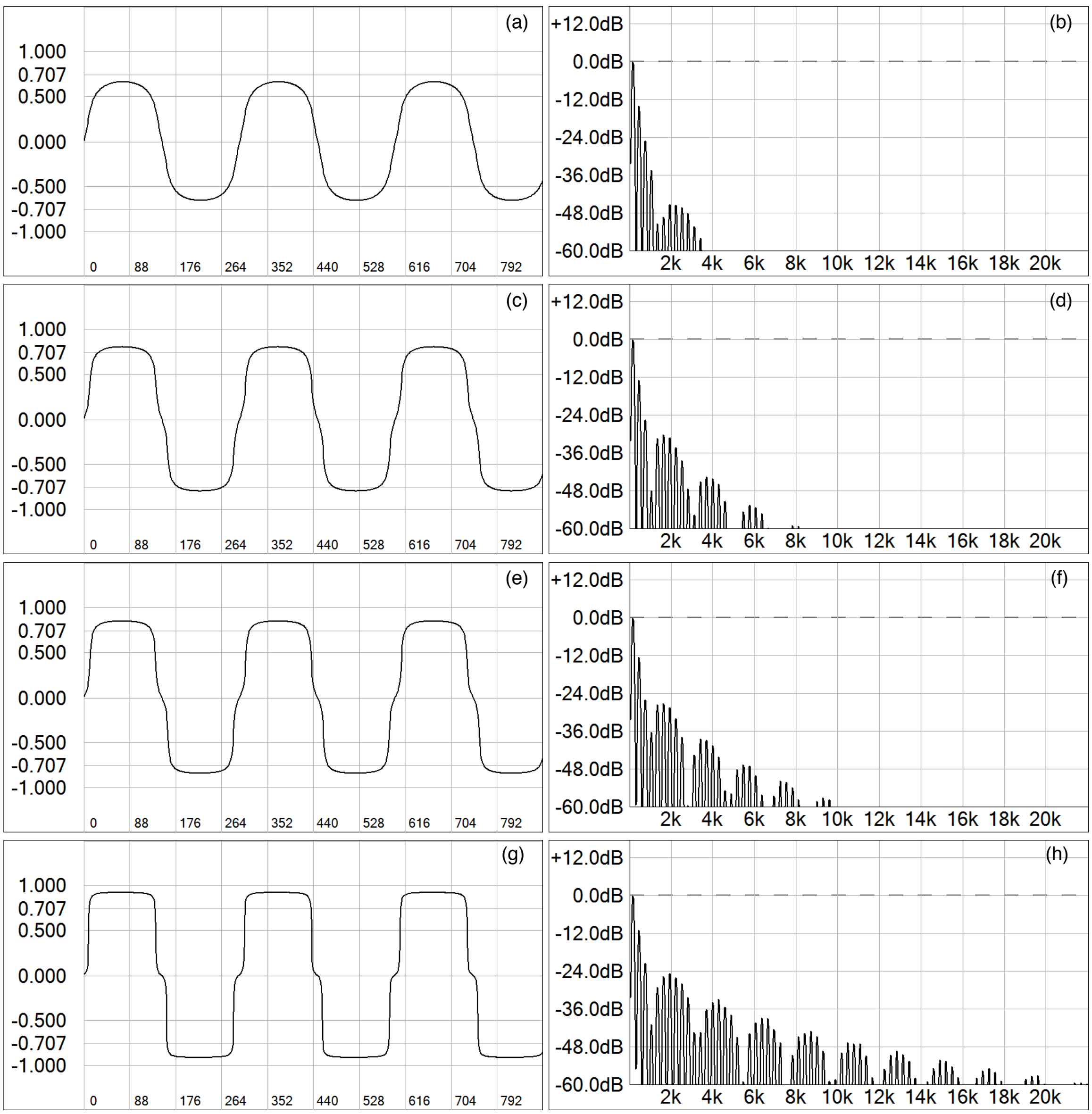


Figure A19.50: time and spectral plots for the Poletti Class-B emulator at 150Hz and varying values of input gain G for unity (a) and (b), $G = 3$ (c) and (d), $G = 5$ (e) and (f), and lastly $G = 25$ (g) and (h); notice how the spectrum changes significantly as the crossover distortion and waveform squaring increase

A19.19.4 Pirkle: Class-B Pentode Object

The method used to generate the improved triode Class-A object can be extended to the pentode (or triode) Class-B case as well. The fundamental differences are the waveshaper equation chosen and the fixed bias point at cutoff. As with all the rest of the models, it uses two branches to process the input signal with inverters in one branch so that all processing is applied to the opposite waveform half. Other than the doubled processing, the rest of the operation is fundamentally identical to the triode Class-A version.

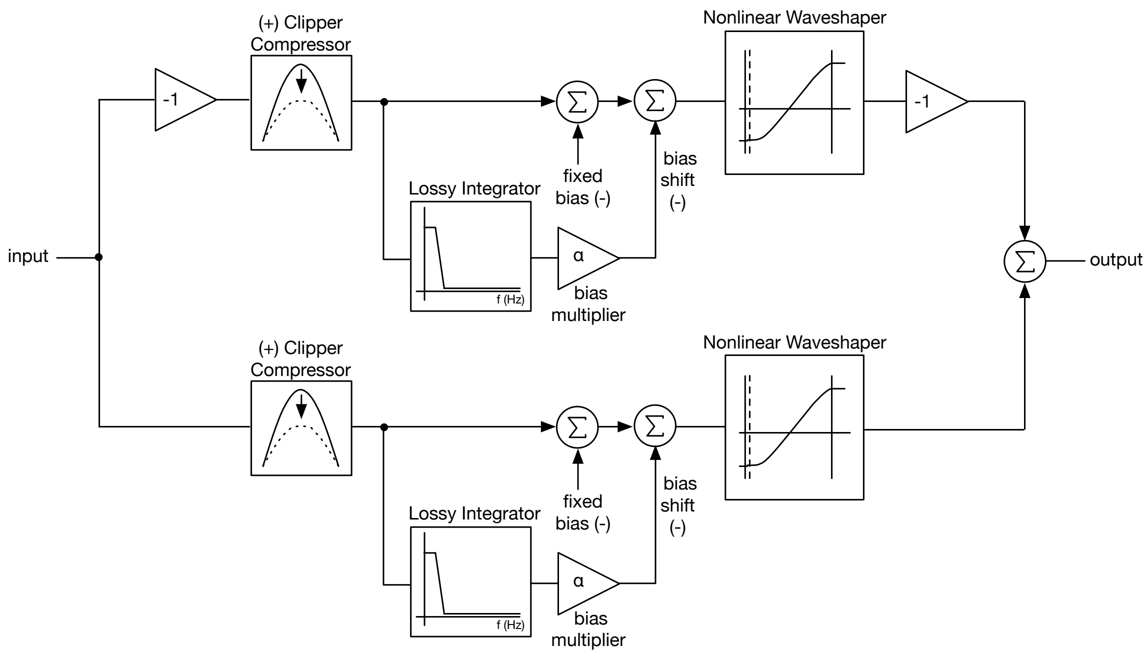


Figure A19.51: Pirkle's overdriven Class-B emulator uses twin branches operating on normal and inverted versions of the input signal uses a fixed bias at cutoff plus a bias offset generated when grid conduction clipping occurs

Figure A19.52 shows the output of this Class-B emulator at a variety of input gain settings. The details of the system will be revealed later when we look at the C++ implementations. You can see that increasing the input gain causes the expected crossover distortion and soft squaring of the waveform and tube compression occurs once the gain is increased above the crossover distortion point. The plots here are only one set of numerous variations. Minor parameter changes can make a big difference in the final output and there are multiple parameters that may be varied so you have plenty of room for your own experimentation.

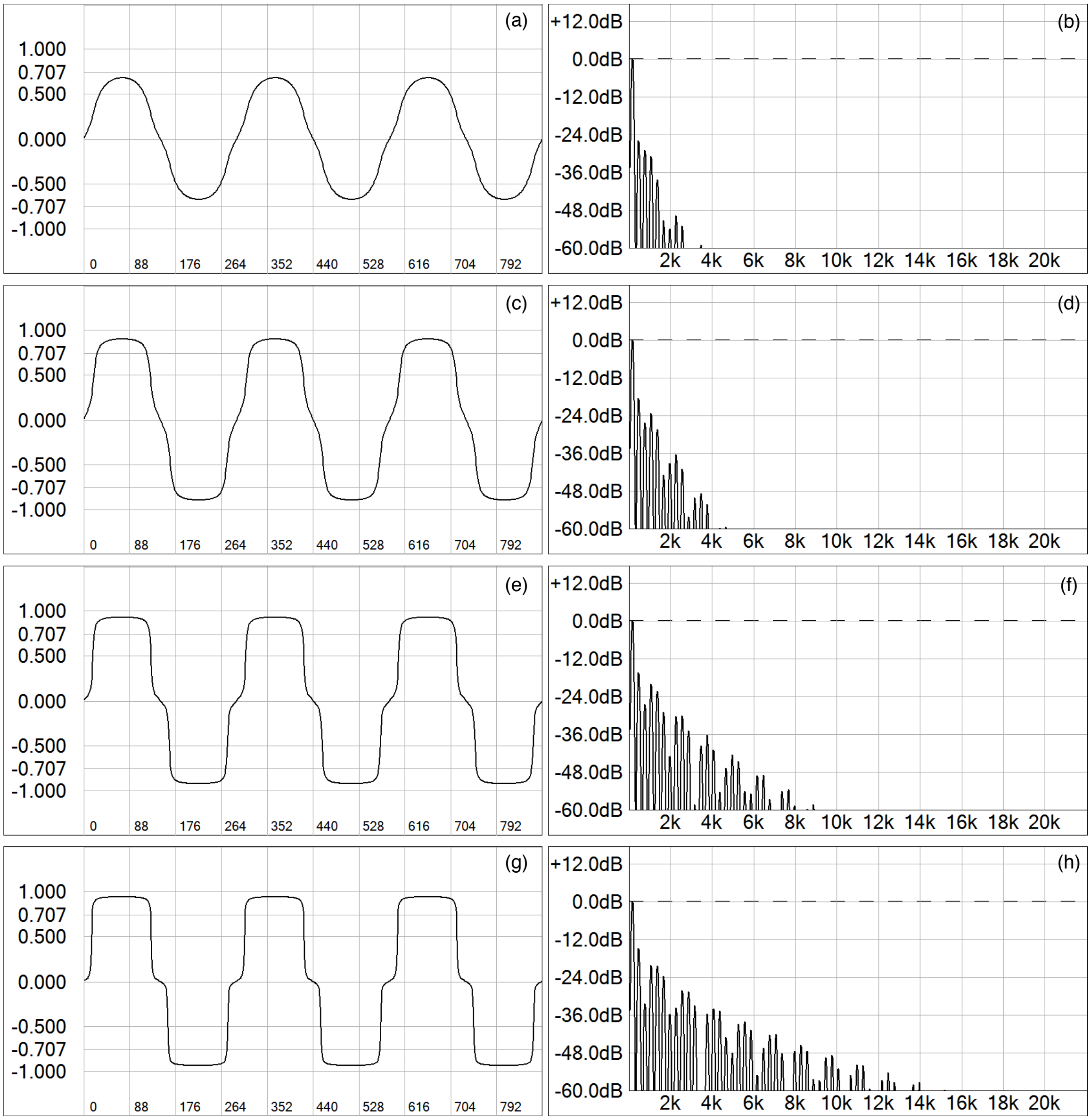


Figure A19.52: time and spectral plots for Pirkle's Class-B emulator at 150Hz and varying values of input gain $G = 3$ (a) and (b), $G = 20$ (c) and (d), $G = 40$ (e) and (f), and lastly $G = 75$ (g) and (h); notice how the spectrum changes significantly as the crossover distortion and waveform squaring increase – compare with the other Class-B plots

Part IV: Filtering and Distortion

Before looking at the C++ implementations in the next section, it is worth taking some time to discuss filtering and distortion algorithms as combined processes that we can also separate out to make the model simpler to generate. It turns out that careful filtering both before and after applying distortion is crucial in obtaining a good distorted guitar sound in both the analog or digital domains, whether you are trying to emulate tubes or not. For example the Sans Amp GT-2® achieves its distortion by simply overdriving an op-amp. Its design is a testament to what good filtering before and after the distortion amplifier can achieve. [Kyttälä 2008]

A19.20 Circuit/Patent Study

As before, we will analyze some circuits and patents.

A19.20.1 Peavey

Peavey patented several inventions before TransTube and one of them called *Superdistortion* is included in the accompanying set of patents and bonus documents. The Superdistortion circuit tries to emulate a heavily overdriven tube amplifier using a diode based clipper circuit. The beauty and genius of this circuit is actually the filter that is part of it, whose characteristics change as the overdrive amount changes. This was a continuation of a previous patent idea involving this dynamic filtering idea.

Pat: 4,811,401 *Superdistorted Amplifier with Normal Gain* [Peavey, 401]

A19.20.2 Scholz R&D

Tom Scholz has numerous patents under his belt going all the way back to his Polaroid days. The Rockman® and the resulting series of products that it spurned were used on countless recordings from the 1980's. It only had four preset distortion sounds (and two of them were 'clean' settings), but you could layer track after track of distorted guitar without it becoming muddy sounding. Sholz's patents should be standard reading material for anyone interested in generating distorted guitar sounds because they include a plethora of background information on guitar distortion and harmonics processing theory. The patents included here are:

Pat: 4,584,700 *Electronic Audio Signal Processor* [Scholz, 700]
Pat: 5,133,015 *Method and Apparatus for Processing an Audio Signal* [Scholz, 015]

A19.20.3 Gallien-Krueger

Gallien-Krueger (GK) produced solid-state guitar amplifiers up until the mid-1990s – in fact the company's very first product was a solid-state guitar amp. The GK 250ML represents an excellent achievement – a two-channel, stereo, 100W guitar amp with chorus and reverb in a package not much larger than a child's lunchbox. The method for generating distortion using matched FET current regulators was (and still is) quite novel and produced a fantastic heavy tube emulation. It also lent itself to cascading series modules to achieve varying degrees of distortion. The power amp was also designed as a soft-clipping amp but it did not emulate overdriven Class-B tubes directly.

A19.20.4 Scholz Rockman Distortion

The Rockman had several variants and the one discussed here is the X100B model. As with the other Rockman products, it had four presets: two of them being distorted and two being clean. In each of the four presets, a different combination of compression, pre-filtering, distortion, and post-filtering are applied. There is only one distortion amplifier circuit that uses a high-gain op-amp to overdrive back-to-back diodes, in the form of red LEDs, which have a V_D of about double that of a silicon diode. The two distortion settings mainly differ in the amount of gain the signal receives prior to the filtering and distortion, which are the same for both settings. Figure A19.52 shows the signal path for the distortion presets. The pre-distortion filter is a bandpass filter (BPF) and the post-distortion filter, named the "complex filter" in the patent, is a quasi-third order, fairly steep lowpass filter with multiple ripples in the pass band and an amplified bass response.

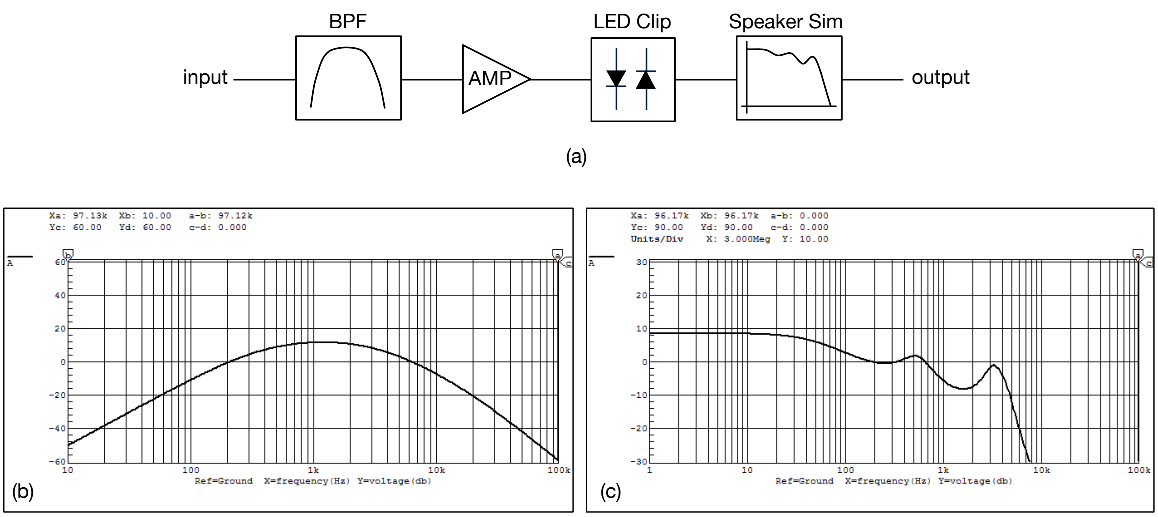


Figure A19.52: (a) the Rockman distortion preset signal chain (b) the input BPF to the distortion amplifier and (c) the speaker simulator filter response

The input bandpass filter has the following specifications:

- f_c: 1.225 kHz (f_L: 500 Hz f_H: 3.0 kHz)
- *Q*: 0.5
- *Midband Gain*: +12dB
- *Rolloff*: 12dB/octave

The output "complex filter" acts as the speaker simulator (with SM57 microphone). It is described in detail in the patent, with carefully chosen peaks and valleys in the passband area. In addition, the bass response is boosted by about +9dB creating a low boost-shelving filter from 250Hz downward. The analog circuit consists of a lowpass filter followed by two resonant peaking filters that produce the peaks and valleys. The overall gain change from low end to about 5kHz is 18dB. The filter is difficult to specify (see the patent) because it is an amalgam of several filters in series. Figure A19.53 shows two more views of the output filter response. One feature common to most speaker simulators is a high order lowpass filter with a cutoff around 3kHz, and this is exactly what we see here. The Rockman also includes a second post-distortion filter that is a simple 2nd order lowpass type, used in the clean preset signal chain. You should be aware that in the grand scheme of things, the ripples you see in Figure A19.53 are not as pronounced when you zoom out. If you make these peaks and notches too deep, it will begin to sound artificial and over-filtered.

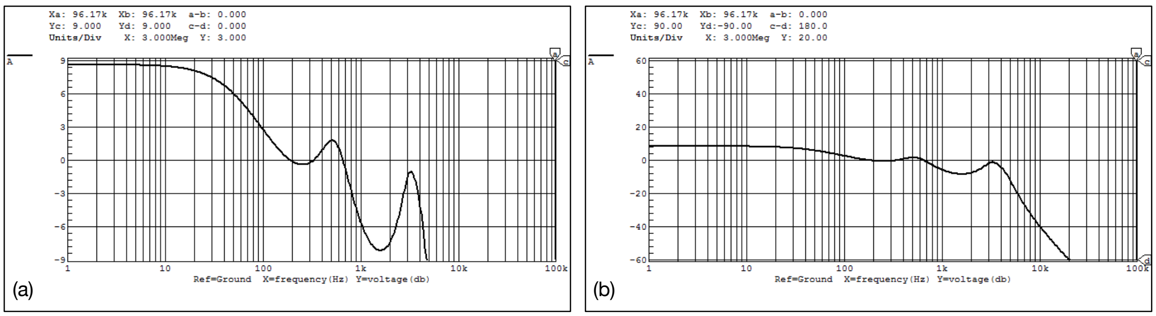


Figure A19.53: two more views of the post-distortion filter (a) detail of passband ripple from -9dB to +9dB and (b) zoom out showing a roll off of about 28dB/octave close to the cutoff edge, tapering down to about 20dB above 10kHz

A19.20.5 Scholz Distortion Generator

Scholz developed more finely detailed versions of the basic Rockman recipe, adding more presets for various degrees of distortion. Figure A19.54 shows the result in another Scholz patent six years following the Rockman patent. This version includes five types of distortion from clean to heavy. Notice that each distortion “channel” has an associated pre-determined pair of filters named *pre-distortion* and *post-distortion*. In addition there are separate pre and post gain stages that surround the single distortion amplifier circuit (though the patent does stipulate that there may be other embodiments that include multiple amplifiers).

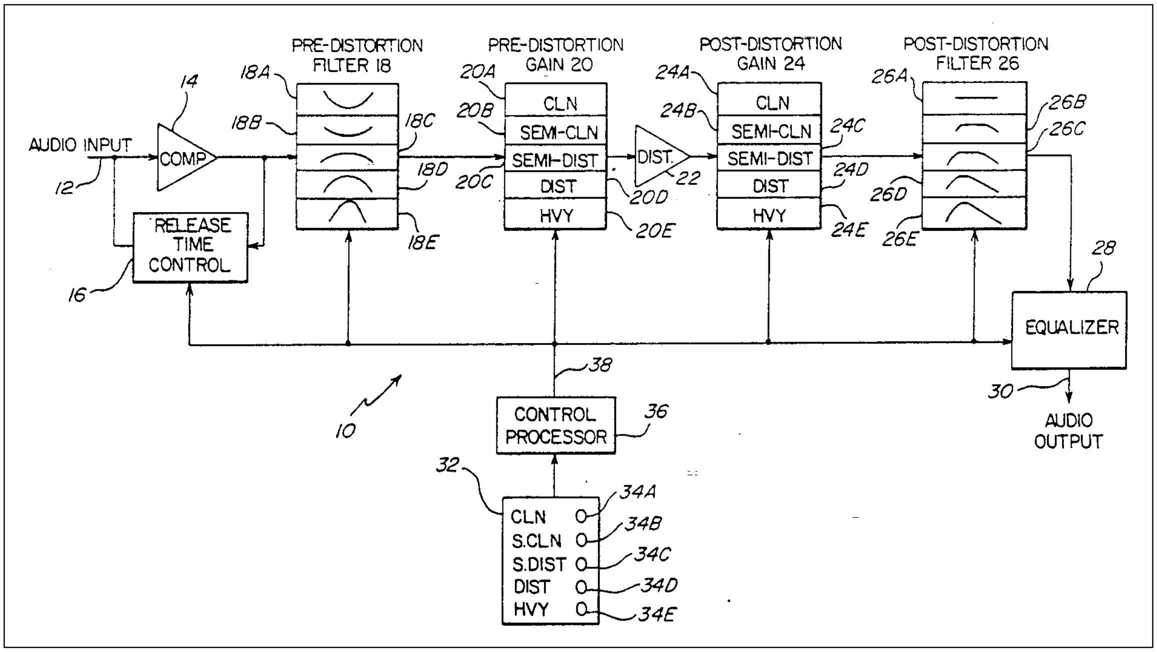


Figure A19.54: from U.S. Patent 5,133,015 *Method and Apparatus for Processing an Audio Signal*

Notice that as the distortion gets heavier going down the channels, the pre-distortion filter's band-pass response gets narrower and the post-distortion filter adds successively more bass while reducing successively more high frequencies. The complimentary filtering here is key and it is clear from the patent that much time was spent setting up these filters. Scholz is very outspoken about the need for the filters in both patents. Here are some selected quotes regarding filtering and distortion from the two sets of patents/circuits. Also, you should read the information about compression before distortion, and the interesting way he biases the compressor with added high frequency content for the clean settings.

Rockman Band-Pass Pre-Filtering:
"The mid band pass filter 14 reduces the high and low signal content before the signal goes through the distortion amp 16. Rolling off the highs results in less noise at the output of the distortion amp and reduces the amount of highs from the input signal heard after the distortion amp 16. *This is important because in this substantial distortion mode it is important that the high end content of the output signal be made up primarily of high harmonics produced by distorting the mid range portion of the signal which are of long duration, rather than by the natural high harmonics contained in the input signal which are of short duration.*"

Rockman Complex Post-Filtering:
"The complex filter 17 which receives the output of the distortion device, processes this output into an output signal having excellent tonal qualities. *Without this filter, the output would be both "harsh' and "muddy' in tonal quality.*"

Distortion Generator Pre-Filtering:
"From FIG. 1, it is seen that *as the level of distortion increases, there is less and less emphasis at the treble and bass ends of the spectrum, with maximum roll off at the treble and bass ends occurring in filter 18E for heavy distortion*."

Distortion Generator Post-Filtering:
Exemplary filter characteristics are shown for each of the filter segments, with the filter characteristic for "clean" filter 26A being substantially flat, and with the base and treble roll-offs on the filters becoming increasingly great, particularly the treble roll off, as the degree of distortion increases. *The filter is thus designed to compensate for the increases in bass and treble harmonic content in the audio signal caused by distortion circuit 22… for typical distortion circuits 22 the bass and treble characteristics of the filter will exhibit increasing roll-off as the distortion level increases*."

A19.20.6 Peavey Superdistortion

The Peavey patent for what they call "superdistortion" is an interesting and clever adaptation of a standard op-amp that drives back-to-back clipping diodes, not unlike the Rockman circuit. It is interesting with respect to the Scholz patent because of a similarity in discussion of the input bandpass filtering operation. The Peavey patent pre-dates the Scholz patent by three years and in the Scholz patent, he comments that the distortion could be made variable (rather than switched presets) with the filters also being made variable as well (presumably with the same knob control). This is what the Peavey super-distortion patent informed and taught three years earlier – it varied a bandpass filter smoothly as the user increased the amount of distortion.

In the Scholz Rockman circuit, the bandpass filter is a passive RC:CR filter that is on the input of the distortion amplifier. In the Peavey super-distortion circuit, the bandpass filter is built into the feedback path, around the single "distortion" potentiometer and incorporating it. In this way, as the user increases the distortion by increasing the gain of the amplifier, they are also changing the bandpass filter, making it narrower and lower in center frequency while achieving higher mid-band gain from the amp. The Peavey circuit includes a fixed pre-distortion filter that is band-pass in nature and the block diagram is shown in Figure A19.55.

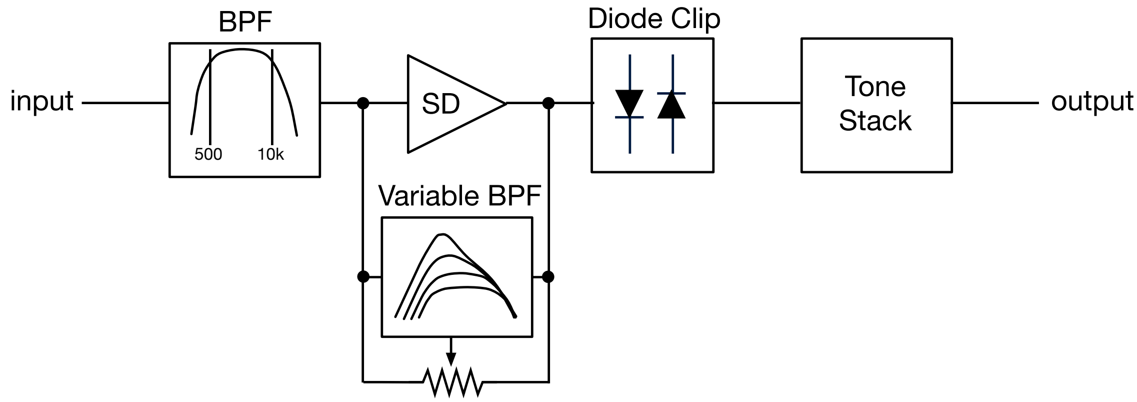


Figure A19.55: block diagram of Peavey superdistortion patent; note the variable BPF inside of the feedback loop of the amplifier

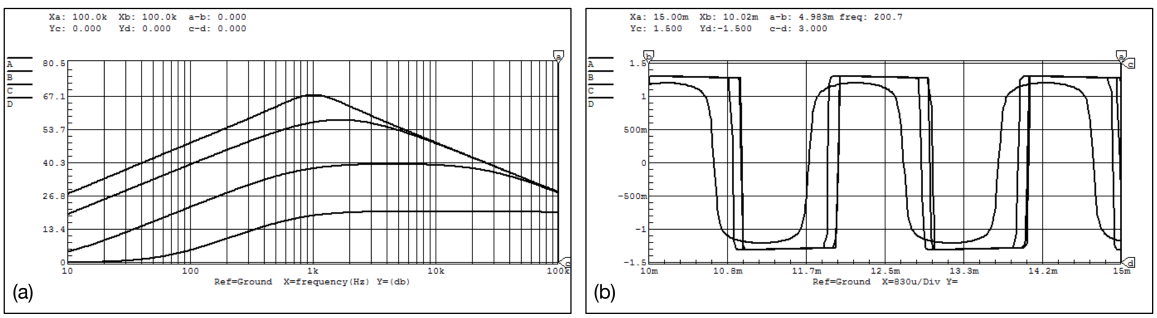


Figure A19.56: (a) the Peavey superdistortion bandpass responses as the overdrive level is increased – notice the bandpass filter becomes narrower, the center frequency moves lower, and the gain increases with increased distortion and (b) the four, time domain outputs at these settings

Figure A19.56 shows how the bandpass filter changes as the distortion level increases. In these simulations, the distortion control is increased by 10x as the potentiometer moves from 1k to 10k to 100k to 1M. The center frequency of the bandpass filter moves from around 14kHz down to 1kHz over the same range. The gain, which saturates the clipping diodes, ranges from about +20dB to almost +70dB at the top of the range.

There are a couple of other features as well in Figure A19.55. There is a fixed input bandpass that precedes the super-distortion circuit with a low-edge roll-off that starts at 500Hz. This means that the signal is doubly-bandpass filtered and has already lost a significant amount of low frequency energy before having more of it removed with the super-distortion variable bandpass filter. The output of the super-distortion circuit then feeds a 3-band passive tone-stack EQ, which may then further shape the signal and acts as the post-distortion filtering module.

Patent Quote:
"The gain of superdistortion amplifier stage 16 is designed to be high so that the output of operational amplifier A2 in combination with the distortion stage 18a is clipped, thereby generating a number of harmonics of the frequencies of the input signals thereto from the output of tone control filter stage 14. *The presence of such harmonics is believed by the*

inventors to enhance the sound output from the amplifier circuitry in the superdistortion control mode, and especially those harmonics associated with the mid-range frequencies. In particular, guitars are known to produce predominantly low frequency notes, and the harmonics generated by the amplification of the superdistortion amplifier stage 16 and distortion stage 18 (described, infra.) are believed to enhance the sound of the output signal generated by the superdistortion amplifier circuitry"

A19.20.7 Gallien-Krueger GK250 ML

The GK 250 ML "lunchbox" amplifier has two input preamps, one for a cleaner signal and another for a high-distortion channel. Each channel is setup in a similar manner, and they share part of a common signal path. The signal path follows what is now a common theme in this section – input pre-filtering, distortion (a little or a lot) and then post-filtering followed by EQ. The GK amps use active tone controls so there is no passive tone-stack. However the pre-distortion filters are designed to mimic, at least in part, the tone stack with all the controls at about 50% that produces a mid-cut type of response. Figure A19.57 shows the preamp's block diagram.

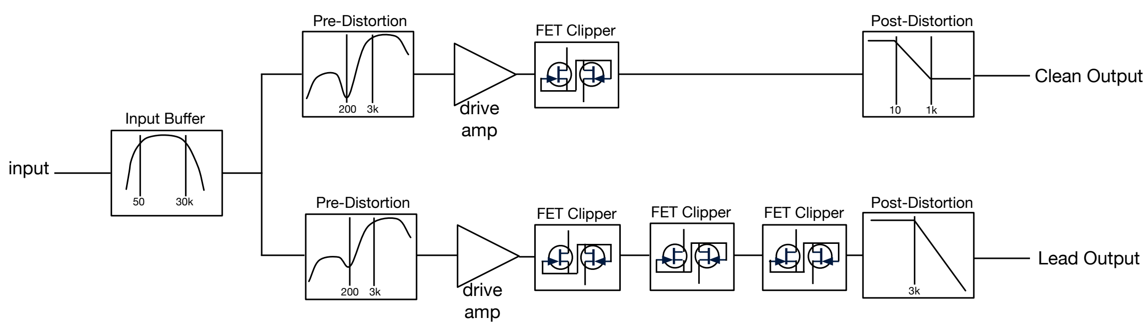


Figure A19.57: the GK 250 ML preamp has two channels and two sets of pre and post-distortion filtering

In Figure A19.58, you can see that the pre-distortion filters are nearly identical, the only main difference being the depth of the notch at 200Hz. The post-distortion filters are quite different, with the clean channel filter adding 30dB of bass boost with a low shelving filter and the lead channel providing a 2nd order lowpass filter at ~3kHz. If you are interested in FET circuits, the GK dual JFET clippers are unique, with their gates tied to their sources and acting as current regulators, in effect clipping the current rather than the voltage of the signal.

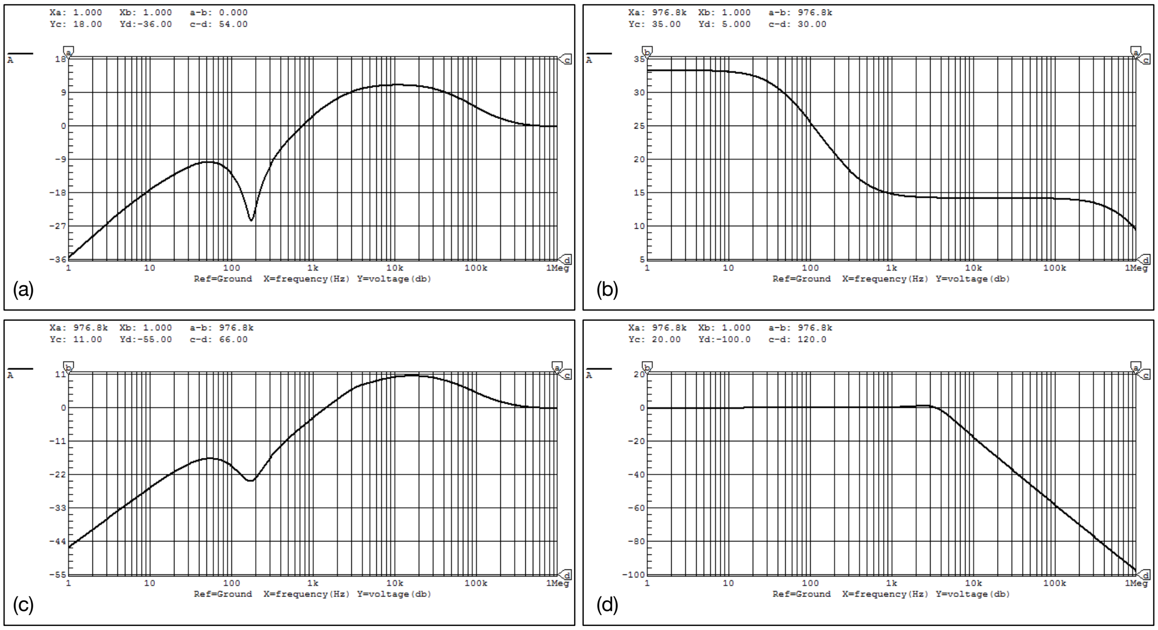


Figure A19.58: SPICE simulations for the pre and post-distortion filters for the GK 250ML (a) clean channel pre-distortion (b) clean channel post-distortion (note massive bass boost) (c) lead channel pre-distortion and (d) lead channel post-distortion

NOTE: in the GK 250-ML case especially, you should remember that these filters were designed to take into account the entire rest of the voicing of the instrument so they might sound strange or un-usable when you first start experimenting with them. Take your time and be patient – the voicing filters are tuned basically by ear so you need to spend time listening and saving presents (early and often).

A19.21 Case Study: Mesa-Boogie DC-2/DC-3 Preamp

We've gone over a lot of information up to this point from basic tube circuit design through the various aspects of overdriving tube stages, then pre and post-distortion filtering. To tie this all together, have a look at the clean and lead channels from my 1996 Mesa-Boogie DC-2 (the preamp is nearly identical to the DC-3). The schematics are available on the Internet so they won't be re-published here in verbatim, however I have a block-diagram convention that can demonstrate the circuit at the top-level design position.

A19.21.1 Mesa-Boogie DC-2 Rhythm Channel

Figure A19.59 shows the block diagram and key functionality of the clean (aka rhythm) channel of the DC-2. It uses three triode stages, all with the same plate voltage and load resistor. Between each stage are a filter and an attenuation mechanism.

A19.21.1.1 Triode 1:

The first triode uses a fully bypassed cathode resistor and drives a tone-stack directly. The mid-band gain of the tone-stack output with all controls set to 50% is about -10dB or 0.32; its output feeds the preamp's gain potentiometer that sets the overall signal level through the rest of the preamp.

A19.21.1.2 Triode 2:

The second triode stage is nearly identical in design to the first one, but has a different cathode bypass cap. It is DC coupled to the next stage via a capacitor that forms a low cutoff of 10Hz. An attenuator follows the blocking cap, attenuating the output by 0.64.

A19.21.1.3 Triode 3:

The third and final triode is identical to the first stage. It is DC coupled to the next stage via a capacitor that forms a low cutoff whose frequency will vary with the potentiometer that follows it which sets the master output level for the clean channel. The circuit produces a maximum gain of 0.035 when the volume control is all the way up.

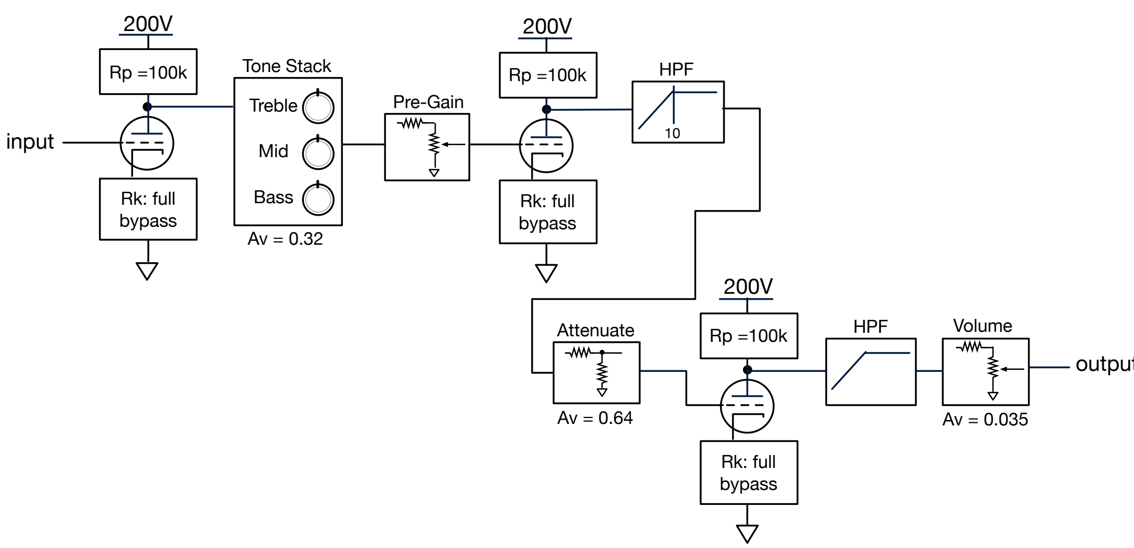


Figure A19.59: the clean channel of a Mesa-Boogie DC-2 in block diagram form

Now have a look at the frequency responses for each stage taken individually in Figure A19.60. You can see the effect of the low frequency shelf (from the bypassed cathode resistor), the inter-stage filtering, and the gain reduction from the attenuator blocks. Notice how the cathode bypass low frequency shelf and the DC blocking cap's responses yield a step (plateau) between 10Hz and 100Hz. Also, be aware that these plots are for each stage taken individually (this is because of how I factor the triode object in the book) so you can see that as the signal goes from stage to stage, it becomes more and more bandpass filtered but in a complex manner involving the low frequency step. Think about the Peavey Superdistortion and the Scholz Distortion Generator's use of pre and post distortion filtering.

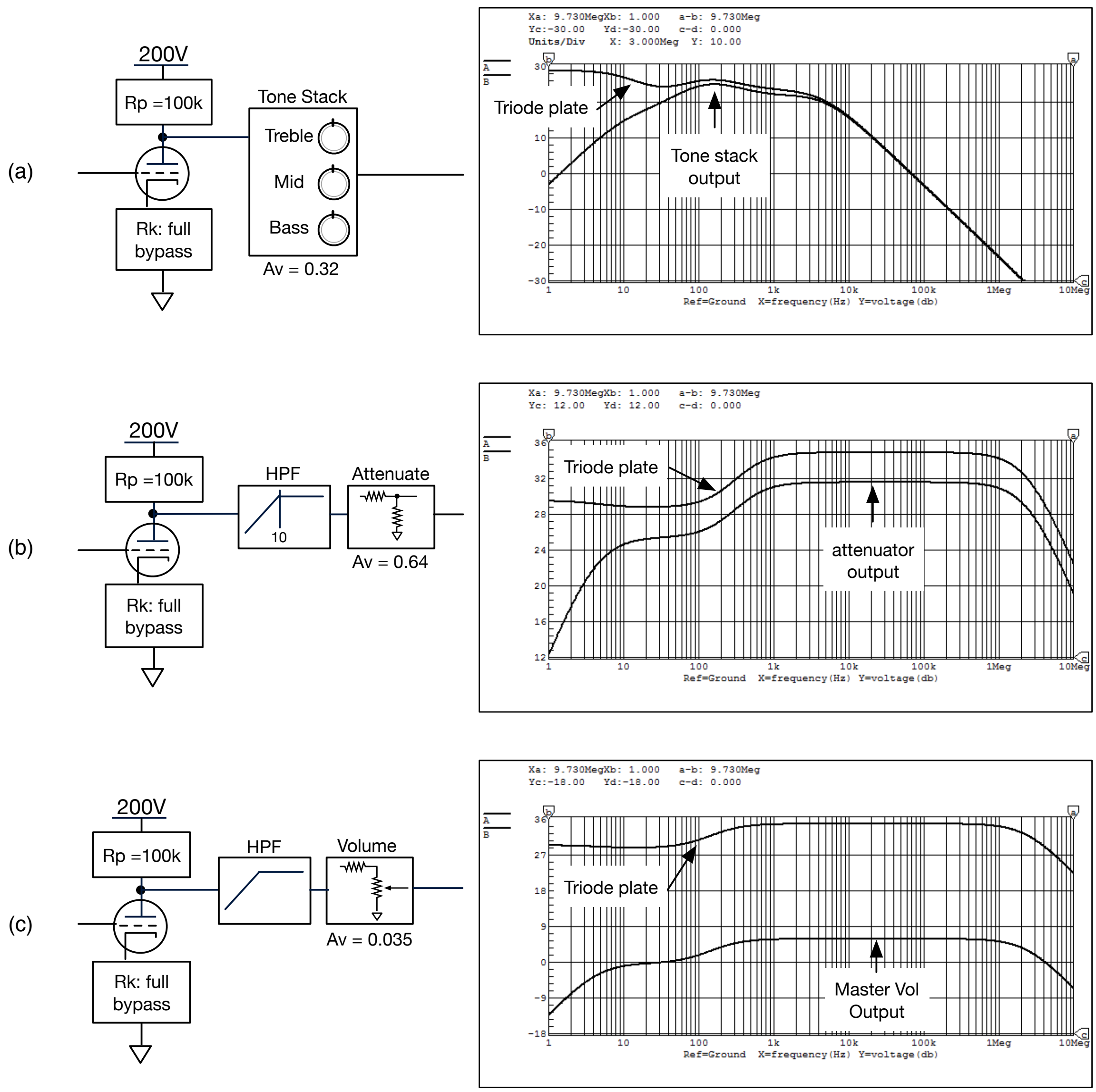


Figure A19.60: analysis of each stage of the DC-2 clean channel (a) input amp and tone stack (the controls are set at 50% for the simulation) (b) the 2nd amplifier triode and (c) the output stage and master volume control

A19.21.2 Mesa-Boogie DC-2/3 Lead Channel

The lead channel is highly overdriven and uses one extra triode stage. It includes an interesting pre-distortion filter along with some more variation in the gain and frequency responses of each block. Figure A19.61 shows the complete lead channel in block diagram form. We'll analyze it just as we did with the clean channel triode modules.

A19.21.2.1 Triode 1:

The first triode is different from the rest in the preamp. It has a lowered plate resistance and a partially bypassed cathode resistor. Its output is DC coupled via a capacitor into a filter/attenuator. The filter is bandpass in nature, and the attenuation is only about -3dB. The result is very similar to the clean channel with the LF step between 10Hz and 100Hz.

A19.21.2.2 Triode 2:

The 2nd triode in the cascade drives an interesting output filter that resembles a tone stack with its potentiometers frozen in some static positions. The output of this pre-distortion filter drives the preamp gain control potentiometer, which sets the overall signal level through the rest of the preamp.

A19.21.2.3 Triode 3:

The third triode acts as the voltage amplifier to overdrive the final stage; it is otherwise relatively bland and is identical to several other stages in the design.

A19.21.2.4 Triode 4:

The last triode is configured generally the same as the others for gain and low frequency response. It drives the passive tone stack. There will be about -10dB of loss through the tone stack (at best) and then the output master volume control whose max gain is 0.09 or -21dB.

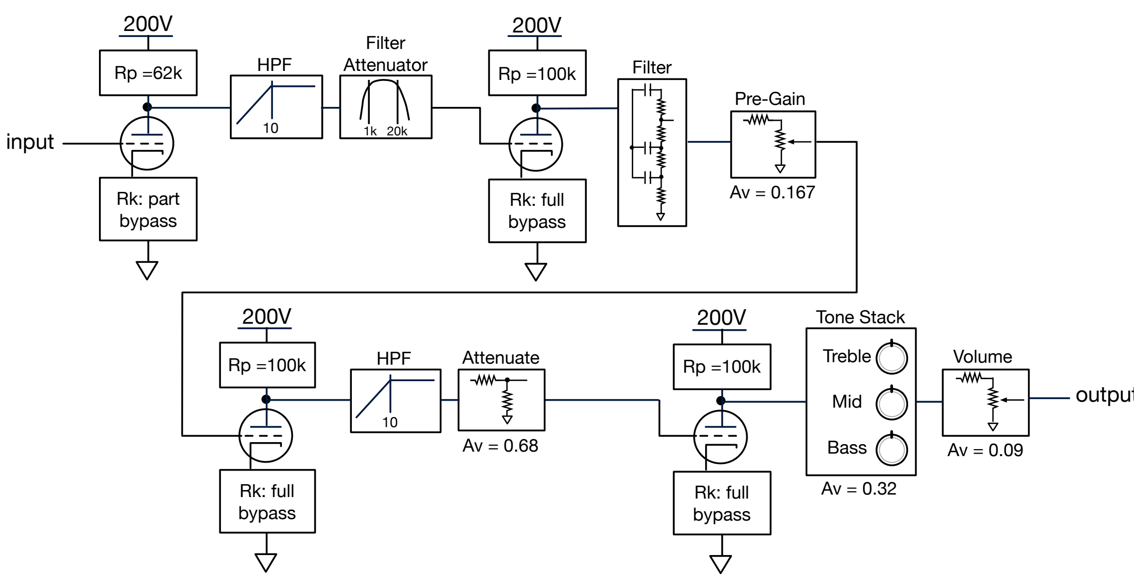


Figure A19.61: the DC-2 lead channel in block diagram form

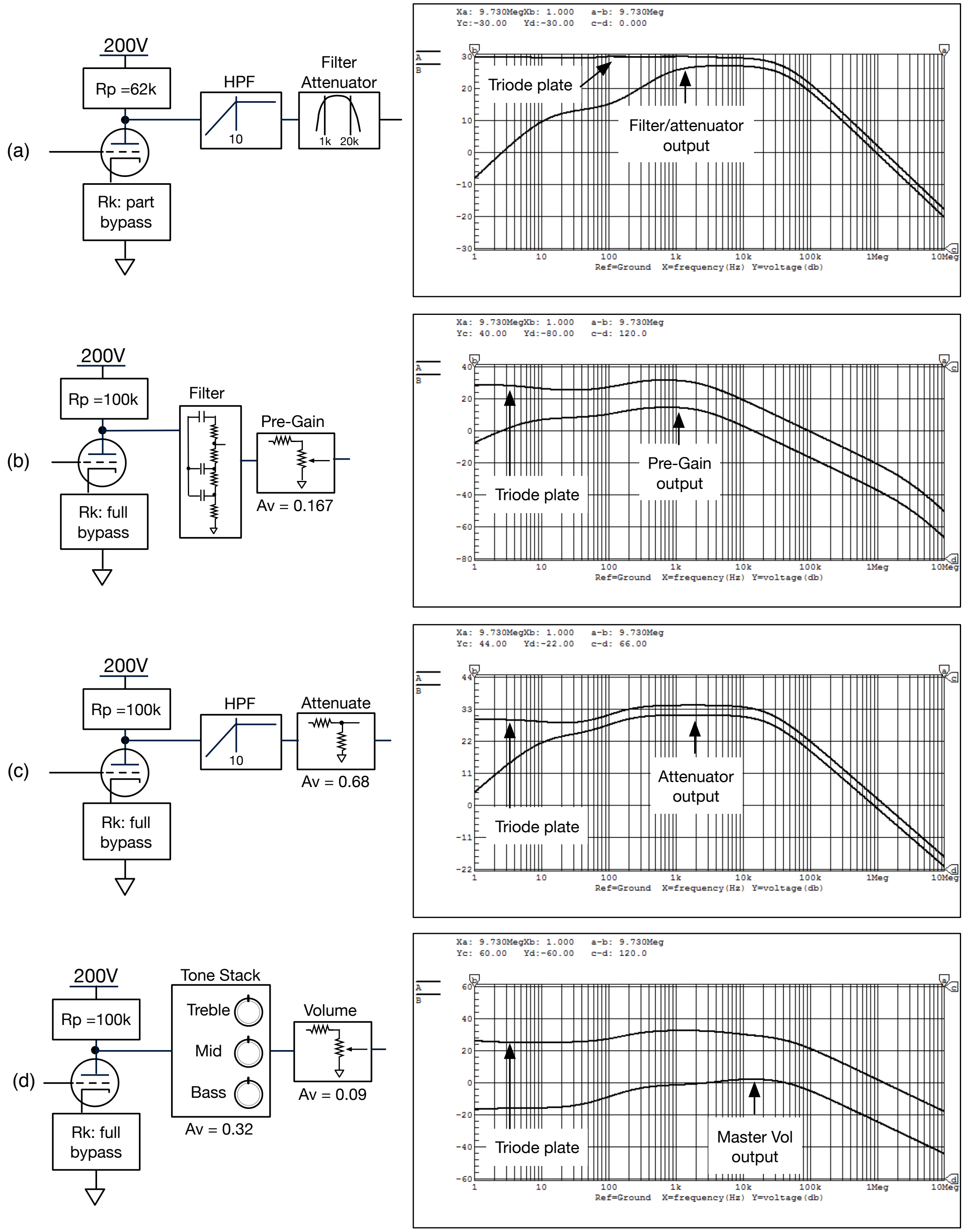


Figure A19.62: analysis of each stage of the DC-2 lead channel (a) input amp and first pre-filter (b) the 2nd amplifier triode and the main pre-distortion filter (c) the third voltage amplifier stage overdrives the final stage (d) which includes the tone stack and master volume control for the channel

A19.21.3 Big Muff π Tone Control

The tone control from the classic Electro-Harmonix Big Muff π distortion box is simple and interesting. It only requires one control knob that produces a variety of responses from lowpass to high-shelf to notch. The circuit is simply a parallel LPF and HPF with a blend knob that sets the ratios of the outputs. My version is shown in Figure A19.63 and is a bit different in construction in that it uses a high shelf filter rather than the circuit's original HPF. The reason is that the filter's passive nature forces signal to flow through the branches at all time, even when the potentiometer is turned all the way to one direction. You can see in the plot that the shelf's final gain is around -18dB. If you check the references, you can find exact plots from the hardware to compare with.

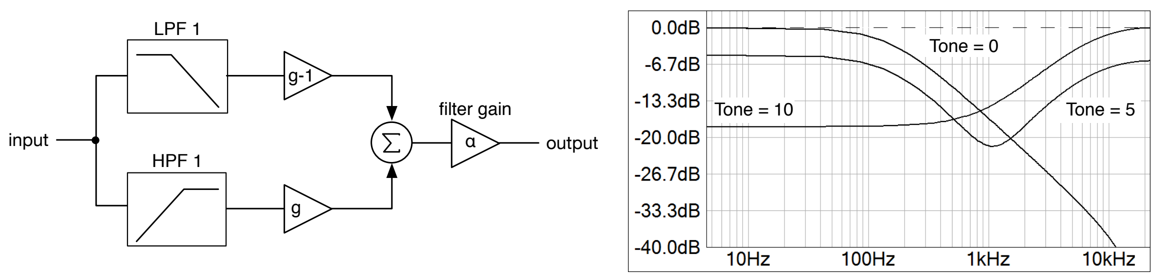


Figure A19.63: the Big Muff tone control block diagram and responses; note that this is my version whose notch at Tone = 5 is at 3kHz; on the original the notch is at 1kHz and there is more loss through the network than my version

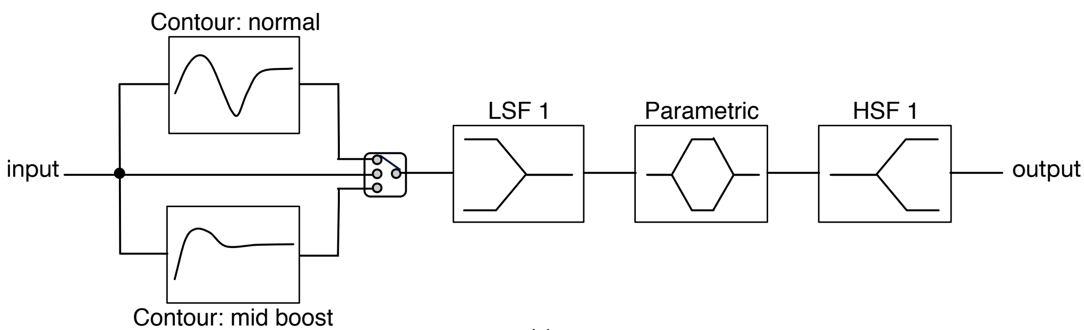
A19.21.4 Guitar Amp Tone Stack

You've probably noticed that I haven't brought up the old passive tone stack designs from the classic tube amps. The reason is that I didn't want to open up a new topic involving these passive tone controls, which could very well have their own Chapter (or whole book). Though the circuits are not too complex, the interaction of the controls is complex. This means that as the user moves the "bass" knob, it will affect the midrange and high frequencies also to some extent. On some tone stacks, the controls are so sensitive that minor changes in them can destroy hours of work trying to get the perfect sound. Have you ever seen guitarists with masking tape markings on their amps for the knobs? The reason is that they've lost valuable settings from the amp being knocked around, or the knob being accidentally altered.

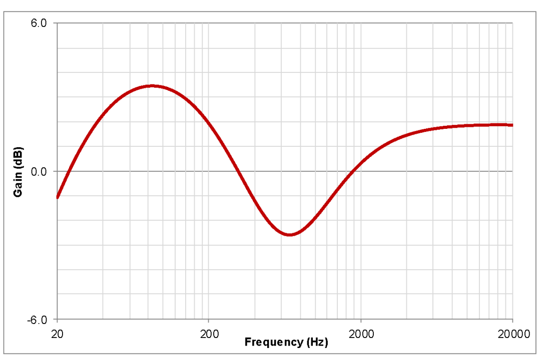
The standard guitar tone stacks produce a noticeable midrange notch when all the controls are set at their 50% (center) positions. In addition, there is about a -10dB loss through the passive network. There are three popular designs: Fender, Marshall and Mesa-Boogie and if you do some googling, you can find circuits for these tone stacks, and perhaps a few algorithms as well. Texas Instruments App Note called *Tone Stack for Guitar Amplifier Reference Design* included with this document includes a nice reference design that doesn't refer to any of the manufacturers themselves.

Instead of messing with a whole new topic, I am going to use a simple three-band "active" EQ but with a contour notch feature to add the "notch when all controls are flat" option if the user wants it. I have prepared the object to match exactly the TI App Note's frequency responses when all controls are set to 5.0 (center position). This is done with a contour filter that processes the audio first, then the 3-band EQ is processed next, which stretches out the original contour filter in a way that is not unlike a tone stack, except that the controls do not interact. In addition, I followed the same boost and cut specifications in the App Note, but I boosted the range of the bass control a bit.

Figure A19.64 shows the block diagram for the tone stack I will use with the guitar amp simulator plugin project. It includes a selectable contour filter that matches the guitar tone stack when all controls are set to the center position named *normal*, as well as an increased mid-gain named *mid-boost*. You may also bypass the contour filter altogether.



(a)



(b)

Table 1. Comparison of Design Goals, Simulation, and Measured Performance

	Goal	Simulated	Measured
THD+N level at 1 kHz	-100 dB (0.001%)	-102.6 dB (0.00074%)	-105.4 dB (0.00054%)
Treble adjustment range	10 dB	10.4 dB	10.6 dB
Mid adjustment range	6 dB	6.1 dB	8.9 dB
Bass adjustment range	15 dB	18 dB	19.2 dB

Table 2. Summary of Tone Stack Component Values and Performance Characteristics

	Components	Values	Behavior
Overall attenuation	P_4	25 kΩ	-7.3 dB
	R_3	33 kΩ	
Bass passband upper cutoff frequency	C_9	22 nF	219 Hz
Bass passband lower cutoff frequency	C_8	22 nF	62 Hz
	P_3	1 MΩ	
Treble high pass cutoff frequency	C_3	470 pF	1.4 kHz
	P_2	250 kΩ	
Mid boost high pass cutoff frequency	C_1	470 pF	700 Hz

(c)

Figure A19.64: (a) block diagram of my guitar tone stack object (b) the settings copied for the all-controls-flat response and (c) tables used to assemble the initial algorithm

Figure A19.65 shows the various frequency responses of the tone stack object. It does a good job of emulating the tone stack without resorting to wave digital filters, where the control interactions could be modeled. Please refer to the TI App Note for all of the details of analog circuit design and operation – plenty of equations to allow you to modify the design for your own purposes. I modeled the contour filter with a combination of bandpass and highpass in parallel and with both DC and Nyquist gains as matched with the App Note. Compare Figure A19.64 (b) and Figure A19.65 (a) which shows the contour filter and all controls flat.

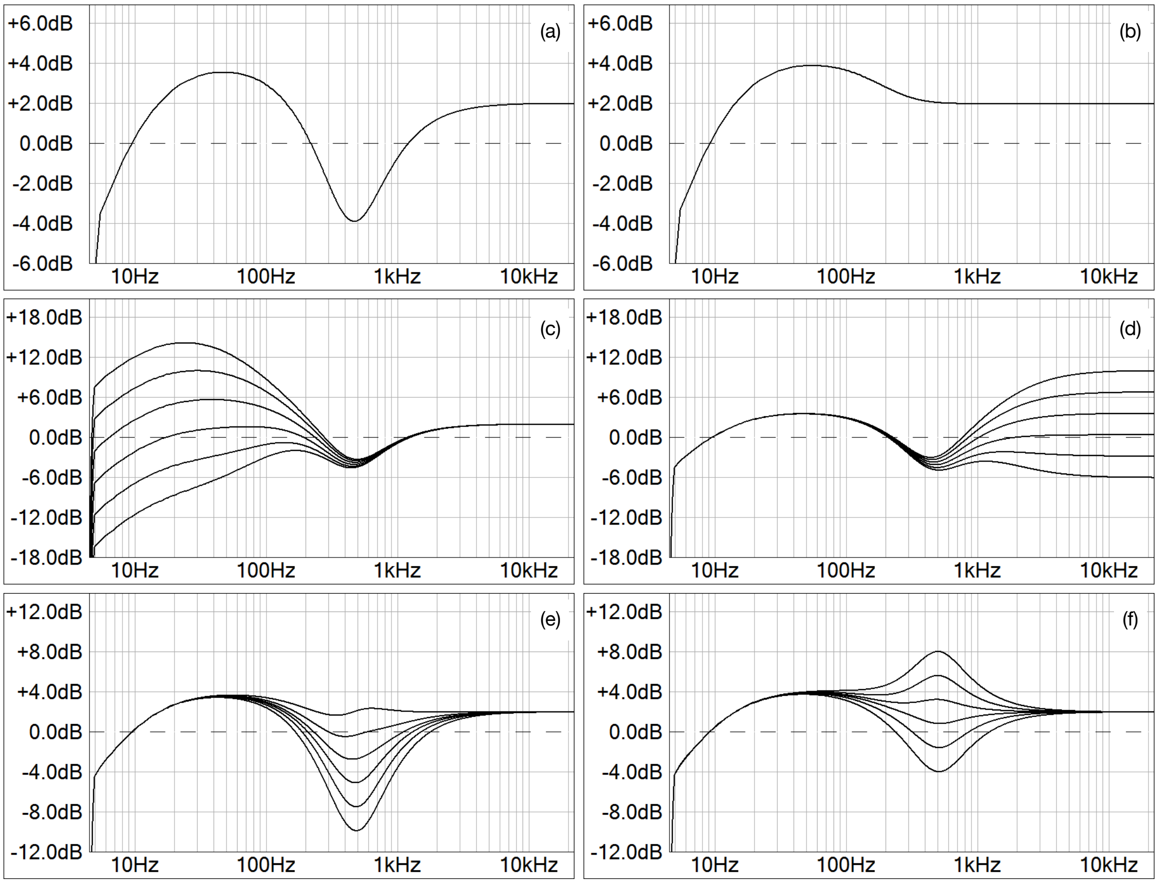


Figure A19.65: frequency responses of my guitar tone stack based on the TI reference design (a) response with all controls flat and normal contour, (b) response with all controls flat and mid-boost contour, (c) response of bass control’s complete excursion with normal contour, (d) response of treble control’s complete excursion with normal contour, (e) response of mid control’s complete excursion with normal contour, and (f) response of mid control’s complete excursion with mid-boost contour

Part V: C++ Objects and Projects

The following collection of C++ objects and sample projects will help get you up and running with your own designs. Remember that we are adding significant harmonics and aliasing will occur even at moderately high frequencies so you will probably want to apply the oversampling techniques of Chapter 22 to these projects (you will need FFTW installed).

All of the objects and code are packaged in the *valves.h* file which #includes the *fxobjects.h* file and uses many of its components

A19.22 Headroom and Knob Control Markings

Before starting, I want to mention a couple of things with these objects and plugins that are different that the other FX projects, including the ones from Chapter 19. I decided that for the Class-A and Class-B algorithms, I would treat audio input levels as volts so that I could keep signal sizes relative to one another and relative to my SPICE simulations and the actual tubes. So we will automatically require headroom in these projects. This means that our internal audio signal's amplitude range can and will be far outside the audio range of [-1.0, +1.0]. In some ways, it is nice that we have the added floating-point range and headroom. We just need to be careful with signal scaling so you will run into several gain and attenuation coefficients. I spent quite some time working on the relative signal sizing through the guitar amp sim plugin and we do need that headroom to work with. The bottom line is to make sure the only distortion you are generating is from your algorithm, and not from clipping the audio waveform on the output.

Secondly, if you play electric guitar, you've probably noticed that many guitar amps do not include numerical values printed on the chassis except for numbers ranging from 0 to 10 or from 1 to 10, or sometimes just tick marks. There are plenty of good reasons for this (how many guitar players really want to know the numerical value of the control?) and we are going to follow suit with many of our parameters operating as GUI knob controls that go from 0 to 10. Internally, we will map these 0 to 10 values to the appropriate ranges used in the algorithms. And, if you want your guitar amp sim plugin to have knobs that go to 11, you are free to do so – just modify the *setParameters* cooking equation.

A19.23 C++ DSP Object: *ClassAValve*

The *ClassAValve* object implements a block diagram that simulates the Class-A valve with grid conduction circuit in Figure A19.45. The object simulates dynamic compression that is inherent in tube circuits, which the FX book objects did not. It uses my method of positive grid compression followed by DC offset extraction and bias shifting described in the previous sections. The *ClassAValve* full block diagram is shown in Figure A19.66.

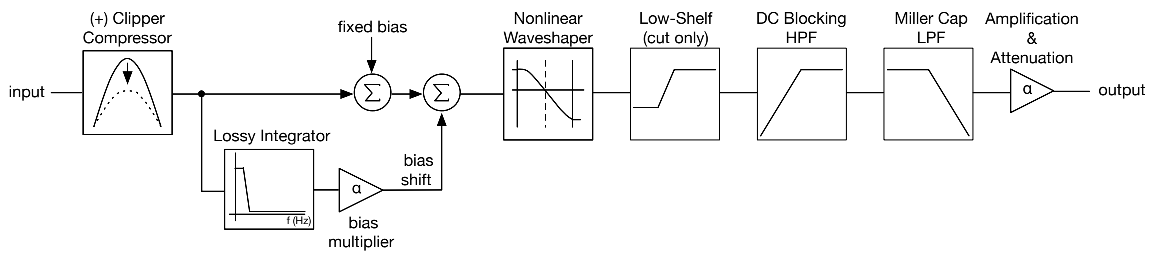


Figure A19.66: block diagram of the *ClassAValve* object

We will discuss the waveshaper in more detail shortly. The other blocks have already been discussed in the text and include:

- Grid-conduction clipper/compressor and integrator to perform DC bias shifting
- Bias shifting can be scaled for sensitivity control
- The low shelf filter is for emulating the cathode bypass cap
- The DC blocking filter is for emulating the DC blocking cap at the plate output
- The Miller Cap LPF is a high-frequency reduction filter that emulates the parasitic capacitance between the plate and grid that forms a negative feedback path for very high frequencies; this and the DC blocking filter together form the valve's band-pass filter response
- Amplification and attenuation controls set the final output level for the valve emulator

A19.23.1 *ClassAValve*: Enumerations and Data Structure

The object is updated via the *ClassAValve Parameters* custom data structure that contains members for adjusting its internal objects, and enabling/disabling its filters. The member naming should suffice in making the connection to the underlying object parameters.

```
struct ClassAValveParameters
{
	ClassAValveParameters() {}

	ClassAValveParameters& operator=()
	(removed here)

	// --- for emulation of LF shelf from cathode bypass cap
	double lowFrequencyShelf_Hz = 100.0;
	double lowFrequencyShelfGain_dB = 0.0;

	// --- for emulation of output HPF for DC blocking
	double dcBlockingLF_Hz = 10.0;

	// --- for emulation of parasitic Miller Cap that reduces HF
	double millerHF_Hz = 10000.0;

	// --- can adjust this too
	double integratorFc = 5.0;
```

```
	// --- clip and threshold points
	double clipPointPositive = 4.0;     // --- SPICE data
	double clipPointNegative = -1.5;    // --- see addendum
	double gridConductionThreshold = 1.5;

	// --- this can scale the added DC offet amount
	double dcShiftCoefficient = 1.0;
	double waveshaperSaturation = 1.0;  // --- the (k) value

	// --- I/O scaling:
	double inputGain = 1.5; // --- this is because our bias is -1.5V
	double outputGain = 1.0;// --- reduce output back to [-1, +1]

	// --- return data (optional)
	double dcOffsetDetected = 0.0;
};
```

A19.23.2 *ClassAValve*: Members

Tables A19.2 and A19.3 list the *ClassAValve* member variables and member functions. As an aggregate object it is fairly simple.

ClassAValve Member Variables		
Type	Name	Description
ClassAValveParameters	*parameters*	Custom data structure
double	*sampleRate*	Current sample rate
ZVAFilter	*lossyIntegrator[2]*	DC detection filters
AudioFilter	*lowShelvingFilter*	Low shelving filter
AudioFilter	*dcBlockingFilter*	Output HPF
AudioFilter	*upperBandwidthFilter*	Output LPF

Table A19.2: the *ClassAValve* member variables

ClassAValve Member Functions		
Returns	Name	Description
ClassAValveParameters	*GetParamerers*	Get all parameters at once
void	*SetParamerers* Parameters: - *ClassAValveParameters* _parameters	Set all parameters at once
bool	*Reset* Parameters: - *double* sampleRate	Resets all filters
double	*processAudioSample* Parameters: - *double* xn	Process input xn through the model
double	*doValveGridConduction* Parameters:	Check the value of the input and apply grid

	- *double* xn - *double* *gridConductionThreshold*	compression as needed; returns effected signal
double	*doValveEmulation* Parameters: - *double* xn	Do the complete valve emulation as per the block diagram

Table A19.3: the *ClassAValve* member functions

A19.23.3 *ClassAValve*: Programming Notes

The *ClassAValve* object combines the three filters, the grid conduction circuit, and a waveshaper in a simple C++ package. Note that the output gain is not given in dB, but as a cooked multiplier value because the gain parameter is mimicking a resistor voltage divider only. The saturation control implements the tube's internal gain function.

- reset the object to prepare it for streaming
- set the object parameters in the *ClassAValveParameters* custom data structure
- call *setParameters* to update the calculation type
- call *processAudioSample*, passing the input value in and receiving the output value as the return variable

Function: *reset & setParameter*
Details: The *reset* function sets up the three filters; the *setParameter* function simply forwards the values to the sub-objects. In the *reset* function, the only thing that is very different from all the other filtering stuff we've done is the lossy integrator setup. The lossy integrator is a LPF with an extremely low cutoff frequency and I use the Zavalishin VA filter here because it already contains the trapezoidal integrator. We just add the lossy part with a non-DC cutoff frequency.

```
virtual bool reset(double _sampleRate)
{
      // --- do any other per-audio-run inits here
      sampleRate = _sampleRate;

      // --- integrators
      lossyIntegrator[0].reset(_sampleRate);
      lossyIntegrator[1].reset(_sampleRate);

      ZVAFilterParameters params = lossyIntegrator[0].getParameters();
      params.filterAlgorithm = vaFilterAlgorithm::kLPF2;
      params.fc = 5.0; // 5 Hz
      lossyIntegrator[0].setParameters(params);
      lossyIntegrator[1].setParameters(params);

      // --- other filters
```

Function: *processAudioSample*
Details: This function first performs the waveshaping, then includes the optional filtering operations. Parameters are set in the custom data structure. There are three different

waveshapers that are selected with the *waveshaper* parameter. The fuzz asymmetrical waveshaper includes an asymmetry control on the range of [-1.0, +1.0] and it should be noted that this waveshaper's output gain is dependent on the *saturation* control as well as the gains.

```
virtual double processAudioSample(double xn)
{
	double yn = 0.0;

	// --- input scaling
	xn *= parameters.inputGain;

	// --- (1) grid conduction check, prior to waveshaping
	xn = doValveGridConduction(xn, parameters.gridConductionThreshold);

	// --- (2) detect the DC offset that the clipping may have caused
	double dcOffset = lossyIntegrator[0].processAudioSample(xn);

	// --- save this - user may indicate it in a meter if they want
	//     This is a bipolar value, but we only do DC shift for
	//     *negative* values so meters should be aware
	parameters.dcOffsetDetected = dcOffset;

	// --- process only negative DC bias shifts
	dcOffset = fmin(dcOffset, 0.0);

	// --- (3) do the main emulation
	yn = doValveEmulation(xn,
	                      parameters.waveshaperSaturation,
	                      parameters.gridConductionThreshold,
	                      dcOffset*parameters.dcShiftCoefficient,
	                      parameters.clipPointPositive,
	                      parameters.clipPointNegative);

	// --- (4) do final filtering
	//
	// --- remove DC
	yn = dcBlockingFilter.processAudioSample(yn);

	// --- LF Shelf
	yn = lowShelvingFilter.processAudioSample(yn);

	// --- HF Edge
	yn = upperBandwidthFilter.processAudioSample(yn);

	// --- (6) final output scaling and inversion
	yn *= -parameters.outputGain;

	return yn;
}
```

A19.23.4 *ClassAValve* Waveshaping and the 3/2 Power Law

The 3/2 power law looks like it would be the obvious answer for a waveshaper and you may certainly modify the *ClassAValve* to use your own version of waveshaping. But, there are a few issues to consider:

- There is no "perfect" 3/2 power law waveshaper. The shape of the 3/2 power law transfer function curve ranges from nearly linear to very curved; lowering the plate voltage will make the transfer function more nonlinear, holding all other variables constant

- Ideally, the tube enters cut-off as soon as the grid-voltage reaches the terminal value that we found from the data sheet load line plot; this area is shown in Figure A19.67 at the bottom. This is the point where the 3/2 power law will produce the most nonlinearity. But, the tube does not cut-off instantly as electrons can "sneak" around the grid for a while as the V_G drops. This means that the tube goes into cut-off in a gradual manner – we would say that it is soft clipped. Because of phase inversion, the negative grid voltage is flipped so that cut-off clipping results in the tops of the waveforms being soft-clipped.

- Ideally, the tube doesn't conduct when V_{GK} is greater than 0.0 but we know that it actually does conduct, and a good amount of the conducted current will flow through the plate since it is at a massively higher voltage level than the grid.

- The tube will not begin clipping the bottoms of the output waveform until the DC offset in the signal has become so large that the shift it creates exceeds the operating limits (voltage rails) of the circuit; after this point, both the tops and bottoms of the resulting waveform will be clipped.

All of these additions, and especially the fact that the 3/2 power law involves multiple variables of operation including the tube-geometry factor, points us in the direction of finding a good solution that we can manipulate to modify the tube emulation behavior. For the *ClassAValve* object I made the decision to follow the block diagram and emulation as closely as possible.

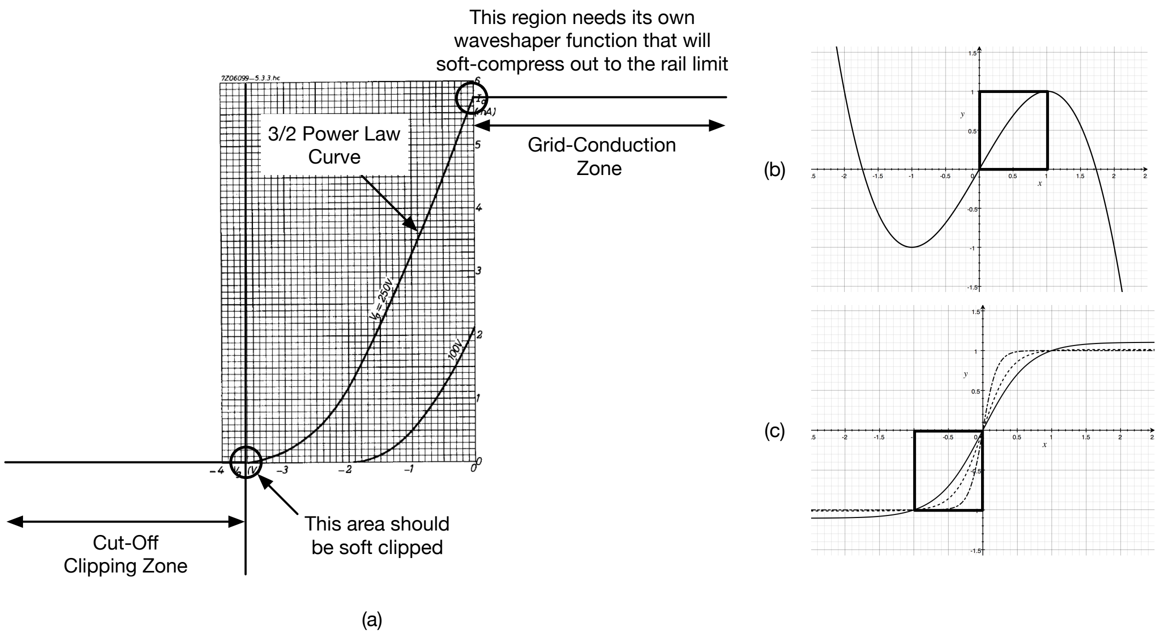


Figure A19.67: (a) the ideal 3/2 power law curve with hard-clipped cut-off and grid conduction zones added (b) the Array waveshaper's positive zone and (c) the hyperbolic tangent waveshaper at a variety of *k* (saturation) values with the negative zone enclosed

Figure A19.67 shows how the 2/3 power law would look if used as a waveshaper. The cut-off and grid conduction zones are simply hard clipped. The V_p = 250V curve is being used and you can see the V_p = 100V curve below it – note that they do not have identical shapes.

A19.23.4.1 Cut-Off & Normal Tube Conduction Zones

The cut-off zone is outside the final cut-off grid voltage and it will cause the signal to clip at the bottom (which becomes clipping at the top of the inverted output signal). But the clipping action is soft. In Figure A19.67 (c) notice the area inside the box – this is the normal tube conduction zone where we want to emulate the 3/2 power law. The *tanh* waveshaper's adjustable *k* (saturation) value will allow a variety of curves that will be normalized to 1.0 at the cut-off input and output voltage. The *tanh* waveshaper will produce soft clipping even at extreme settings of the saturation value. Emulating with this waveshaper gives us that degree of freedom in playing with the design. Note that you are encouraged to create your own 3/2 power law version of the waveshaper – you don't need to rely on my version. But remember to make the clip points soft.

A19.23.4.2 Grid-Conduction Zone

The grid-conduction zone is modeled with two algorithms. The first is a peak waveform compressor that applies the grid conduction, compressing the tops of the waveforms down according to a soft-knee compression curve that I found empirically with SPICE simulations. Figure A19.68 shows the two blocks of processing that are performed when the input signal makes $V_{GK} > 0$.

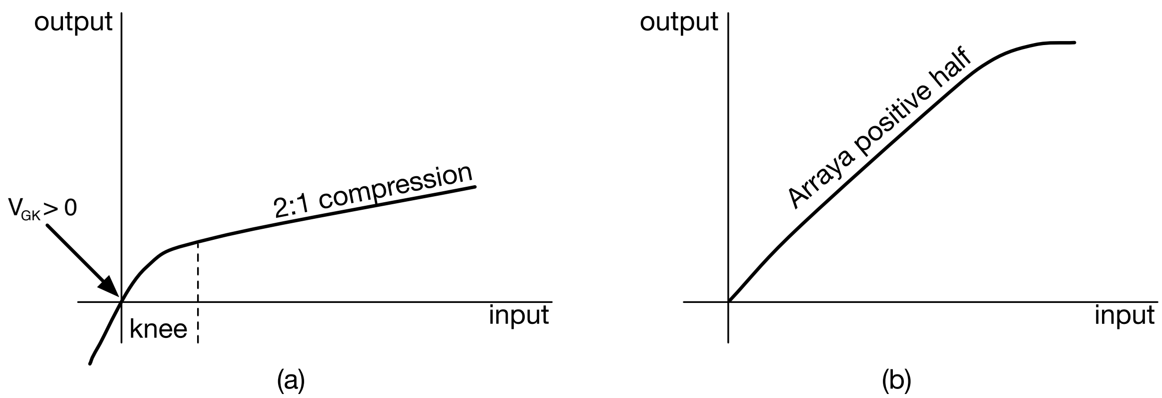


Figure A19.68: (a) the 2:1 peak compressor used on the positive portions of the waveform when $V_{GK} > 0$ and (b) the top portion of the Arraya waveshaper (see FX book)

The majority of the nonlinear processing on the positive peaks comes from the peak compressor. One compressed, the DC offset is extracted with the lossy integrator. The variable DC offsets are added and the whole waveform is shifted accordingly. The top portions of that waveform will be processed through the quite-linear and ultra-soft-clip Array waveshaper function. I have also included the code for a 4:1 compression curve that clips more of the input signal but found that it was not sonically different enough to warrant more work there.

Figure A19.69 shows the complete waveshaper and all ranges of values for the input signal. You can see that there are four zones:

1. cut-off when $V_{GK} < -3V$
2. normal tube conduction for negative input values above -3V
3. Array waveshaper for grid conduction
4. Saturation of operating limits when $V_{GK} > 2.5V$

You can see from the plot that it is nearly symmetrical, with upper clipping occurring when $V_{GK} > 2.5V$ and lower clipping when $V_{GK} < -3.0V$. I set the system up this way so that it is easy to scale out any of the waveshapers to cover whatever input/output voltage sizes you wish to implement. The hybrid approach here lets you substitute in different nonlinear sections and then scale the input/output as you like.

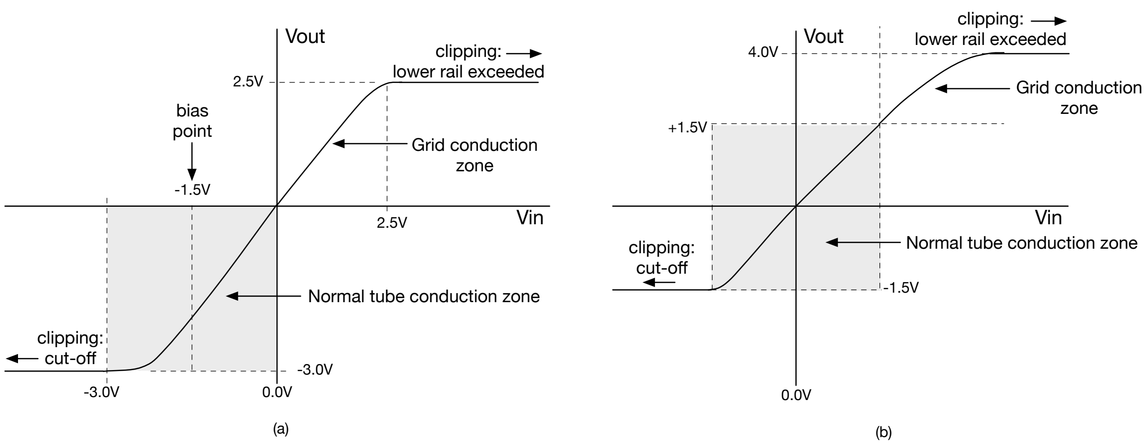


Figure A19.69: (a) the complete waveshaper for the *ClassAValve* object and its four areas of operation and (b) the same transfer function shifted by +1.5V so that the bias value is 0.0 (this is discussed later) – the *y*-axis has normalized values as the tube circuit will have some amplification factor; since we adjust gain between stages, it is easier to have the transfer function operate with normal values that are easily scaled

A19.23.5 *ClassAValve* Implementation Notes

The majority of the code for the *ClassAValve* is fairly straightforward since much of it involves filtering that we've already covered. If you want to poke around the code in *valves.h*, here are some highlights to look for.

A19.23.5.1 DC Detection

I am using a DC detector built as a lossy integrator, which is a lowpass filter with a very low cut-off frequency. I am using the Zavalishin structures because they are already setup as integrator-based topologies. Another option for DC detection would be a moving average (MA) filter. One drawback to extracting DC is the impulse response of the lossy integrator. The longer this impulse response time is, the slower the DC value will move around. If you set the lossy integrator to a ridiculously low f_c value like 0.1Hz, then the impulse response time is heavily stretched out. If the amplitude of the signal suddenly changes, for example due to the user adjusting the volume knob very quickly, then the DC tracking will be too slow to keep up and the system will seem to shut down, then slowly come back up to volume, as you rapidly reduce the amplitude to a new fixed location.

To get over this issue, we simply raise the cut-off frequency f_c value to something around 1Hz to 5 Hz. This speeds up the DC extraction. Of course the higher we make this value, the more non-DC information it will transmit, however if used with the 6-string guitar, whose low E hits about 88 Hz, then there is plenty of room to experiment. I achieved very acceptable results with the f_c value at 5.0Hz. If you wanted to use a moving average filter, you would have a similar issue because it operates on a sliding window of data. The larger the window, the more accurate the averaged value will be (and the average value is the DC component), but the slower the extracted DC value will move. So, you would respond by making the window shorter, however if it is too short, then the DC value will tend to bound around with a lot of ripple.

A19.23.5.2 *doValveGridConduction*

This simple function implements the peak compressor that operates on absolute sample values; there are no detectors or other conversion. This compression scheme came directly from my own SPICE simulation data and a simple exponential curve fit was applied.

```
inline double doValveGridConduction(double xn,
                                    double gridConductionThreshold)
{
      if (xn > 0.0)
      {
            // --- check how far above clip level we are
            double clipDelta = xn - gridConductionThreshold;

            // --- negative only check
            clipDelta = fmax(clipDelta, 0.0);

            // --- compression value
            double compressionFactor = 0.4473253 +
                              0.5451584*exp(-0.3241584*clipDelta);

            return compressionFactor*xn;
      }
      else
            return xn;
}
```

A19.23.5.3 *doValveEmulation*

This function applies the waveshaping and bias shifting that we've discussed so far. The *k* variable is the waveshaper saturation value. Now there is one issue with trying to mimic the fixed bias in the tube – if we add a fixed DC bias to the signal, it will usually cause a loud thump or click at the very beginning of the audio selection that is being rendered. It will take some samples before the HPF can properly remove the added DC. To accommodate this and prevent the turn-on thump, I have shifted the entire transfer function in Figure 19.69 (a) by +1.5V so that it looks like Figure 19.69 (b). This simply shifts the reference point so that the fixed DC offset effectively becomes 0.0 – everything else is the same. You can see the reference shifting code in the valve emulation function.

The two clip points then become -1.5V for cut-off and +4.0V for grid-conduction saturation. To emulate the 3/2 power law, we observe that for the positive portion of the input waveform, the amplification is nearly linear. For the lower (-) portion of the waveform, I use a *tanh* waveshaper. The ability to adjust the *k* (saturation) value allows you to make higher or lower gain models. For the positive portion that is in grid conduction, I use the grid compression and the Arraya waveshaper (described in the FX book).

You can also notice the scaling that is done – as a reader exercise, make sure you follow this part in the code as well because I use a lot of headroom for the *ClassAValve* modeling – that is, the internal signals may be well above [-1.0, +1.0] and can reach values in the 100's because I am modeling the signal in volts here.

```
inline double doValveEmulation(double xn,
                               double k,
                               double gridConductionThreshold,
                               double variableDCOffset,
                               double clipPointPos,
                               double clipPointNeg)
{
      // --- fixed DC offset
      xn += variableDCOffset;
      double yn = 0.0;

      // --- NOTE: the whole transfer function is shifted so the normal
      //           DC operating point is 0.0 to prevent massive
      //           clicks at the beginning of a selection due to the
      //           temporary DC offset that occurs before the HPF can
      //           react
      //
      // --- top portion is arraya
      if (xn > gridConductionThreshold)
      {
            if (xn > clipPointPos)
                  yn = clipPointPos;
            else
            {
                  // --- get into the first quadrant for Arraya @ (0,0)
                  xn -= gridConductionThreshold;

                  // --- signal should be clipped/compressed prior
                  if (clipPointPos > 1.0)
                        xn /= (clipPointPos - gridConductionThreshold);

                  // --- arraya
                  yn = xn*(3.0 / 2.0)*(1.0 – (xn*xn) / 3.0);

                  // --- scale by clip point positive
                  yn *= (clipPointPos – gridConductionThreshold);

                  // --- undo scaling
                  yn += gridConductionThreshold;
            }
      }
      else if (xn > 0.0) // --- ultra linear region
      {
            // --- fundamentally linear region of 3/2 power law
            yn = xn; // feel free to experiment here
      }
      else // botom portion is tanh( ) waveshaper - EXPERIMENT!!
      {
            if (xn < clipPointNeg)
                  yn = clipPointNeg;
            else
            {
```

```
                // --- clip normalize
                if (clipPointNeg < -1.0)
                      xn /= fabs(clipPointNeg);

                // --- the waveshaper
                yn = tanh(k*xn) / tanh(k);

                // --- undo clip normalize
                yn *= fabs(clipPointNeg);
          }
      }
      return yn;
}
```

A19.23.5.4 *processAudioSample*

This function is where all the theory comes together. Make sure you can follow it along with the block diagrams in this text. The last part of the function is just filtering and scaling – not too difficult. The first part is where the clipping and DC extraction occurs.

```
// --- do the valve emulation
virtual double processAudioSample(double xn)
{
      double yn = 0.0;

      // --- input scaling
      xn *= parameters.inputGain;

      // --- (1) grid conduction check, done prior to waveshaping
      xn = doValveGridConduction(xn,
                                parameters.gridConductionThreshold);

      // --- (2) detect the DC offset that the clipping may have caused
      double dcOffset = lossyIntegrator[0].processAudioSample(xn);

      // --- save this - user may indicate it in a meter if they want
      //     Note that this is a bipolar value, but only DC shift for
      //     *negative* values so meters should be aware
      parameters.dcOffsetDetected = dcOffset;

      // --- process only negative DC bias shifts
      dcOffset = fmin(dcOffset, 0.0);

      // --- (3) do the main emulation
      yn = doValveEmulation(xn,
                            parameters.waveshaperSaturation,
                            parameters.gridConductionThreshold,
                            dcOffset*parameters.dcShiftCoefficient,
                            parameters.clipPointPositive,
                            parameters.clipPointNegative);

      // --- (4) do final filtering
      //
      // --- remove DC
      yn = dcBlockingFilter.processAudioSample(yn);
```

```
        // --- LF Shelf
        yn = lowShelvingFilter.processAudioSample(yn);

        // --- HF Edge
        yn = upperBandwidthFilter.processAudioSample(yn);

        // --- (5) final output scaling and inversion
        yn *= -parameters.outputGain;

        return yn;
}
```

A19.23.6 Results and SPICE Simulations

The *ClassAValve* object was built, in part, using SPICE simulation data. The main data used was the positive compression information when $V_{GK} > 0$. Figure A19.70 shows some SPICE output data compared with data from the *ClassAValve* object driven with a sinusoidal signal that is made to be larger and larger, eventually exceeding $V_{GK} > 0$ and then shortly after that, exceeding the operating range (rails) of the amplifier. Figures A19.71 – A19.74 shows the SPICE results and actual object performance. Note that the SPICE FFTs have a longer time base (out to 30kHz) when making comparisons in frequency. In the grid-conduction cases, notice how the duty cycle of the output waveforms change – a clear indication of the grid-conduction positive waveform compression.

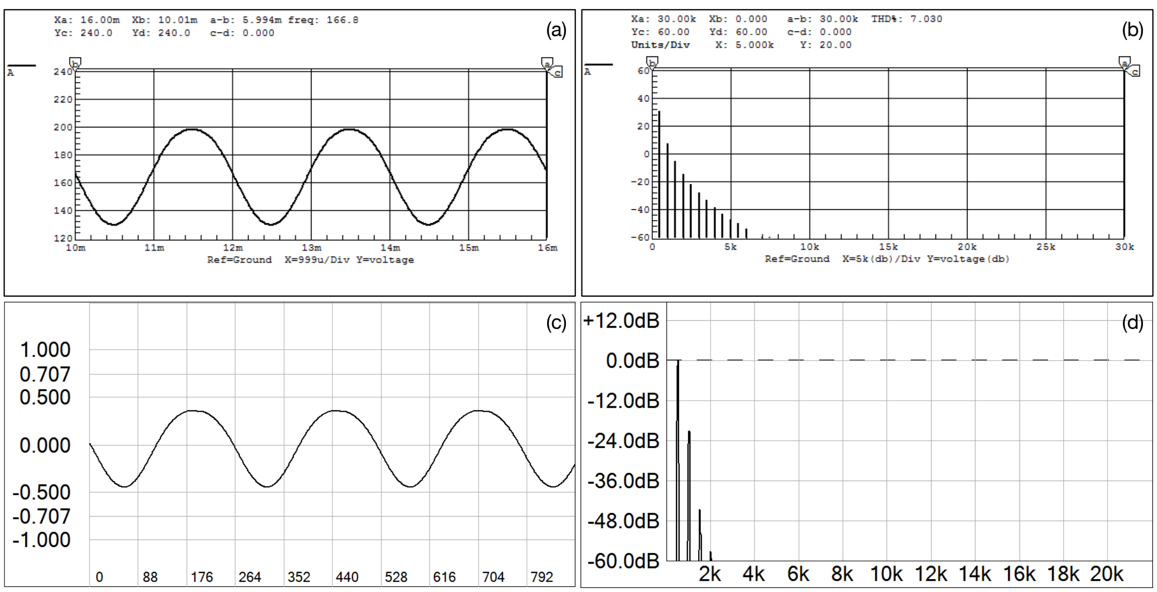


Figure A19.71: SPICE plots vs. *ClassAValve* object for an input level that just hits the grid voltage level prior to cut-off clipping (a) SPICE time and (b) frequency with actual object performance in time (c) and frequency (d)

For the next group of plots, the only thing I have changed for both SPICE and actual performance testing is the input amplitude of the driving signal. All products you see here are a result of input amplitude changes only. In Figure A19.72 the tube has just entered the cut-off clipping state.

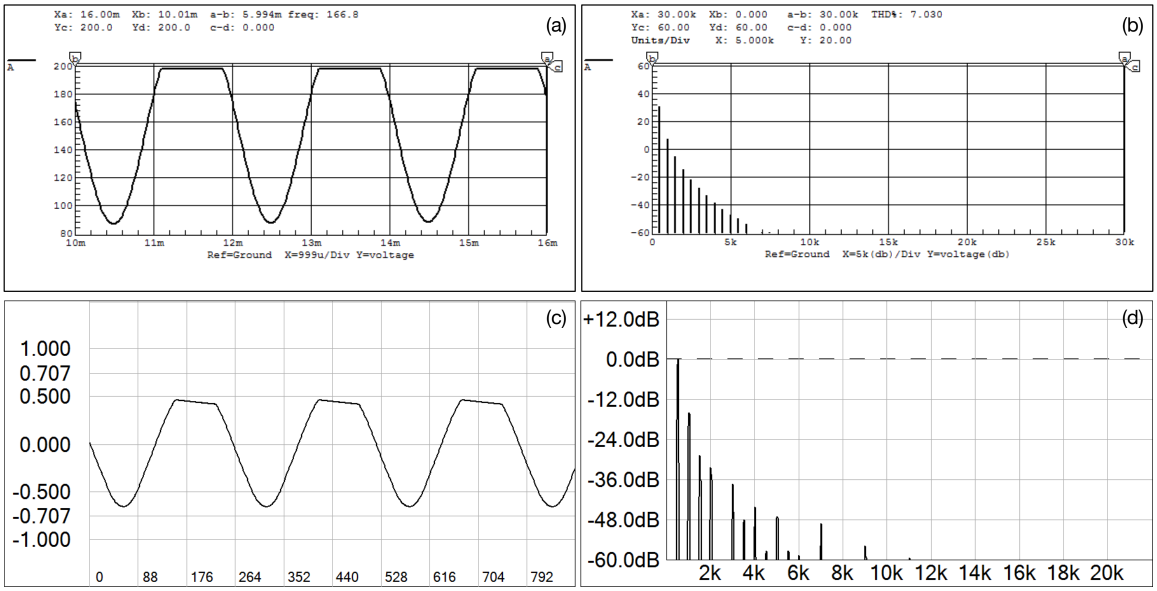


Figure A19.72: SPICE plots vs. *ClassAValve* object for an input level that has just created tube cut-off clipping (a) SPICE time and (b) frequency with actual object performance in time (c) and frequency (d)

Increasing the signal input size further creates grid-conduction and upper waveform compression. In Figure A19.73 the tube is in full grid-conduction and is just on the verge of clipping at both rails (0V and B+).

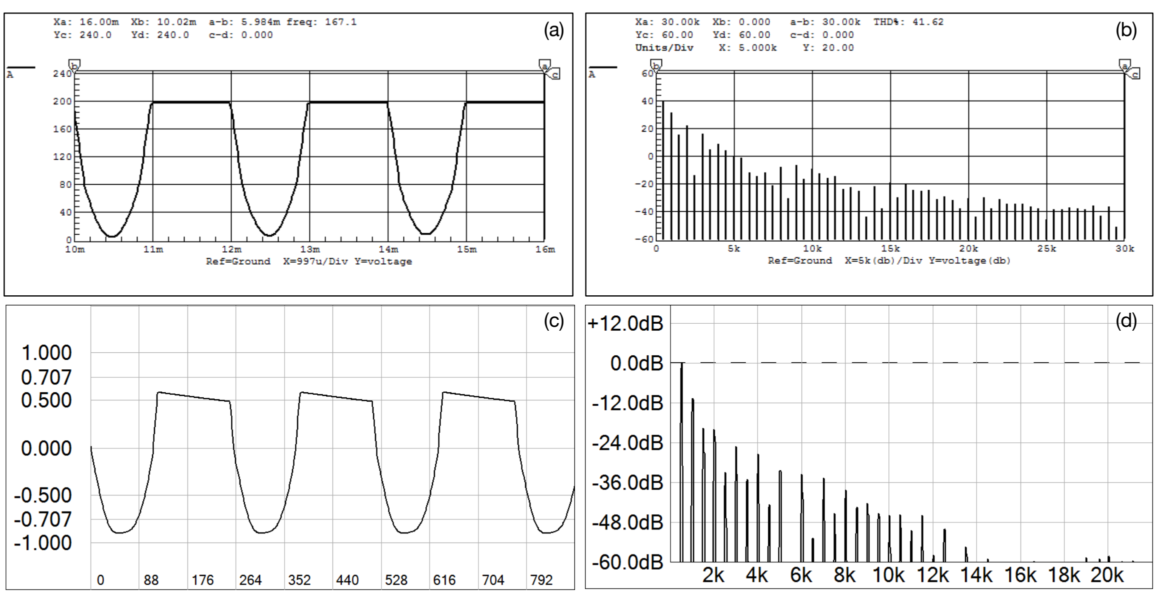


Figure A19.73: SPICE plots vs. *ClassAValve* object for an input level that has started grid-conduction; notice the elongated bottom halves for (a) SPICE time and (b) frequency with actual object performance in time (c) and frequency (d)

Finally, in Figure A19.73 the tube has hit the operating limits in both cut-off as well as grid-conduction hard clip. Increasing the input signal level beyond this point will only result in more squared-off with the 2nd harmonic still more emphasized.

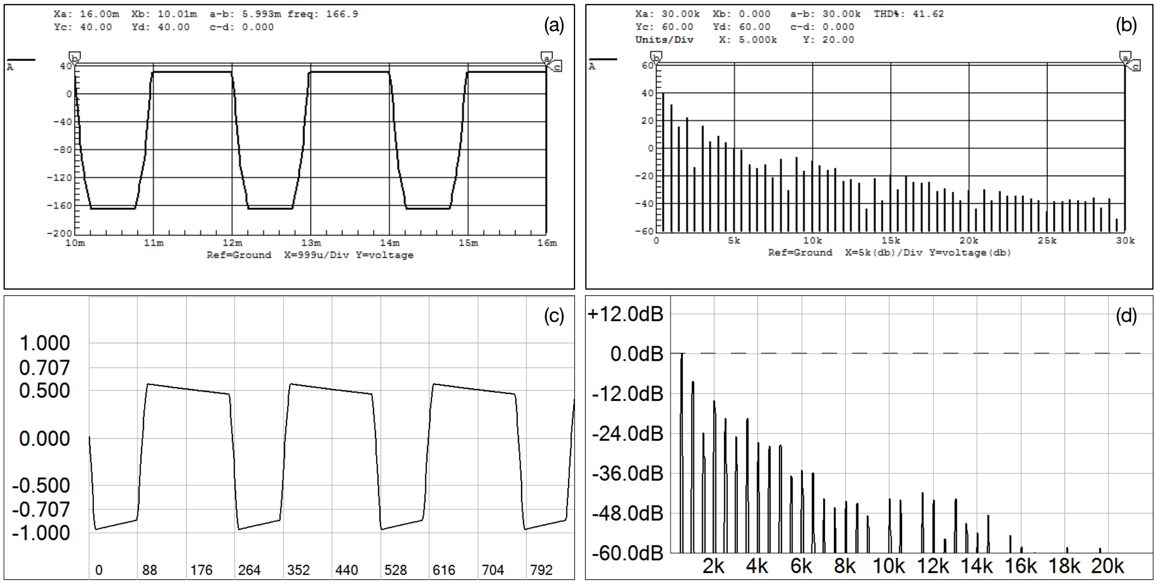


Figure A19.74: SPICE plots vs. *ClassAValve* object for an input level that has caused the tube to clip at both the cut-off and grid-saturation levels (a) SPICE time and (b) frequency with actual object performance in time (c) and frequency (d)

A19.24 C++ DSP Object: *ClassBValvePair*

The *ClassBValvePair* object aims to emulate the Class-B pentode output circuitry of a tube amplifier. As its name implies, it emulates the pair of triodes together performing the Class-B amplification. In this object, I made some changes to make the coding simpler. The DC bias shifts are done in a different manner because the waveshaper is designed to operate on bipolar data (i.e. not fully negative DC shifted signals). The reason is that the Class-B amplifier will produce symmetrical distortion, and the behavior of the crossover distortion (or more accurately “crossover compression”) on the signal makes the waveshaper simpler. There is no need for upper and lower waveshaper portions in this design (though you could certainly implement it that way if you wish. In addition, this object implements two different Class-B emulation algorithms; the Poletti algorithm that is fully documented in the attached patent and my own Pirkle emulation, which is similar to a Peavey TransTube approach blended with the Yamaha paradigm of equating signal amplitude with the eventual DC offset that will occur due to grid-conduction. Figure A19.75 shows the two block diagrams for the algorithms with the DC blocking and HF compensation filters (HPF and LPF respectively) added to the output section. These components are shared between the two algorithms. Numerous time and frequency output plots for both algorithms have already been presented in Figures A19.50 and A19.52 so they will not be repeated here.

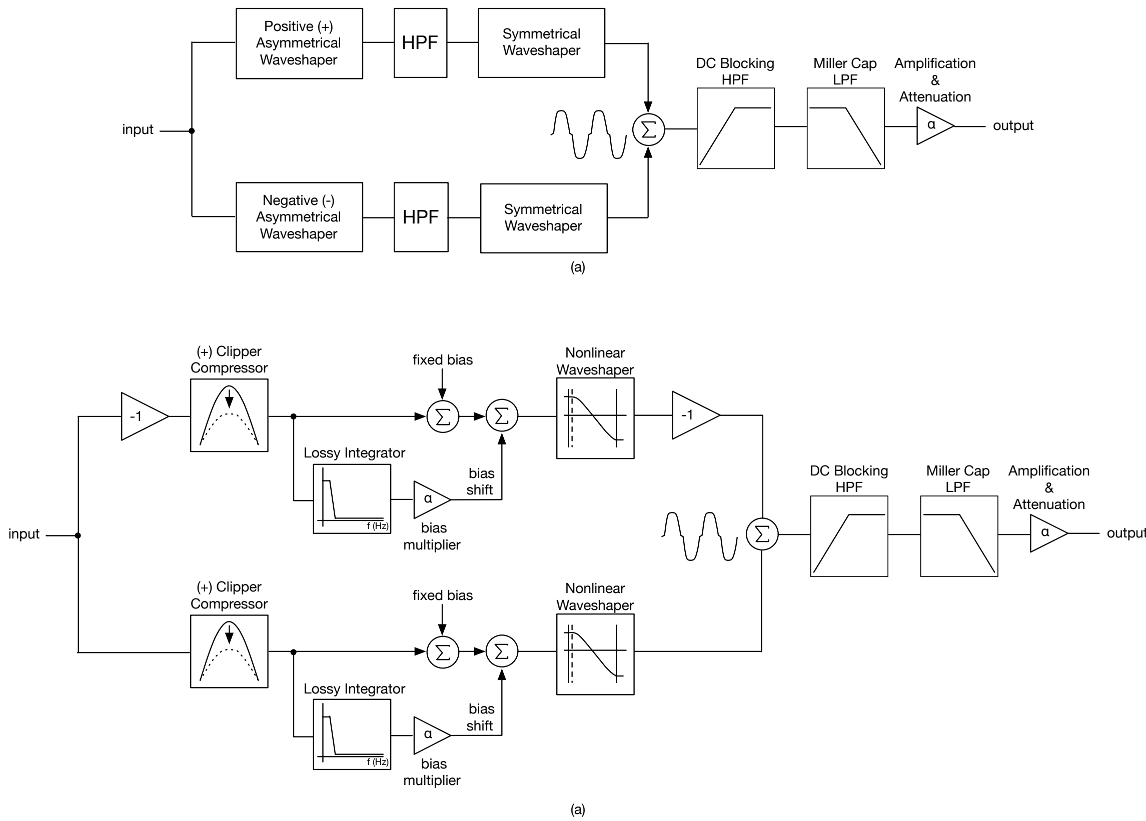


Figure A19.75: the two Class-B emulation block diagrams for (a) Poletti and (b) Pirkle's Class-B emulator.

A19.24.1 *ClassBValvePair*: Enumerations and Data Structure

The object is updated via the *ClassBValvePairParameters* custom data structure that contains members for adjusting its internal objects, and enabling/disabling its filters. The member naming should suffice in making the connection to the underlying object parameters. There is one custom strongly typed *enum* for the algorithm named *classBType*.

```
// --- chooser for algorithm
enum class classBType{ pirkle, poletti };
```

The custom data structure is larger than the previous one, but this is due to the fact that it supports two algorithms. In reality, the operation is actually a bit simpler than the ClassA case. There are some things to notice:

- The *inputGain* variable sets the sensitivity of the Class-B module, meaning how far into crossover distortion/compression a full-range digital input signal will drive the algorithm

- For this model, I do not run the system in a biased-down configuration so that the whole normal tube conduction zone is in the 3rd quadrant of the I/O function; here I am using a symmetrical waveshaper about 0.0 to mimic the symmetrical distortion that will result when the ½ waveforms are recombined at the output – the result of this is that my bias voltages and clip points are ½ of that for the triode case

- The Poletti algorithm has two waveshapers, one asymmetrical and the other symmetrical so there are multiple sets of coefficients

- I run the algorithm as a purely symmetrical system but Poletti notes that by altering the asymmetrical waveshaper coefficients (one set for the positive branch, the other for the negative) you can create many more distortion algorithms including a hybrid version that has Class-B compression but Class-A nonlinear amplification.

- For the Poletti algorithm, I made changes to the coefficients because the values from the patent were obviously incorrect; you can certainly experiment with these values for many coding sessions

```
struct ClassBValveParameters
{
	ClassBValveParameters() {}

	ClassBValveParameters& operator=() (removed)

	// --- filter stuff
	classBType algorithm = classBType::pirkle;

	// --- blocks DC and bandpasses the signal
	double outputHPF_fc = 1.0;
	double outputLPF_fc = 20480.0;

	// --- I/O scaling
	double inputGain = 50.0;      // --- effective sensitivity
	double outputGain = 0.53;     // --- reduce output to [-1, +1]

	// --- Pirkle Coefficients this is 1/2 of the actual cutoff bi
	double fixedBiasVoltage = -1.5;
	double clipPointPositive = 1.5;
	double dcShiftCoefficient = 0.5;
	double waveshaperSaturation = 1.2;  // --- the (k) value

	// --- Poletti Coefficients
	// --- not same as patent - alot to experiment with here!!
	//
	// --- asymmetrical waveshaper
	double asymWaveshaper_g = 1.70;     // --- gain
	double asymWaveshaper_Lp = 23.6;    // --- positive limit
	double asymWaveshaper_Ln = 0.5;     // --- negative limit

	// --- symmetrical waveshaper
	double symWaveshaper_g = 4.0;       // --- gain
```

```
    double symWaveshaper_LpLn = 1.01;   // --- pos, neg limit

    // --- return data (optional, Pirkle algorithm only)
    double dcOffsetDetectedPos = 0.0;
    double dcOffsetDetectedNeg = 0.0;
};
```

A19.24.2 *ClassBValvePair*: Members

Tables A19.4 and A19.5 list the *ClassBValvePair* member variables and member functions. As an aggregate object it is fairly simple.

ClassBValvePair Member Variables		
Type	Name	Description
ClassBValveParameters	*parameters*	Custom data structure
double	*sampleRate*	Current sample rate
ZVAFilter	*lossyIntegrator[2]*	DC detection filters
AudioFilter	*dcBlockingFilter[2]*	DC blocking from output transformer
AudioFilter	*upperBandwidthFilter*	Upper bandwidth limit filter

Table A19.4: the *ClassBValvePair* member variables

ClassBValvePair Member Functions		
Returns	Name	Description
ClassBValveParameters	*GetParameters*	Get all parameters at once
void	*SetParameters* Parameters: - *ClassBValveParameters* _parameters	Set all parameters at once
bool	*Reset* Parameters: - *double* sampleRate	Resets all filters
double	*processAudioSample* - *double* xn	Process input xn through the model
double	*doPolettiWaveshaper* Parameters: - *double* xn - *double* g - *double* Lp - *double* Ln	Perform Poletti waveshapers (both); the parameters match the names in the Patent
double	*doPirkleWaveshaper* Parameters: - *double* xn - *double* g	Perform the Pirkle waveshaper for Class-B emulation; the fixed DC offset is the tube bias

	- *double* fixedDCOffset - *double* variableDCOffset	voltage and the variable offset is what is detected as a result of grid-conduction

Table A19.5: the *ClassBValvePair* member functions

A19.24.3 *ClassBValvePair*: Programming Notes

The *ClassBValvePair* object combines all of the ingredients needed for both the Poletti and Pirkle algorithms. There is really only one parameter you would allow the user to vary – the *inputGain* value which sets the sensitivity of the Class-B emulator. You can also think of this as a bias condition as well – if the amp is under biased, this value will be low, while if over biased, this value would be high. The operation is fairly straightforward:

- reset the object to prepare it for streaming
- set the object parameters in the *ClassBValveParameters* custom data structure
- call *setParameters* to update the calculation type
- call *processAudioSample*, passing the input value in and receiving the output value as the return variable

Function: *reset & setParameter*
Details: The *reset* function sets up the lossy integrator, DC blocking filter and HF compensation filters identically as with the *ClassAValve* object so there is no need to repeat it here. The *setParameter* function is likewise simple – it just sets the DC blocking and HF compensation filter cutoff values, which will likely never change and will be hard coded.

Function: *processAudioSample*
Details: This function first decodes the algorithm, and then applies the chosen function. The output filtering is the same for each algorithm.

```
virtual double processAudioSample(double xn)
{
	double yn = 0.0;

	// --- add input gain
	xn *= parameters.inputGain;

	// --- two choices here: Poletti or Pirkle
	if (parameters.algorithm == classBType::poletti)
	{
```

The Poletti code exactly follows the patent block diagram that is also reprinted in this document. Notice how everything doubles, as there are two branches to process now. And in the asymmetrical clipping block, note how the arguments are swapped (Lp and Ln) so that each waveshaper performs an identical operation. You can experiment with these values for many more types of distortion outputs. The sequence is: asymmetrical waveshaping -> HPF -> symmetrical waveshaping.

```
// --- (1) asymmetrical waveshaping
double yn_pos = doPolettiWaveShaper(xn, parameters.asymWaveshaper_g,
                                    parameters.asymWaveshaper_Lp,
                                    parameters.asymWaveshaper_Ln);


double yn_neg = doPolettiWaveShaper(xn, parameters.asymWaveshaper_g,
                                    parameters.asymWaveshaper_Ln,
                                    parameters.asymWaveshaper_Lp);

// --- (2) block DC
yn_pos = dcBlockingFilter[0].processAudioSample(yn_pos);
yn_neg = dcBlockingFilter[1].processAudioSample(yn_neg);

// --- (3) symmetrical waveshaping
yn_pos = doPolettiWaveShaper(yn_pos, parameters.symWaveshaper_g,
                             parameters.symWaveshaper_LpLn,
                             parameters.symWaveshaper_LpLn);

yn_neg = doPolettiWaveShaper(yn_neg, parameters.symWaveshaper_g,
                             parameters.symWaveshaper_LpLn,
                             parameters.symWaveshaper_LpLn);

// --- (4) combine output branches
yn = yn_pos + yn_neg;
}
```

The Pirkle code follows exactly with the block diagram in Figure A19.75 (b) and should be relatively simple to follow. Note that my algorithm requires one branch to be inverted because I use only one type of waveshaper algorithm.

```
else // pirkle
{

// --- (1) create two branches, invert signal in one
double xn_Pos = xn;
double xn_Neg = -xn; // (-) branch input

// --- (2) check grid conduction
xn_Pos = doValveGridConduction(xn_Pos);
xn_Neg = doValveGridConduction(xn_Neg);

// --- (3) detect DC offset or pos and neg branches
double dcOffsetPos = lossyIntegrator[0].processAudioSample(xn_Pos);
double dcOffsetNeg = lossyIntegrator[1].processAudioSample(xn_Neg);


parameters.dcOffsetDetectedPos = dcOffsetPos;
parameters.dcOffsetDetectedNeg = dcOffsetNeg;

// --- only use (-) DC offset
dcOffsetPos = fmin(dcOffsetPos, 0.0);
dcOffsetNeg = fmin(dcOffsetPos, 0.0);
```

```
// --- (4) do the shaper
double yn_Pos = doPirkleWaveShaper(xn_Pos,
                        parameters.waveshaperSaturation,
                        parameters.fixedBiasVoltage,
                        dcOffsetPos*parameters.dcShiftCoefficient);

double yn_Neg = doPirkleWaveShaper(xn_Neg,
                        parameters.waveshaperSaturation,
                        parameters.fixedBiasVoltage,
                        dcOffsetNeg*parameters.dcShiftCoefficient);

// --- (5) combine branches (with inversion)
yn = yn_Pos - yn_Neg;
}
```

From this point on, it is just application of filtering, in part to emulate the output transformer. You can do your final voicing tweaking on these filters to adjust the final bass and treble response.

```
// --- adjust the bandpass nature of the output transformer if you like
//
// --- LF Edge
yn = dcBlockingFilter[0].processAudioSample(yn);

// --- HF Edge
yn = upperBandwidthFilter.processAudioSample(yn);

// --- final output scaling
yn *= parameters.outputGain;

return yn;
}
```

A19.24.4 *ClassBValvePair* Implementation Notes

The Poletti algorithm is completely documented in the patent and I tried to make the code here match that patent as closely as I could.

A19.24.4.1 *doPolettiWaveshaper*

This function simply implements the pair of waveshapers from the Poletti patent, one for negative input values and the other for positive ones.

```
double doPolettiWaveShaper(double xn, double g, double Ln, double Lp)
{
      double yn = 0.0;
      if (xn <= 0)
            yn = (g * xn) / (1.0 - ((g * xn) / Ln));
      else
            yn = (g * xn) / (1.0 + ((g * xn) / Lp));
      return yn;
}
```

A19.24.4.2 doPirkleWaveshaper

The waveshaper I chose was the arctangent waveshaper. I like the ability to control the saturation value and thus the hardness (or softness) of the Class-B waveform edges. In addition because of the Class-B nature, I did not need separate upper and lower shapers. After biasing the signal, the waveshaping occurs.

```
double doPirkleWaveShaper(double xn, double g, double fixedDCoffset,
                          double variableDCOffset)
{
      xn += fixedDCoffset;
      xn += variableDCOffset;

      double yn = 1.5*atan(g*xn) / atan(g);

      return yn;
}
```

A19.25 Distortion Filters

In addition to the ClassA and ClassB emulation objects, I've also prepared a set of distortion filters for you to use. These are based in part on the filters we discussed in Part IV. Since this is really a filtering topic that has been covered in great detail in the FX book, I won't rehash all of that stuff again. These filters are fundamentally simple and easy to use and of course, you can use them in other plugins as well.

A19.25.1 C++ DSP Object: *BigMuffToneControl*

The *BigMuffToneControl* implements the LPF/HPF pair as discussed in Section 19.21.3 and shown in the Block diagram in Figure A19.76. It only has one user-control that adjusts the relative mixture of the filter outputs.

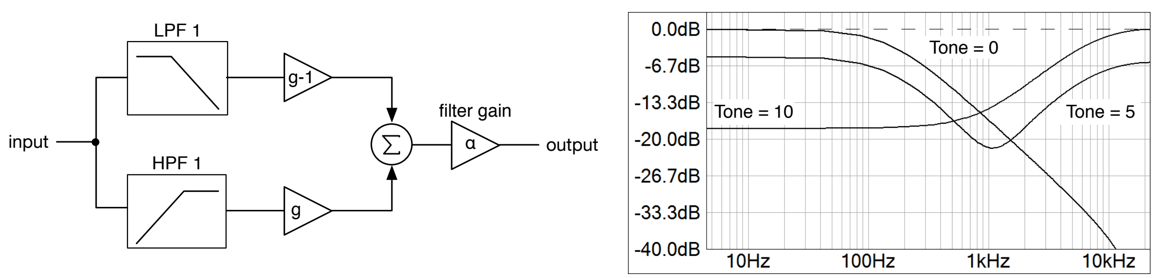


Figure A19.76: the block diagram and frequency response plots for the *BigMuffToneControl* object

A19.25.2 *BigMuffToneControl*: Enumerations and Data Structure

The object is updated via the *BigMuffToneControlParameters* custom data structure that contains members for adjusting its gain control. The custom data structure is simple with only the tone and gain controls needing updates. Notice the labeling of the tone variable with **_010** appended to the name. This indicates that this control is labeled from 0 to 10 on the GUI and will require some kind of re-mapping internally.

```
struct BigMuffToneControlParameters
{
	BigMuffToneControlParameters() {}

	BigMuffToneControlParameters& operator=() // removed

	// --- filter stuff
	double toneControl_010 = 5.0; // tone, 1->10
	double filterGain_dB = 0.0;
};
```

A19.25.3 *BigMuffToneControl*: Members

Tables A19.6 and A19.7 list the *BigMuffToneControl* member variables and member functions. As an aggregate object it is fairly simple.

BigMuffToneControl Member Variables		
Type	Name	Description
BigMuffToneControlParameters	*parameters*	Custom data structure
double	*sampleRate*	Current sample rate
double	*filterGain*	Cooked filter gain variable
double	*filterBlend*	Cooked filter blend variable
AudioFilter	*lpf*	The LPF part of the tone control
AudioFilter	*hsf*	The HSF part of the tone control

Table A19.6: the *BigMuffToneControl* member variables

BigMuffToneControl Member Functions		
Returns	Name	Description
BigMuffToneControlParameters	*GetParameters*	Get all parameters at once
void	*SetParameters* Parameters: - *BigMuffToneControl Parameters* _parameters	Set all parameters at once
bool	*Reset* Parameters: - *double* sampleRate	Resets all filters
double	*processAudioSample* - *double* xn	Process input xn through the model

Table A19.7: the *BigMuffToneControl* member functions

A19.25.4 *BigMuffToneControl*: Programming Notes

The *BigMuffToneControl* object is too simple to warrant wasting a lot of space with. It simply runs the two filters in parallel and blends the output, scaled with the one and only tone control. An optional filter output gain is added since the original was passive – you may ignore it if you wish. But the method for setting up the filters is interesting because I am using a high shelf filter to mimic part of the "problem" with the passive design. For my system, the two filters used have the following specifications:

Low Pass Filter:

- f_c = 150 Hz
- DC filter gain = 0.0dB

High Shelf Filter:

- f_c = 3000 Hz
- *shelf boost/cut:* +18dB
- DC filter gain = -18.0dB

Note the way I use the shelf boost value combined with the opposite DC gain value. This places the shelf at the correct amplitude which is key for getting that notch at 1kHz when the control is at 50%.

A19.25.5 C++ DSP Object: *ComplexLPF*

The *ComplexLPF* implements three preset lowpass filters that may also be run in 2nd or 4th order configurations. This object is designed as a post-distortion filter, or as a speaker simulation filter (or both, since the two are often combined). The name Complex LPF comes from the original Scholz Rockman patent and refers to the rippling lowpass filter used in the highest distortion settings. The filter's three presets are name as follows:

- *normal*: filter with f_c = 3.2kHz and Q = 1, this is based off of the GK 250-ML post distortion filter for the lead channel
- *resonant*: stagger tuned LPFs produce ripples and two peaks in the pass band
- *bright*: brighter than normal with emphasis at 2kHz; note that this filter is only subtly more bright sounding than the others so use good monitors when auditioning them

My implementation makes no effort to copy that particular filter, though my "resonant" variation does include the pass band ripping of the Scholz filter. Figure A19.77 (a) shows the block diagram of stagger tuned LPFs. A preset table stores the f_c and Q values which the user may manipulate via presets only. Each filter may also have its 2nd filter bypassed for a total of six (6) different output responses all centered around a f_c of about 3kHz.

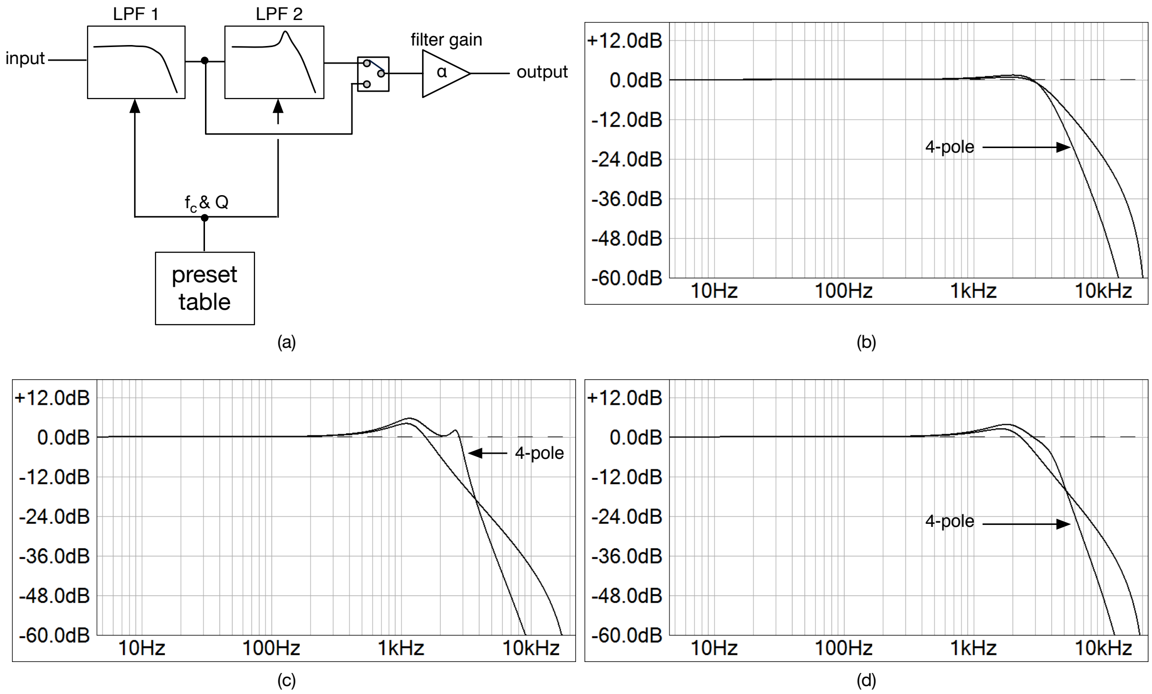


Figure A19.77: the *ComplexLPF* block diagram (a) and frequency response plots for normal (a), resonant (b) and bright (c) presets

A19.25.6 *ComplexLPF*: Enumerations and Data Structure

The object is updated via the *ComplexLPF*P*arameters* custom data structure that contains members for adjusting its gain control. There are two custom, strongly typed enums to handle the preset and filter order selection.

```
enum class complexLPFPreset { resonant, normal, bright };
enum class complexLPFOrder { twoPole, fourPole };
```

The custom data structure is minimal and only contains three members for the preset, order, and final filter gain scalar. The default is for the normal, 4 pole version and if this is all you need, you may use the object without dealing with the parameter updates.

```
struct ComplexLPFParameters
{
	ComplexLPFParameters() {}

	ComplexLPFParameters& operator=()   // removed

	// --- filter stuff
	complexLPFPreset algorithm = complexLPFPreset::normal;
	complexLPFOrder filterOrder = complexLPFOrder::fourPole;
	double filterGain_dB = 0.0;
};
```

A19.25.7 *ComplexLPF*: Members

Tables A19.8 and A19.9 list the *ComplexLPF* member variables and member functions. As an aggregate object it is fairly simple.

ComplexLPF Member Variables		
Type	Name	Description
ComplexLPFParameters	*parameters*	Custom data structure
double	*sampleRate*	Current sample rate
double	*filterGain*	Cooked filter gain variable
AudioFilter	*lowpassFilters[2]*	The two LPF objects

Table A19.8: the *ComplexLPF* member variables

ComplexLPF Member Functions		
Returns	Name	Description
ComplexLPFParameters	*GetParameters*	Get all parameters at once
void	*SetParameters* Parameters: - *ComplexLPFParameters* _parameters	Set all parameters at once
bool	*Reset* Parameters: - *double* sampleRate	Resets all filters
double	*processAudioSample* - *double* xn	Process input xn through the model
void	*updateFilters* - *double* xn	Recalculate filter coefficients due to user preset change

Table A19.9: the *ComplexLPF* member functions

A19.25.8 *ComplexLPF*: Programming Notes

As with the previous object, the *ComplexLPF* object is too simple to warrant a big discussion. It simply runs the two filters in series and scales the output while allowing the user to ignore one of the two filters in the two-pole mode.

A19.25.9 C++ DSP Object: *VariBPF*

One of the most interesting filters from all of the patents might be the Peavey superdistortion sliding bandpass filter. We also observed very similar pre and post-distortion filters in the Scholz Distortion Generation patent. Having a filter whose parameters are set with a single control that is also used to control signal level and/or overdrive/distortion is very powerful and is key to the success of the Peavey patent. By using a pair of these filters, you may setup the same kind of system Scholz teaches in the Distortion Generation patent with morph-able filters that change continuously with the user's selection.

The *VariBPF* object may be used as a morphing band pass filter (mBPF) using the single morph knob control that is labeled 0 to 10. The *VariBPF* is implemented with two filters in series, HPF into LPF, which creates the BPF response. Figure A19.78 (a) shows how the filters are specified. You set starting and ending points for the three parameters of the filter as it morphs from start to end as the control moves from 0 to 10. You may also set the high and low edge slopes as either 6dB/oct or 12dB/oct which allows you to create variable-slope BPFs where one band edge is steeper than the other (if you do this, then the starting slopes must match; i.e. the filter can not change its order while morphing).

In Figure A19.78 (a) you can see that the filters are set up to morph from high frequencies to low frequencies, while smoothing increasing the gain. There are no big rules here – you may set the filters up to morph from high to low or low to high in any of the parameters. But, the big rule is that for any BPF, the f_L value must be lower than the f_H value otherwise it won't be a band pass filter. You may also set up the filters to have a constant gain, or constant common break frequency.

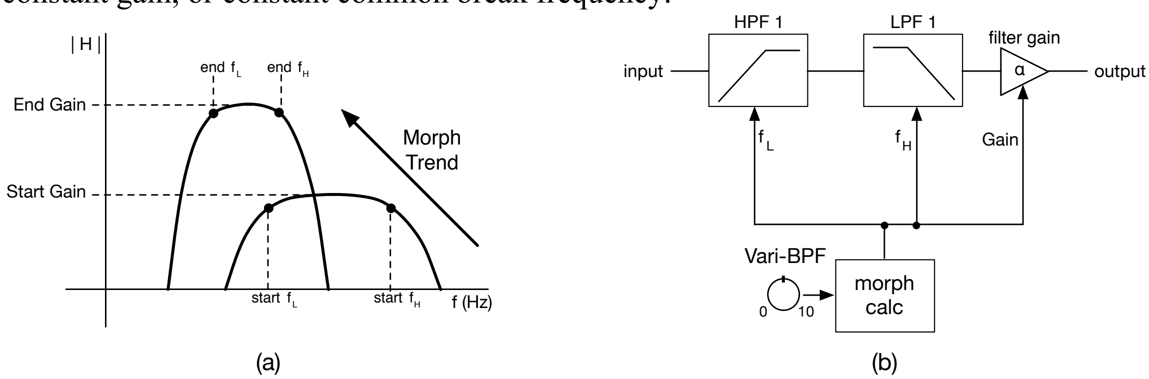


Figure A19.78 (a) the morphing filter specifications and (b) block diagram

Figure A19.78 (b) shows the block diagram which consists of the filters and a morphing calculator that converts the Vari-BFP control's 0 to 10 value into a set of the three filter parameters: gain, f_L and f_H. There is a post filter gain block to move the entire response up or down in amplitude as per the configuration. Figure A19.79 shows two different sets of responses for two different specifications. One thing to note is that in Figure A19.79, the control is linear and the snapshots of the filter responses are taken at even intervals (0, 2, 4, 6, 8, 10). Notice that the gain morphs *linearly in dB* and that the frequencies all morph linearly as well. You can change this behavior with your GUI control selection (linear, log, anti-log) or you can write a simple modification to apply the taper of choice (HINT: see the *PluginParameter* object's tapering code).

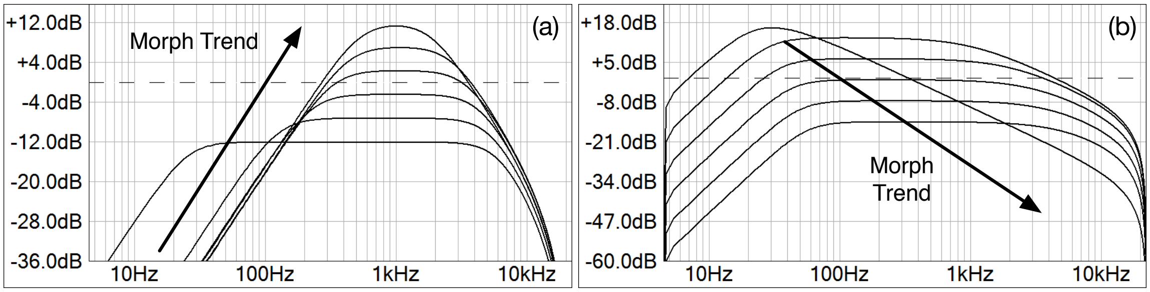


Figure A19.79: two versions of the *VariBPF*; in (a) the filter starts low amplitude and wide bandwidth, then morphs to a narrow, higher center frequency while (b) morphs in the opposite direction

A19.25.10 *VariBPF*: Enumerations and Data Structure

The object is updated via the *VariBPFParameters* custom data structure that contains members for adjusting its morph control, as well as all of the other parameters for the starting and ending filters. However, it is likely that you will want to pre-program several morph curves into the object and allow the user to select from them. Allowing the user to adjust the edge frequencies can be dangerous because they may come up with a combination that is dis-allowed. Adding GUI control code to catch and notify the user would probably be a kludge at best. For the code here, the items in bold are probably the only ones you want to allow the user to modify.

```
struct VariBPFParameters
{
	VariBPFParameters() {}

	VariBPFParameters& operator=()		// removed

	// --- allows customization of BPF slopes
	variBPFEdgeSlope lfEdgeSlope = variBPFEdgeSlope::twelvedB_oct;
	variBPFEdgeSlope hfEdgeSlope = variBPFEdgeSlope::twelvedB_oct;

	// --- single control
	double variBPFControl_010 = 5.0;

	// --- filter spec: start and stop low/high band edges
	double start_FL_Hz = 50.0;
	double end_FL_Hz = 500.0;
	double start_FH_Hz = 5000.0;
	double end_FH_Hz = 2000.0;

	// --- start and end gains
	double startFilterGain_dB = 0.0;
	double endFilterGain_dB = 0.0;
};
```

A19.25.11 *VariBPF*: Members

Tables A19.10 and A19.11 list the *VariBPF* member variables and member functions. As an aggregate object it is fairly simple.

VariBPF Member Variables		
Type	Name	Description
VariBPFParameters	*parameters*	Custom data structure
double	*sampleRate*	Current sample rate
double	*currentFilterGain*	Cooked filter gain variable
double	*current_FL*	Cooked filter FL gain variable
double	*current_FH*	Cooked filter FH gain variable
AudioFilter	*lpf*	The LPF object
AudioFilter	*hpf*	The HPF object

Table A19.10: the *VariBPF* member variables

VariBPF Member Functions		
Returns	Name	Description
VariBPFParameters	*GetParameters*	Get all parameters at once
void	*SetParameters* Parameters: - *VariBPFParameters* _parameters	Set all parameters at once
bool	*Reset* Parameters: - *double* sampleRate	Resets all filters
double	*processAudioSample* - *double* xn	Process input xn through the model

Table A19.11: the *VariBPF* member functions

A19.25.12 *VariBPF*: Programming Notes

As with the previous object, the *VariBPF* object is too simple to warrant a big discussion. It uses the two filters in series to create the bandpass edge. The only interesting code is in the *setParameters* method where the morphed parameters are updated. Notice the use of the *doUnipolartModulationFromMin* function, which is described in more detail in Chapter 13 of the FX book. The second part of the function chooses the filter order based on band edge selection.

```
void setParameters(const VariBPFParameters& params)
{
	parameters = params;

	// --- this maps variBPF control = 0 -> 10   to   G = 0 -> 1
	double variBlend = (parameters.variBPFControl_010) / (10.0);

	// --- find modulated gain in dB
	double modFilterGain_dB = doUnipolarModulationFromMin(
					variBlend,
					parameters.startFilterGain_dB,
					parameters.endFilterGain_dB);

	currentFilterGain = pow(10.0, modFilterGain_dB / 20.0);

	// --- current modulated values
	current_FL = doUnipolarModulationFromMin(
					variBlend,
					parameters.start_FL_Hz,
					parameters.end_FL_Hz);

	current_FH = doUnipolarModulationFromMin(
					variBlend,
					parameters.start_FH_Hz,
					parameters.end_FH_Hz);

	// --- LPF (upper edge)
	AudioFilterParameters filterParams = lpf.getParameters();
	filterParams.fc = current_FH;

	if (parameters.hfEdgeSlope == variBPFEdgeSlope::sixdB_oct)
		filterParams.algorithm = filterAlgorithm::kLPF1;
	else
		filterParams.algorithm = filterAlgorithm::kLPF2;

	lpf.setParameters(filterParams);

	// --- HPF (lower edge)
	filterParams = hpf.getParameters();
	filterParams.fc = current_FL;

	if (parameters.lfEdgeSlope == variBPFEdgeSlope::sixdB_oct)
		filterParams.algorithm = filterAlgorithm::kHPF1;
	else
		filterParams.algorithm = filterAlgorithm::kHPF2;

	hpf.setParameters(filterParams);
}
```

The processAudioSample method is super simple – it just processes the filters in series.

```
virtual double processAudioSample(double xn)
{
	double hpOut = hpf.processAudioSample(xn);
	double yn = lpf.processAudioSample(hpOut);
	return currentFilterGain * yn;
}
```

A19.25.13 *VariBPF* as Speaker Simulator

The *VariBPF* object may be modified for use as a loudspeaker simulator. The loudspeaker and enclosure combination produces a frequency response that can be modeled with a resonant band pass filter. In this case, you allow the *Q* of one or both edges to take on a value greater than 0.707. This will produce two resonant humps. On the low frequency band edge, the resonant hump can simulate the resonance that the enclosure helps to cause. On the high frequency edge, the resonant hump can simulate a resonance within the speaker itself. For a great website full of loudspeaker simulators employed in analog guitar amps, see: http://www.hexeguitar.com/diy/techinfo/cabsims

A19.26 C++ DSP Object: *ToneStack*

The last filter object to discuss is the *ToneStack* object that I designed straight from the analog Tone Stack design in the Texas Instruments App Note called *Tone Stack for Guitar Amplifier Reference Design* that is included with this document. Most of the details of operation were already discussed and Figures A19.64 and A19.65 show the basic block diagram, tables used during design, and the filter outputs at various settings of the controls. Figure A19.80 shows the complete block diagram once more for reference. You can see that the filter is an amalgam of sub-filters. The contour filter itself is also an amalgam consisting of a bandpass and high pass filter in parallel. The interface for the user consists of only three knob controls: *bass*, *mid* and *treble*. Optionally, you may allow the user to set the contour filter: *none, normal,* and *mid_boost*. The mid-boost control is sometimes done with a pull-switch potentiometer and labeled “Pull Fat” or “Pull Thick.”

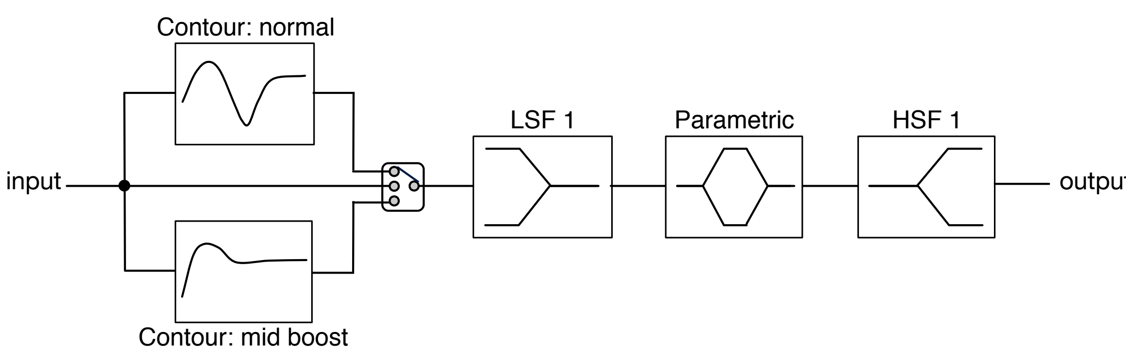


Figure A19.80: the *ToneStack* C++ object in block diagram form

A19.26.1 *ToneStack*: Enumerations and Data Structure

The object is updated via the *ToneStackParameters* custom data structure that contains members for adjusting its three band boost-cut controls in a [0, 10] ranged value, and a contour enumeration that describes the contour filter type.

```
enum class contourType { none, normal, mid_boost };
```

In the custom data structure, notice the naming convention once again with the _010 to signify that this control is labeled and transmits values on the range of [0, +10].

```
struct ToneStackParameters
{
      ToneStackParameters() {}
      ToneStackParameters& operator=()    // removed

      // --- filter stuff
      contourType contour = contourType::normal;
      double LFToneControl_010 = 5.0; // tone, 0->10
      double HFToneControl_010 = 5.0; // tone, 0->10
      double MFToneControl_010 = 5.0; // tone, 0->10
};
```

A19.26.2 *ToneStack*: Members

Tables A19.12 and A19.13 list the *ToneStack* member variables and member functions. As an aggregate object it is fairly simple.

ToneStack Member Variables		
Type	Name	Description
ToneStackParameters	*parameters*	Custom data structure
double	*sampleRate*	Current sample rate
double	*currentFilterGain*	Cooked filter gain variable
double	*current_FL*	Cooked filter FL gain variable
double	*current_FH*	Cooked filter FH gain variable
AudioFilter	*lpf*	The LPF object
AudioFilter	*hpf*	The HPF object

Table A19.12: the *ToneStack* member variables

ToneStack Member Functions		
Returns	Name	Description
ToneStackParameters	*GetParameters*	Get all parameters at once
void	*SetParameters* Parameters: - *ToneStackParameters* _parameters	Set all parameters at once
bool	*Reset* Parameters: - *double* sampleRate	Resets all filters
double	*processAudioSample* - *double* xn	Process input xn through the model

Table A19.13: the *ToneStack* member functions

A19.26.3 *ToneStack*: Programming Notes

The *ToneStack* is really an alaborate combination of sub-filters and is nearly completely specified in the TI App Note. I created the contour filter by taking measurements directly from the plot shown in Figure A19.63. The three-band filter consists of three series filters: low shelf, constant Q mid parametric, and high shelf. All of these filters and their specifications have already been documented in the FX book so I won't bother repeating it here.

The contour filter is designed with two filters in parallel having the following specifications:

Bandpass Filter:

- $f_c = 50.0$ Hz
- $Q = 0.222$
- *Gain* = +3.5 dB

High pass Filter (normal):

- $f_c = 750.0$ Hz
- *Gain* = +2.0 dB

High pass Filter (mid boost):

- $f_c = 250.0$ Hz
- *Gain* = +2.0 dB

The *processAudioFrame* function does most of the work running the signal through the desired filters. Notice the code at the top for selecting the contour filters.

```
virtual double processAudioSample(double xn)
{
      double cn = xn;
      double ynCFB = contourBPF.processAudioSample(cn);
      double ynCFH = contourHPF.processAudioSample(cn);

      if (parameters.contour != contourType::none)
      {
            // --- preset gains
            //
            // --- BPF: fc = 50Hz Q = 0.222 Gain = +3.5dB
            // --- HPF: fc = 750Hz            Gain = +2.0dB
            //
            //       double contourBPFGain = pow(10.0, 3.5 / 20.0);
            //       double contourHPFGain = pow(10.0, 2.0 / 20.0);;
            cn = contourBPFGain * ynCFB + contourHPFGain * ynCFH;
      }

      double ynLP = lowShelfFilter.processAudioSample(cn);
      double ynHP = highShelfFilter.processAudioSample(ynLP);
      double ynMB = midParametricFilter.processAudioSample(ynHP);
      return ynMB;
}
```

The *setParameters* method translates the knob controls' [0, 10] range to the proper range required for the filters.

```
void setParameters(const ToneStackParameters& params)
{
	parameters = params;

	// --- contour presets
	AudioFilterParameters filterParams = contourHPF.getParameters();

	if(parameters.contour == contourType::normal)
		filterParams.fc = 750.0;
	else if (parameters.contour == contourType::mid_boost)
		filterParams.fc = 250.0;

	// --- this will reject same settings...
	contourHPF.setParameters(filterParams);

	// --- this maps qControl = 1 -> 10   to   K = 0 -> 1
	double LF_gain = parameters.LFToneControl_010  / 10.0;
	double MF_gain = parameters.MFToneControl_010 / 10.0;
	double HF_gain = parameters.HFToneControl_010 / 10.0;

	// --- convert to dB
	double LF_gain_dB = doUnipolarModulationFromMin(LF_gain,
							-12.0, +12.0);
	double MF_gain_dB = doUnipolarModulationFromMin(MF_gain,
							-6.0, +6.0);
	double HF_gain_dB = doUnipolarModulationFromMin(HF_gain,
							-8.0, +8.0);

	// --- LSF
	filterParams = lowShelfFilter.getParameters();
	filterParams.boostCut_dB = LF_gain_dB;
	lowShelfFilter.setParameters(filterParams);

	// --- HSF
	filterParams = highShelfFilter.getParameters();
	filterParams.boostCut_dB = HF_gain_dB;
	highShelfFilter.setParameters(filterParams);

	// --- Para
	filterParams = midParametricFilter.getParameters();
	filterParams.boostCut_dB = MF_gain_dB;
	midParametricFilter.setParameters(filterParams);
}
```

A19.27 C++ Effect Object: *TurboDistorto*

The *TurboDistorto* plugin object emulates a simple but highly saturated distortion box. The word "turbo" comes from the turbo switch that you use to engage a second waveshaper that adds more distortion. The *TurboDistorto* block diagram is shown in Figure A19.81. The plugin has three knobs: *Distortion*, *Tone* and *Volume* and two switches: *Turbo* and *Filter*. The *Distortion, Tone,* and *Volume* controls use the guitarist friendly range of [0, 10] and the two switches may be implemented with Boolean variables. An optional fourth control is available if needed called *Trim* to make an adjustment on the input level from -20dB to +20dB. This optional control may be omitted as the *Trim* value defaults to 0dB.

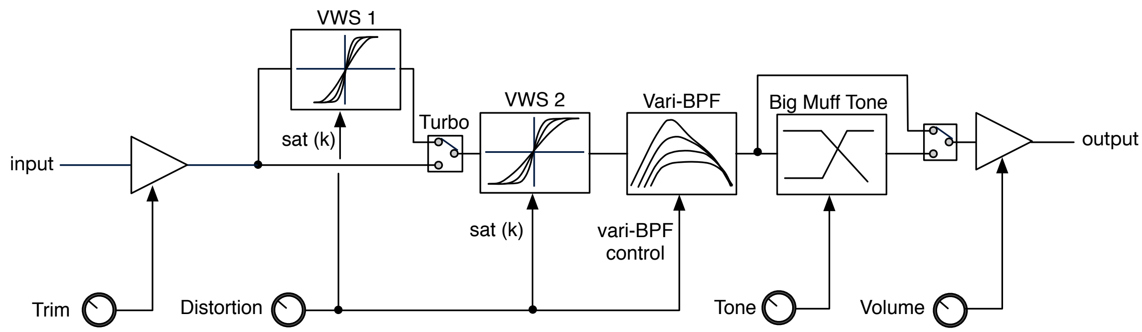


Figure A19.81: the *TurboDistorto* block diagram

The distortion mechanism is a variable waveshaper (VWS) whose saturation (k) value is taken from the user control marked "Distortion." If the *Turbo* button is engaged, a second waveshaper pre-distorts the signal first. The pre-distortion VWS uses a saturation value that is ¼ that of the normal value that is sent to the normal waveshaper, but it still varies with the control knob. In this way, you can create cascading distortion of increasing intensity. The equations for the two variable wave shapers are:

$$\begin{array}{ll} VWS\ 1 & VWS\ 2 \\ m = k/4 & \\ y = \dfrac{\tanh(mx)}{\tanh(m)} & y = \dfrac{\tanh(kx)}{\tanh(k)} \end{array}$$

The output of the waveshapers is fed to the *VariBPF* object with the following specifications:

- Start f_L: 150Hz
- Start f_H: 5.8kHz
- Start gain: -10 dB

- End f_L: 500Hz
- End f_H: 2.5kHz
- End gain: -2.0 dB

- Low Edge slope: 6 dB/oct
- High Edge slope: 12dB/oct

The *Distortion* knob connects directly to the *VariBPF* control on the object. Therefore, as you turn up the distortion knob, you increase the gain through the waveshapers and simultaneously set the distortion post-filter which is a BPF that gets narrower as the control is increased. The curves I use are similar but not identical to the Peavey superdistortion patent circuit.

The output of the *VariBPF* is then fed into the simple Big Muff tone control for more tonal shaping. The user can optionally bypass this tone control with a switch. The output is then processed through a final simple gain stage to set the output volume.

A19.27.1 *TurboDistorto*: Enumerations and Data Structure

The object is updated via the *TurboDistortoParameters* custom data structure that contains members for adjusting its four controls in a [0, 10] ranged value, and two Booleans that handle the filter engagement and turbo boost mode. In the custom data structure, notice the naming convention once again with the _010 to signify that this control is labeled and transmits values on the range of [0, +10].

```
struct TurboDistortoParameters
{
	TurboDistortoParameters () {}
	TurboDistortoParameters & operator=()     // removed

	// --- distortion controls
	double distortionControl_010 = 0.0;
	double toneControl_010 = 0.0;
	double outputGain_010 = 5.0;

	// --- input volume tweaker
	double inputGainTweak_010 = 5.0; // [-20, +20] dB

	// --- switches
	bool engageTurbo = false;
	bool bypassFilter = false;
 };
```

A19.27.2 *TurboDistorto*: Members

Tables A19.14 and A19.15 list the *TurboDistorto* member variables and member functions. As an aggregate object it is fairly simple.

TurboDistorto Member Variables		
Type	Name	Description
TurboDistortoParameters	*parameters*	Custom data structure
double	*sampleRate*	Current sample rate
double	*inputGain*	Cooked filter gain variable for input tweak
double	*outputGain*	Cooked filter gain

		variable for output
double	*current_FH*	Cooked filter FH gain variable
VariBPF	*variBPF*	The variable BPF object
BigMuffToneControl	*toneFilter*	The filter object

Table A19.14: the *TurboDistorto* member variables

TurboDistorto Member Functions		
Returns	Name	Description
TurboDistortoParameters	*GetParameters*	Get all parameters at once
void	*SetParameters* Parameters: - *TurboDistortoParameters* _parameters	Set all parameters at once
bool	*Reset* Parameters: - *double* sampleRate	Resets all filters
double	*processAudioSample* - *double* xn	Process input xn through the model

Table A19.15: the *TurboDistorto* member functions

A19.27.3 *TurboDistorto*: Programming Notes

The *TurboDistorto* is an amalgam of existing robust C++ objects and its operation is straightforward. One interesting function you will find in *valves.h* is a range mapper. This function takes an input control value over an input range, and produces an output control value on the mapped output range. This function is used to easily convert the [0, 10] knob control ranges to arbitrary ranges such as [-20, +20] for the input tweaker or [-40, +20] for the output gain. There is also a flag that will process the mapped control value as dB and return the cooked value that your plugin will use.

```
inline double calcMappedVariableOnRange(double inLow, double inHigh,
                                        double outLow, double outHigh,
                                        double control,
                                        bool convertFromDB = false)
{
      // --- mapper
      double slope = (outHigh - outLow) / (inHigh - inLow);
      double yn = outLow + slope * (control - inLow);

      if (convertFromDB)
            return pow(10.0, yn / 20.0);
      else
            return yn;
}
```

The *processAudioFrame* function does most of the work running the signal through the desired filters. Notice the code at the top for selecting the contour filters.

```
virtual double processAudioSample(double xn)
{
      // --- convert distortion knob to waveshaper value from 1 -> 50
      double saturation = calcMappedVariableOnRange(0.0, 10.0,
                              1.0, 50.0,
                              parameters.distortionControl_010);

      double yn = inputGain * xn;

      // --- turbo adds pre-distortion
      if (parameters.engageTurbo)
            yn = tanh((saturation / 4.0)*yn) / tanh(saturation / 4.0);

      // --- normal shaper
      yn = tanh(saturation*yn) / tanh(saturation);

      // --- VariBPF
      yn = variBPF.processAudioSample(yn);

      // --- optional filter
      if (!parameters.bypassFilter)
            yn = toneFilter.processAudioSample(yn);

      return outputGain*yn;
}
```

The *setParameters* method translates the knob controls' [0, 10] range to the proper range required for the filters.

```
void setParameters(const TurboDistortoParameters& params)
{
      parameters = params;

      // --- find the output gain amount
      outputGain = calcMappedVariableOnRange(0.0, 10.0,
                                    40.0, +20.0,
                                    parameters.outputGain_010,
                                    true);

      // --- find the input tweak amount
      inputGain = calcMappedVariableOnRange(0.0, 10.0,
                              20.0, +20.0,
                              parameters.inputGainTweak_010,
                              true);

      // --- update the VariBPF
      VariBPFParameters vbpfParams = variBPF.getParameters();

      vbpfParams.variBPFControl_010 = parameters.distortionControl_010;
      variBPF.setParameters(vbpfParams);

      // --- update the Big Muff Tone
      BigMuffToneControlParameters bmParams =
```

```
                                toneFilter.getParameters();

      bmParams.toneControl_010 = parameters.toneControl_010;
      toneFilter.setParameters(bmParams);
}
```

A19.28 Chapter Plugin: *SuperSaturator*

Our *SuperSaturator* plugin object will implement the block diagram in Figure A19.82. It is very simple because there only a few controls and the *TurboDistorto* object does all the work.

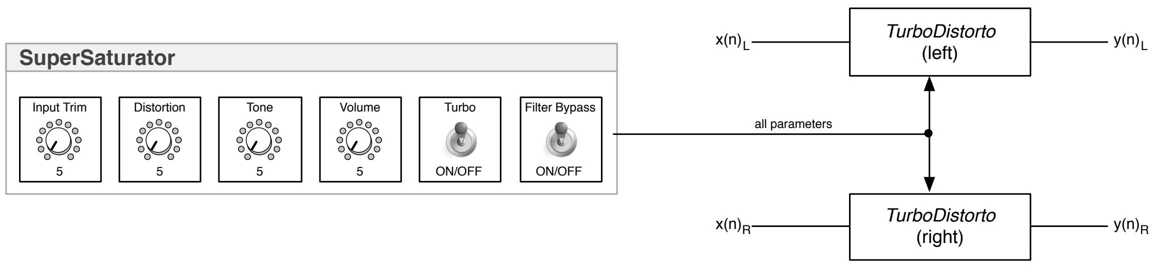


Figure A19.82: the GUI and block diagram for the *SuperSaturator* plugin

A19.28.1 *SuperSaturator* GUI Parameters

The GUI parameter table is shown in Table A19.16 – use it to declare your GUI parameter interface for the plugin framework you are using. If you are using ASPIK, remember to first create your enumeration of control ID values at the top of the *plugincore.h* file and use automatic variable binding to connect the linked variables to the GUI parameters.

ASPiK: top of *plugincore.h* file:

```
enum controlID {
      inputTrim_010 = 0,
      distortionControl_010 = 1,
      toneControl_010 = 2,
      volumeControl_010 = 3,
      engageTurbo = 4,
      filterBypass = 5
};
```

Control Name	Units	min/max/def or string-list	Taper	Linked Variable	Linked Variable type
Input Trim		-40 / 0 / -10	linear	*inputTrim_010*	*double*
Distortion		0 / 20 / 5	linear	*distortionControl_010*	*double*
Tone		1 / 20 / 1	linear	*toneControl_010*	*double*
Volume		1 / 100 / 5	linear	*volumeControl_010*	*double*
Turbo	-	"off, on"	-	*engageTurbo*	*int*
Filter Bypass	-	"off, on"	-	*filterBypass*	*int*

Table A19.16 the *SuperSaturator* plugin's GUI parameter and linked-variable list

A19.28.2 *SuperSaturator* Object Declarations and Reset

The *SuperSaturator* plugin uses two *TurboDistorto* objects to implement the processing. We use a simple array of two, one for each of the left and right channels. Of course you would need to declare more members to support more channels. In your plugin's processor (*PluginCore* for ASPiK) declare two members plus the *udpateParameters* method:

```
// --- in your processor object's .h file
#include "fxobjects.h"

// --- in the class definition
protected:
      TurboDistorto distortion[2];
      void updateParameters();
```

You only need to reset the objects with the current sampling rate – it will reset its sub-components and so on.

ASPiK: *reset( )*

```
// --- reset left and right processors
distortion[0].reset(resetInfo.sampleRate);
distortion[1].reset(resetInfo.sampleRate);
```

A19.28.3 *SuperSaturator* GUI Parameter Update

The *TurboDistorto* object requires only six parameters assuming you include the input trim control. The object does the translational mapping from [0, 10] so the update is simple – just pass the control values directly.

ASPiK: *updateParameters( )* function:

```
// --- get params
TurboDistortoParameters turboParams = distortion[0].getParameters();

// --- switches
turboParams.bypassFilter = filterBypass == 1;
```

```
turboParams.engageTurbo = engageTurbo == 1;

// --- 0 -> 10 controls
turboParams.distortionControl_010 = distortionControl_010;
turboParams.toneControl_010 = toneControl_010;
turboParams.outputGain_010 = volumeControl_010;
turboParams.inputGainTweak_010 = inputTrim_010;

// --- do the update BAM
distortion[0].setParameters(turboParams);
distortion[1].setParameters(turboParams);
```

A19.28.4 *SuperSaturator* Process Audio

The *TurboDistorto* objects do the work. All we need to do is send and receive the audio data and we're done.

ASPiK: *processAudioFrame( )* function:

```
// --- read input (adapt to your framework)
double xnL = processFrameInfo.audioInputFrame[0]; //< input sample L
double xnR = processFrameInfo.audioInputFrame[1]; //< input sample R

// --- process Left and Right channels
double ynL = distortion[0].processAudioSample(xnL);
double ynR = distortion [1].processAudioSample(xnR);

// --- write output (adapt to your framework)
processFrameInfo.audioOutputFrame[0] = ynL; //< output sample L
processFrameInfo.audioOutputFrame[1] = ynR; //< output sample R
```

A19.29 C++ Effect Object: *OneMarkAmp*

The *OneMarkAmp* object is a complete tube amp channel. You can combine two of them together with some effects to make a pretty cool plugin. This fundamentally models an early Mesa Boogie amplifier. It has plenty of gain and crunch. You can easily modify it for very high gain sounds by adding one or two more triode preamp tube stages. The block diagram is shown in Figure A19.83.

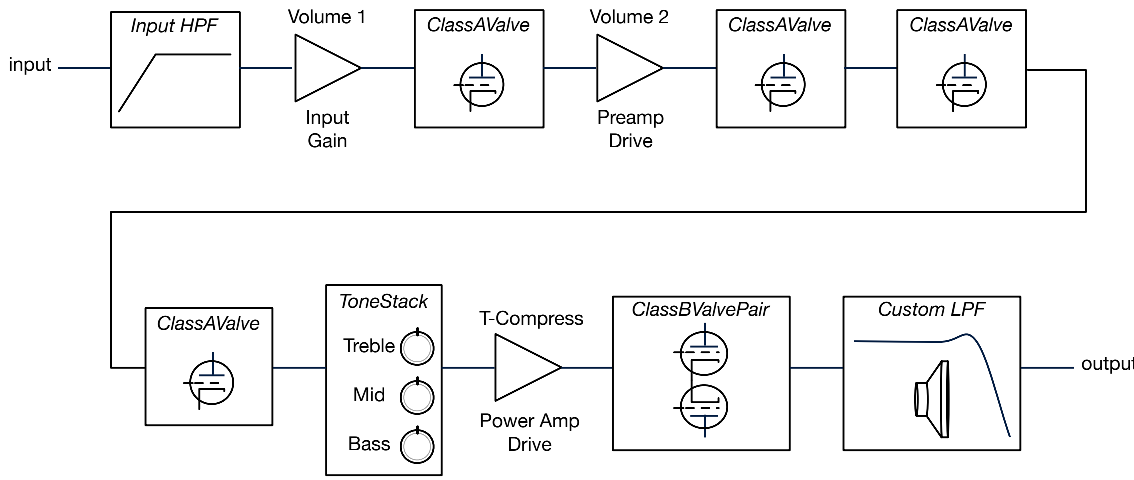


Figure A19.83: block diagram for the *OneMarkAmp* tube guitar amplifier emulator

The *OneMarkAmp* emulator has the following features and modules:

- Input HPF, user adjustable for eliminating DC offsets as well as removing some bass (see Scholz patents) prior to distortion
- Two "volume" knobs: the first is an input attenuator/amplifier with up to +20dB of gain while the second is the preamp drive control with another +/-20dB of gain
- Tone stack with selectable contour
- Four triodes in the preamp
- One Class-B pentode pair
- Class-B tube compression emulation with adjustable overdrive/compression gain
- Custom designed output LPF for speaker emulation

The amp emulator works with three primary gain and distortion controls. The first control marked "Volume 1" adjusts the input level into the first triode model. It provides up to +20dB of gain. The idea here is to drive the input tube just to the point of grid-conduction, but not past it (of course you can do this if you like). At a full-scale input, this triode will output a harmonically enhanced, but not clipped waveform when driven right up to this point.

The "Volume 2" control amplifies or attenuates the first triode's output by +/-20 dB and drives the rest of the chain of triode models. This is the first thing you can modify – the location of the preamp drive control. These two gain controls are how you achieve the preamp distortion. You might also want to increase the top end of the control to +30 or even +40 dB. We are using the headroom that floating-point operation allows, so the internal signal may become very large. A fixed output attenuator brings everything back to the normal operating range (it is omitted from the block diagram).

The "T-Compress" control adjusts the level into the Class-B emulator with the Pirkle model. This control adds +/-12 dB of gain allowing you to go from no tube compression to quite a bit of it (which you can very easily adjust in the model). You will notice that as you turn the control past "5" the output level will increase (get louder) but the amplitude of the signal won't change if you monitor the output signal with a meter – this is the Class-B tube compression effect. NOTE: if you want to experiment with the Poletti model, you will need to play around with the gain levels into and out of the algorithm, including adjustments of the waveshaper limit (L_x) values as they are very different than my version.

The tone stack is identical to the one presented previously without alterations. One simple modification is to change the boost/cut maximum limits and increase them. The output LPF was custom designed just for this preamp. It is a second order LPF with f_c = 2.45 kHz and Q = 1.4 with an optional "bright" setting that raises the f_c to 3.2 kHz and sets the Q at 0.707. There is plenty of room to experiment here from designing your own speaker model to convolving with an impulse response of an actual guitar speaker, which is sort of the norm today.

A19.29.1 DC Shift Monitoring

The DC shifting in the triode preamp tube emulators is reported back to the owning object via the custom data structure. You can monitor the DC shifts this way and provide optional indicators for the user to allow them to know which tubes are being overloaded at their inputs. I designed the preamp so that the last three tubes each overdrive the next in sequence in a successively higher manner, which changes the DC shifting point of each tube; this adds harmonic complexity to the signal as its harmonics come and go depending on the amplitude of the input. It kind of reminds me of how harmonics change in FM synthesis – they undulate rather than simply being filtered to a specific harmonic envelope.

I setup the output gain of each tube to overdrive the next into grid-conduction in a smooth and pleasing way when the first input triode is driven up-to, but not in-to, grid-conduction. So, the way to get the most saturated sound with the most pleasing harmonic quality (to my ears at least) is to increase the Volume 1 control until the DC shift value just starts to become non-zero into the very first triode. For an LED monitor, you could light it when this happens to let the user know to back off the Volume 1 a bit. After that is set, you can crank up the Volume 2 knob all the way to 10 and achieve DC shifting and grid-conduction distortion in the last three triodes, with the last one in the chain receiving the highest shifts and generating the most harmonic distortion. If you follow my example, then you can add one or two (or three or four) more triode stages to get heavier, higher gain sounds of today's more modern amplifiers. The Class-B output stage can also report the DC shifts in each of its two branches but I have found that this is much simpler to set by ear and visual indicators are not necessary.

A19.29.2 Triode Preamp Parameters

To develop this plugin, I used a special "distortion kit" plugin I designed several years ago that allows me to combine distortions, filters, and gain structures together. I can "solo" certain structures to hear them individually and I can tweak all parameters of the models I am working with. The triode preamp parameters were a result of observation of the DC-2 preamp details from Section A19.21 in combination with a lot of time spent tweaking out the models. Tube amp emulation requires a lot of tweaking by ear, more than just about any other algorithm except perhaps some reverb designs. The parameters here are my tweaks and of course you will want to start modifying them – but at least you will have a firm starting point.

Parameter	Triode 1	Triode 2	Triode 3	Triode 4
f_H	20 kHz	9 kHz	7 kHz	6.4 kHz
f_L	8 Hz	32 Hz	40 Hz	43 Hz
Shelving filter	-10dB @ 10 Hz	-10dB @ 10 Hz	-10dB @ 10 Hz	-10dB @ 10 Hz
Output Gain	-3 dB	+5 dB	+6 dB	-20 dB
DC Shift Coeff.	1.0	0.20	0.50	0.52

Table A19.17: individual *ClassAValve* object settings for each emulated triode

A19.29.3 Class-B Power Amp Parameters

The Class-B emulation using my algorithm only requires one parameter to be set called *inputGain*. This value will set the maximum gain into the device for a full-scale signal and sets how far the tube model will go into crossover distortion compression. The higher the value, the more easily the amp model is overdriven. If the value is too high, then the power amp drive control will have little to no effect so there is definitely a sweet setting area that lets you move from clean and undistorted to over-biased and crossover compression. In addition, the signal entering the Class-B model is already amplified somewhat, so this input gain value must also take that into account.

The values I have found to be best are between +20 dB and +30 dB. The output gain must be reduced to get the signal back down into the proper [-1.0, +1.0] range so an output attenuator that inverts the gain value will work well. For the power amp drive (called "T-compress") the parameters are:

- *inputGain* = +20 dB = 10
- *outputGain* = -20 dB = 0.1

These values will give a modest level of gain and Class-B compression. If you want to experiment with higher settings on the input gain, go for it. It just gets crunchier!

A19.29.4 Input HPF

The input HPF is a simple 2nd order HPF with Q = 0.707. The range of fc values is:

[5.0 Hz, 1kHz]

You may change this range in the *setParameters* function that is discussed in a few sections ahead.

A19.29.5 Output/Post-Distortion LPF

The post-distortion filter, or speaker simulator, is a very simple 2nd order LPF and you could certainly improve on it by increasing the roll-off slope to 4th order. Figure A19.84 shows the two responses. The normal response has f_c = 2.45 kHz and Q = 1.4. The "bright" setting f_c to 3.2 kHz and sets the Q at 0.707 and this makes an enormous difference in the sound of the upper harmonics.

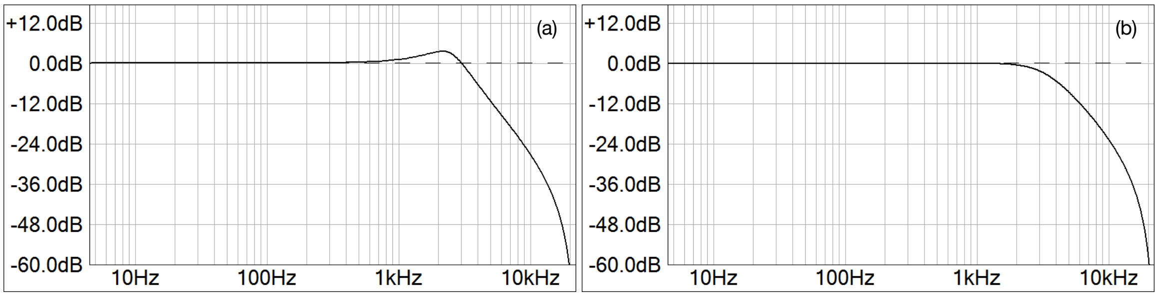


Figure A19.84: (a) the normal and (b) the "bright" output LPF responses

A19.29.6 *OneMarkAmp*: Enumerations and Data Structure

The object is updated via the *OneMarkAmpParameters* custom data structure that contains members for adjusting its controls in a [0, 10] ranged value. Since the *ToneStack* object already has its own custom data structure, we can just make one of those structures a member of the amp parameters. In addition there is a *bright* switch for the output LPF, and there are four DC shift return variables in an array that indicate how much negative DC bias shifting has occurred in the tube model.

The waveshaper saturation value will be used to set the simulated amp's tubes – low, medium and high gain are simulated with saturation = 1.0, 2.0 and 5.0 respectively. This will give the amp model a wide range of tones. It should be noted that all voicing development was done with the saturation at 1.0 (low).

```
enum class ampGainStructure { low, medium, high };

struct OneMarkAmpParameters
{
	OneMarkAmpParameters() {}

	OneMarkAmpParameters& operator=()   // removed

	// --- amp controls are all 0 -> 10
	double volume1_010 = 0.0;
	double volume2_010 = 0.0;
	double inputHPF_010 = 0.0;
	double tubeCompression_010 = 0.0;
	double masterVolume_010 = 0.0;

	// --- tone stack can take care of itself
	ToneStackParameters toneStackParameters;
```

```
    // --- for monitoring preamp tube DC shifts
    double dcShift[PREAMP_TRIODES] = { 0.0, 0.0, 0.0, 0.0 };

    // --- switches
    bool bright = false;
    ampGainStructure ampGainStyle = ampGainStructure::medium;
};
```

A19.29.7 *OneMarkAmp*: Members

Tables A19.18 and A19.19 list the *OneMarkAmp* member variables and member functions. As an aggregate object it is fairly simple.

OneMarkAmp Member Variables		
Type	Name	Description
OneMarkAmpParameters	*parameters*	Custom data structure
double	*sampleRate*	Current sample rate
double	*inputGain*	Raw input gain into first triode model
double	*driveGain*	Raw input gain between 1st and 2nd triode models
double	*tubeCompress*	Amount of gain feeding the Class-B emulator
double	*outputGain*	The final scaled output of the plugin form the master volume control
ClassAValve	*triodes[4]*	The four triode models
ClassBValvePair	*outputPentodes*	The power amp model
ToneStack	*toneStack*	The tone stack
AudioFilter	*inputHPF*	Input DC blocker and bass-reduction
AudioFilter	*outputLPF*	Post-distortion filter

Table A19.18: the *OneMarkAmp* member variables

OneMarkAmp Member Functions		
Returns	Name	Description
OneMarkAmpParameters	*GetParameters*	Get all parameters at once
void	*SetParameters* Parameters: - *OneMarkAmpParameters*	Set all parameters at once

	_parameters	
bool	*Reset* Parameters: - *double* sampleRate	Resets all sub objects
double	*processAudioSample* - *double* xn	Process input xn through the model

Table A19.19: the *OneMarkAmp* member functions

A19.29.8 *OneMarkAmp*: Programming Notes

The *OneMarkAmp* combines many modules together. The triode preamp models are all set using Table A19.17.

A19.29.8.1 reset

The *reset* function is where the models are all initialized. The code is straightforward and doesn't need much commenting. Notice that the pentode gain parameters are calculated from dB values – I left this calculation in place so that you may experiment with how crunchy or smooth you want the Class-B amp emulator to sound. Also, the fourth triode's output gain is set to -20dB to make up for all of the previous gain staging, however this signal is still amplified. You can also experiment with this value to increase or decrease the default drive into the Class-B model.

```
virtual bool reset(double _sampleRate)
{
      // --- do any other per-audio-run inits here
      sampleRate = _sampleRate;

      triodes[T1].reset(sampleRate);
      ClassAValveParameters triodeParams = triodes[T1].getParameters();
      triodeParams.lowFrequencyShelf_Hz = 10.0;
      triodeParams.lowFrequencyShelfGain_dB = -10.0;
      triodeParams.integratorFc = 1.0;
      triodeParams.highCompress = false;
      triodeParams.millerHF_Hz = 20000.0;
      triodeParams.dcBlockingLF_Hz = 8.0;
      triodeParams.outputGain = pow(10.0, -3.0 / 20.0);
      triodeParams.dcShiftCoefficient = 1.0;
      triodes[T1].setParameters(triodeParams);

      triodes[T2].reset(sampleRate);
      triodeParams = triodes[T2].getParameters();
      triodeParams.lowFrequencyShelf_Hz = 10.0;
      triodeParams.lowFrequencyShelfGain_dB = -10.0;
      triodeParams.integratorFc = 1.0;
      triodeParams.highCompress = false;
      triodeParams.millerHF_Hz = 9000.0;
      triodeParams.dcBlockingLF_Hz = 32.0;
      triodeParams.outputGain = pow(10.0, +5.0 / 20.0);
      triodeParams.dcShiftCoefficient = 0.20;
      triodes[T2].setParameters(triodeParams);

      triodes[T3].reset(sampleRate);
```

```
	triodeParams = triodes[T3].getParameters();
	triodeParams.lowFrequencyShelf_Hz = 10.0;
	triodeParams.lowFrequencyShelfGain_dB = -10.0;
	triodeParams.integratorFc = 1.0;
	triodeParams.highCompress = false;
	triodeParams.millerHF_Hz = 7000.0;
	triodeParams.dcBlockingLF_Hz = 40.0;
	triodeParams.outputGain = pow(10.0, +6.0 / 20.0);
	triodeParams.dcShiftCoefficient = 0.50;
	triodes[T3].setParameters(triodeParams);

	triodes[T4].reset(sampleRate);
	triodeParams = triodes[T4].getParameters();
	triodeParams.lowFrequencyShelf_Hz = 10.0;
	triodeParams.lowFrequencyShelfGain_dB = -10.0;
	triodeParams.integratorFc = 1.0;
	triodeParams.highCompress = false;
	triodeParams.millerHF_Hz = 6400.0;
	triodeParams.dcBlockingLF_Hz = 43.0;
	triodeParams.outputGain = pow(10.0, -20.0 / 20.0);
	triodeParams.dcShiftCoefficient = 0.52;
	triodes[T4].setParameters(triodeParams);

	outputPentodes.reset(sampleRate);
	ClassBValveParameters pentodeParams =
	                              outputPentodes.getParameters();

	pentodeParams.algorithm = classBType::pirkle;
	pentodeParams.inputGain = pow(10.0, 20.0 / 20.0);
	pentodeParams.outputGain = pow(10.0, -20.0 / 20.0);
	outputPentodes.setParameters(pentodeParams);

	toneStack.reset(sampleRate);

	inputHPF.reset(sampleRate);
	AudioFilterParameters hpfParams = inputHPF.getParameters();
	hpfParams.algorithm = filterAlgorithm::kHPF2;
	inputHPF.setParameters(hpfParams);

	outputLPF.reset(sampleRate);
	AudioFilterParameters lpfParams = outputLPF.getParameters();
	lpfParams.algorithm = filterAlgorithm::kLPF2;
	outputLPF.setParameters(lpfParams);

	return true;
}
```

A19.29.8.2 setParameters

This function is important because it maps the non-tone-stack controls [0, 10] ranges into their proper values. If you want to modify how those controls operate, this is the place to do it.

```
void setParameters(const OneMarkAmpParameters& params)
{
	parameters = params;
```

```
// --- simulate gain structures with waveshaper saturation
double saturation = 1.0;
if (parameters.ampGainStyle == ampGainStructure::medium)
      saturation = 2.0;
else if (parameters.ampGainStyle == ampGainStructure::high)
      saturation = 5.0;

// --- update
for (int i = 0; i < PREAMP_TRIODES; i++)
{
      ClassAValveParameters triodeParams =
                              triodes[i].getParameters();
      triodeParams.waveshaperSaturation = saturation;
      triodes[i].setParameters(triodeParams);
}

// --- input gain
if (parameters.volume1_010 == 0.0)
      inputGain = 0.0;
else
{
      inputGain = calcMappedVariableOnRange(0.0, 10.0,
            -60.0, +20.0,
            parameters.volume1_010,
            true);
}

// --- drive gain
driveGain = calcMappedVariableOnRange(0.0, 10.0,
            -20.0, +20.0,
            parameters.volume2_010,
            true);

// --- HPF
AudioFilterParameters hpfParams = inputHPF.getParameters();
hpfParams.fc = calcMappedVariableOnRange(0.0, 10.0,
      5.0, 1000.0,
      parameters.inputHPF_010);
inputHPF.setParameters(hpfParams);

// --- ToneStack
toneStack.setParameters(parameters.toneStackParameters);

// --- output gain amount
outputGain = calcMappedVariableOnRange(0.0, 10.0,
      -60.0, +12.0,
      parameters.masterVolume_010,
      true);

tubeComress = calcMappedVariableOnRange(0.0, 10.0,
      -6.0, +24.0,
      parameters.tubeCompression_010,
      true);

// --- speaker simulator
AudioFilterParameters lpfParams = outputLPF.getParameters();
lpfParams.fc = parameters.bright ? 3200.0 : 2450.0;
lpfParams.Q = parameters.bright ? 0.707 : 1.4;
```

```
	outputLPF.setParameters(lpfParams);
}
```

A19.29.8.3 processAudioSample

The processing function ties everything together and pushes the signal from input to output. Here is where you can re-wire stuff to your own liking. For example, it is simple to re-locate the tone stack, but it can have a huge affect on the sound of the amp modeler. You can also re-arrange the gain staging, but you will need to do some signal metering to make sure your DC shifts are acceptable. Also, notice that the first triode meter is binary – if it goes above 0.125 then the meter will fully light up. That is because of how I set the gain staging and you might want to disable that code if/when you change the structuring.

```
virtual double processAudioSample(double xn)
{
	// --- remove DC, remove bass
	double hpfOut = inputHPF.processAudioSample(xn);

	// --- "volume 1" control
	double t1In = hpfOut * inputGain;

	// --- first triode
	double t1Out = triodes[0].processAudioSample(t1In);

	// --- add pre-drive
	t1Out *= driveGain;

	// --- cascade of triodes
	// leaving this verbose - experiment, use less or more triodes...
	double t2Out = triodes[1].processAudioSample(t1Out);
	double t3Out = triodes[2].processAudioSample(t2Out);
	double t4Out = triodes[3].processAudioSample(t3Out);

	// --- class B drive gain
	t4Out *= tubeComress;

	// --- class B model
	double classBOut = outputPentodes.processAudioSample(t4Out);

	// --- tone stack (note: relocating makes a big difference)
	double toneStackOut = toneStack.processAudioSample(classBOut);

	// --- speaker sim
	double outputLPFout = outputLPF.processAudioSample(dcBlock);

	// --- for metering the DC Shifts only - can ignore if you want
	for(int i=0; i< PREAMP_TRIODES; i++)
		parameters.dcShift[i] =
				triodes[i].getParameters().dcOffsetDetected;

	// --- the FIRST meter is binary
	if (parameters.dcShift[0] > 0.125)
		parameters.dcShift[0] = 1.0;

	// --- final output
	return outputLPFout * outputGain;
}
```

A19.30 C++ Effect Object: *WickerAmp*

The *WickerAmp* is a C++ effect object that is a simple combination of two *OneMarkAmp* objects for the tube processing (left and right) and one *ReverbTank* object for the reverb processing. The *ReverbTank* plugin from the FX book comes with two presets, one for a medium room and the other for a hall. We can create a great stereo reverb for the *WickerAmp* by using the *Hall* setting on the *ReverbTank*. All I did was cut and paste the preset values into the plugin's initialization code.

A19.30.1 *WickerAmp*: Enumerations and Data Structure

The object is updated via the *WickerAmpParameters* custom data structure that contains members for adjusting its controls in a [0, 10] ranged value. The large majority of work is already done in the *OneMarkAmp* object so we simply declare a member structure that will cover the majority of the GUI controls. We only need two additional parameters, onf for the single reverb control and the other as a return variable that indicates the DC shift in the first triode, as previously discussed.

```
struct WickerAmpParameters
{
	WickerAmpParameters() {}

	WickerAmpParameters& operator=()    // removed

	// --- only need one since both channels are identical
	OneMarkAmpParameters ampParameters;

	// --- our reverb
	double masterReverb_010 = 0.0;

	// --- DC shift warning meter
	double dcShiftTriode_0 = 0.0;
};
```

A19.30.2 *WickerAmp*: Members

Tables A19.20 and A19.21 list the *WickerAmp* member variables and member functions. As an aggregate object it is fairly simple.

WickerAmp Member Variables		
Type	Name	Description
WickerAmpParameters	*parameters*	Custom data structure
double	*sampleRate*	Current sample rate
OneMarkAmp	*tubeAmpChannel[2]*	A pair of tube amp channels
ReverbTank	*reverb*	The reverb module

Table A19.20: the *WickerAmp* member variables

WickerAmp Member Functions		
Returns	Name	Description
*WickerAmp*Parameters	*GetParameters*	Get all parameters at once
void	*SetParameters* Parameters: - *WickerAmpParameters* _parameters	Set all parameters at once
bool	*Reset* Parameters: - *double* sampleRate	Resets all sub objects
double	*processAudioSample* - *double* xn	Process input xn through the model

Table A19.21: the *WickerAmp* member functions

A19.30.3 *WickerAmp*: Programming Notes

The *WickerAmp* is really simple because it just chains together the two tube channels and the reverb processing. It forwards most of the work to the member-objects.

A19.30.3.1 reset

The *reset* function is where the models are all initialized. The main thing we do here is to initialize the reverb unit with the Hall settings from that plugin project.

```
virtual bool reset(double _sampleRate)
{
      // --- do any other per-audio-run inits here
      sampleRate = _sampleRate;

      tubeAmpChannel[0].reset(sampleRate);
      tubeAmpChannel[1].reset(sampleRate);

      reverb.reset(sampleRate);

      ReverbTankParameters verbParams = reverb.getParameters();
      verbParams.kRT = 0.901248;
      verbParams.lpf_g = 0.300000;
      verbParams.lowShelf_fc = 150.000000;
      verbParams.lowShelfBoostCut_dB = -20.000000;
      verbParams.highShelf_fc = 4000.000000;
      verbParams.highShelfBoostCut_dB = -6.000000;
      verbParams.wetLevel_dB = -12.000000;
      verbParams.dryLevel_dB = 0.000000;
      verbParams.apfDelayMax_mSec = 33.000000;
      verbParams.apfDelayWeight_Pct = 85.000000;
      verbParams.fixeDelayMax_mSec = 81.000000;
      verbParams.fixeDelayWeight_Pct = 100.000000;
      verbParams.preDelayTime_mSec = 25.000000;
      verbParams.density = reverbDensity::kThick;
      reverb.setParameters(verbParams);
```

```
	return true;
}
```

A19.30.3.2 setParameters

Here we just forward the amp parameters to the member objects. We only have one control to deal with for converting the reverb controls [0, 10] range into the wet mix value in dB.

```
void setParameters(const WickerAmpParameters& params)
{
	parameters = params;

	tubeAmpChannel[0].setParameters(parameters.ampParameters);
	tubeAmpChannel[1].setParameters(parameters.ampParameters);

	double reverbWet_dB = parameters.masterReverb_010 == 0.0 ?
	                  -96.0 : calcMappedVariableOnRange(0.0, 10.0,
	                                                    -60.0, 0.0,
	                                       parameters.masterReverb_010);

	ReverbTankParameters verbParams = reverb.getParameters();
	verbParams.wetLevel_dB = reverbWet_dB;
	reverb.setParameters(verbParams);
}
```

A19.30.3.3 processAudioSample & processAudioFrame

All of the work is done in *processAudioFrame*. Function calls to *processAudioSample* are forwarded to the frame processing function. This function then routes the audio through the tube channels, then processes it through the reverb unit. Notice how it takes two DC shift metering values from the channels and picks the larger of the two to display – this makes sense since there is only one input control knob for both channels.

```
// --- do the valve emulation
virtual double processAudioSample(double xn)
{
	float input[2] = { xn, xn };
	float output[2] = { 0.0, 0.0 };
	processAudioFrame(input, output, 1, 1);
	return output[0];
}

/** process STEREO audio amp in frames */
virtual bool processAudioFrame(const float* inputFrame,
	float* outputFrame,
	uint32_t inputChannels,
	uint32_t outputChannels)
{
	// --- make sure we have input and outputs
	if (inputChannels == 0 || outputChannels == 0)
		return false;

	// --- pick up inputs
```

```
    //
    // --- LEFT channel - know we must have one
    double xnL = inputFrame[0];
    double ynL = tubeAmpChannel[0].processAudioSample(xnL);

    OneMarkAmpParameters ampParams =
                    tubeAmpChannel[0].getParameters();
    double leftDCMeter = ampParams.dcShift[0];

    // --- RIGHT channel (duplicate left input if mono-in)
    double xnR = inputChannels > 1 ? inputFrame[1] : xnL;

    // --- right
    double ynR = tubeAmpChannel[1].processAudioSample(xnR);

    ampParams = tubeAmpChannel[1].getParameters();
    double rightDCMeter = ampParams.dcShift[0];

    // --- LED meter
    parameters.dcShiftTriode_0 = fmax(leftDCMeter, rightDCMeter);

    float reverbInput[2] = { ynL , ynR };
    float reverbOutput[2] = { 0.0 , 0.0 };

    reverb.processAudioFrame(reverbInput, reverbOutput, 2, 2);

    // --- set left channel
    outputFrame[0] = reverbOutput[0];

    // --- set right channel
    outputFrame[1] = reverbOutput[1];

    return true;
}
```

A19.31 Chapter Plugin Project: *WickerComboAmp*

The *WickerAmpCombo* is a combo amp simulator that uses a single *WickerAmp* object for the audio channel processing. Although the guitar amp is meant for mono signals, the plugin will actually have two identical channels that will feed the stereo reverb unit. In this way, it can be used with stereo sources and samples.

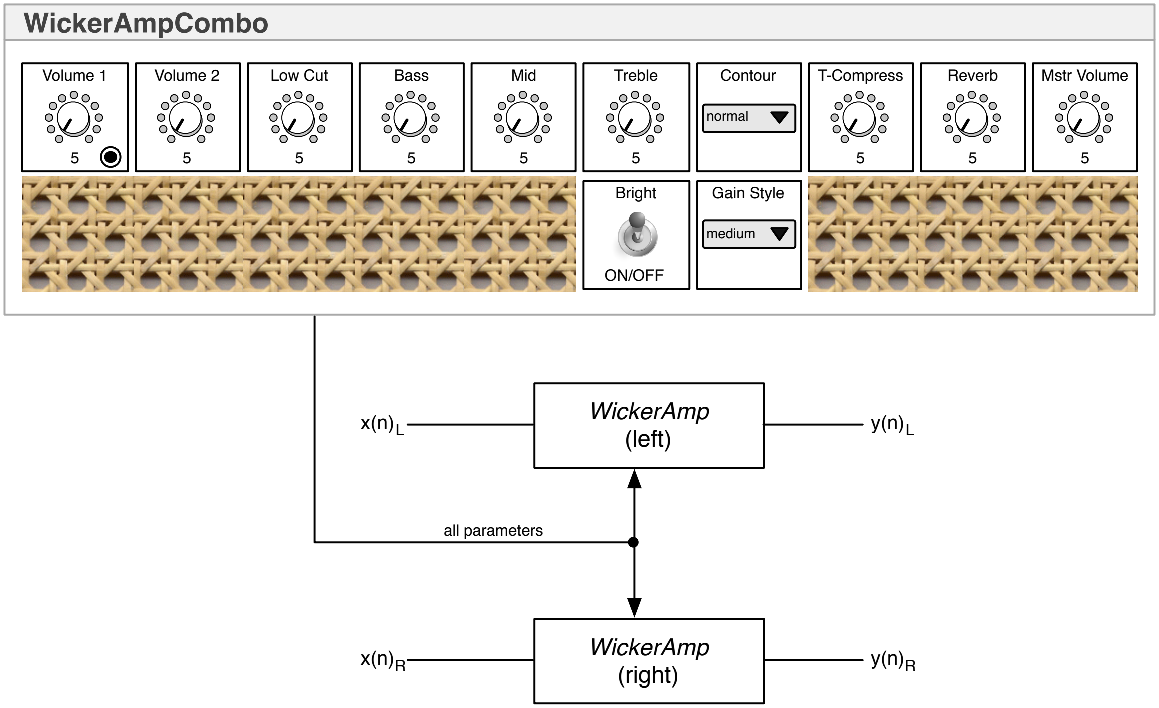


Figure A19.85: the *WickerAmpCombo* plugin GUI and connection; notice the LED meter on the lower right of the Volume 1 knob' this will light when the first triode goes into grid-conduction on either channel

A19.31.1 *WickerAmpCombo* GUI Parameters

The GUI parameter table is shown in Table A19.22 – use it to declare your GUI parameter interface for the plugin framework you are using. If you are using ASPIK, remember to first create your enumeration of control ID values at the top of the *plugincore.h* file and use automatic variable binding to connect the linked variables to the GUI parameters.

ASPiK: top of *plugincore.h* file:

```
enum controlID {
      volume1_010 = 0,
      volume2_010 = 1,
      lowCut_010 = 2,
      bassControl_010 = 3,
      midControl_010 = 4,
      trebleControl_010 = 5,
      toneStackContour = 6,
      masterVolume_010 = 9,
      tubeCompress_010 = 7,
      enableBright = 8,
      gainStructure = 17,
      masterReverb_010 = 19,
      triode1_DCMeter = 10
};
```

Control Name	Units	min/max/def or string-list	Taper	Linked Variable	Linked Variable type
Volume 1		0/ 10 / 5	linear	*volume1_010*	*double*
Volume 2		0/ 10 / 5	linear	*volume2_010*	*double*
Low Cut		0/ 10 / 2	linear	*lowCut_010*	*double*
Bass		0/ 10 / 5	linear	*bassControl_010*	*double*
Mid		0/ 10 / 5	linear	*midControl_010*	*double*
Treble		0/ 10 / 5	linear	*trebleControl_010*	*double*
T-Compress		0/ 10 / 5	linear	*tubeCompress_010*	*double*
Master Reverb		0/ 10 / 0	linear	*masterReverb_101*	*double*
Master Volume		0/ 10 / 5	linear	*masterVolume_101*	*double*
Bright	-	"off, on"	-	*enableBright*	*int*
Contour	-	"none, normal, mid_boost"	-	*toneStackContour*	*int*
Gain	-	"low, medium, high"	-	*gainStructure*	*int*

Table A19.22 the *WickerAmp* plugin's GUI parameter and linked-variable list

A19.31.2 *WickerAmpCombo* Object Declarations and Reset

The *WickerAmpCombo* plugin uses one *WickerAmp* object that does all the work so the plugin is just a wrapper.

```
// --- in your processor object's .h file
#include "valves.h"

// --- in the class definition
protected:
      WickerAmp wickerAmp;
      void updateParameters();
```

You only need to reset the objects with the current sampling rate – it will reset its sub-components and so on.

ASPiK: *reset( )*

```
// --- reset left and right processors
wickerAmp.reset(resetInfo.sampleRate);
```

A19.31.3 *WickerAmpCombo* GUI Parameter Update

The *WickerAmpCombo* object's custom data structure has several member structures, one of which has more member structures. You only need to get the proper GUI control information into the correct structure slot. Most of these are simple forwards with a few string list variables as well that require the conversion function.

ASPiK: *updateParameters( )* function:

```
WickerAmpParameters params = wickerAmp.getParameters();

params.ampParameters.volume1_010 = volume1_010;
params.ampParameters.volume2_010 = volume2_010;
params.ampParameters.inputHPF_010 = lowCut_010;
params.ampParameters.toneStackParameters.LFToneControl_010 =
                                                bassControl_010;
params.ampParameters.toneStackParameters.MFToneControl_010 =
                                                midControl_010;
params.ampParameters.toneStackParameters.LFToneControl_010 =
                                                bassControl_010;

params.ampParameters.toneStackParameters.contour =
                  convertIntToEnum(toneStackContour, contourType);

params.ampParameters.bright = enableBright == 1;

params.ampParameters.tubeCompression_010 = tubeCompress_010;
params.ampParameters.masterVolume_010 = masterVolume_010;

params.ampParameters.ampGainStyle = convertIntToEnum(gainStructure,
                                          ampGainStructure);
params.masterReverb_010 = masterReverb_010;

wickerAmp.setParameters(params);
```

A19.31.4 *WickerAmpCombo* Process Audio

The *WickerAmp* object does the work. All we need to do is send and receive the audio data and we're done. For

ASPiK: *processAudioFrame( )* function:

```
bool success =
      wickerAmp.processAudioFrame(processFrameInfo.audioInputFrame,
                                  processFrameInfo.audioOutputFrame,
                                  processFrameInfo.numAudioOutChannels,
                                  processFrameInfo.numAudioOutChannels);

WickerAmpParameters ampParams = wickerAmp.getParameters();
triode1_DCMeter = ampParams.dcShiftTriode_0;
```

Non-ASPiK:
You gather your frame input and output samples into small arrays (you can statically declare them to minimize cost). You use the same code above to pick up the DC shift indicator, if you are supporting it with your framework.

```
double xnL = //< get from your framework: input sample L
double xnR = //< get from your framework: input sample R

float inputs[2] = { xnL, xnR };
float outputs[2] = { 0.0, 0.0 };
```

```
// --- process (in, out, 2 inputs, 2 outputs)
wickerAmp.processAudioFrame(inputs, outputs, 2, 2);

//< framework left output sample = outputs[0];
//< framework Right output sample = outputs[1];
```

A19.32 References/Bibliography (in addition to Patents)

A19.32.1 Books

Hood, J. 1995. Audio Electronics, 2nd ed., Newnes, Inc. Oxford, UK

Hood, J. 1997. Valve and Transistor Audio Amplifiers, Newnes, Inc. Oxford, UK

Jeruchim, M, Balaban, P. and Shanmugan, K. 2002. Simualation of Communication Systems., Kluwer. Newe York

Kyttälä, T. 2008. Solid-State Guitar Amplifiers, self-published, origin unknown

Pirkle, W. 2009. Analog Audio Electronics v3.2, Titone Systems, Inc. Lighthouse Point

Pittman, A. 1995. The Tube Amp Book, 4.1th ed., Groove Tubes, Inc., Sylmar

Self, D. 1995. Self on Audio, Newnes, Inc. Oxford, UK

Self, D. 2015. Small Signal Audio Design, 2nd ed., Newnes, Inc. Oxford, UK

Rozenblit, B. 1997. Beginner's Guide to Tube Audio Design, Amateur Audio Press, Peterborough

Zottola, T. 1995. Vacuum Tube Guitar and Bass Amplifier Theory, Chap. 3-5, Bold Strummer, Inc. Westport

Zottola, T. 1995. Vacuum Tube Guitar and Bass Amplifier Servicing, Chap. 4, Bold Strummer, Inc. Westport

A19.32.2 Websites

- ValveWizard: great site from an established designer; lots of information here! http://www.valvewizard.co.uk/

- HexGuitar: every analog loudspeaker simulation known (well almost) http://www.hexeguitar.com/diy/techinfo/cabsims

- Google patents: https://patents.google.com/

- The AX-84 Project: a DIY tube guitar site that several of my students have used to build their own amplifiers; contains a wealth of information and analysis, including a nice tone stack circuit analysis: http://www.ax84.com/

- The Big Muff π Analysis page: everything you ever wanted to know about every version of the Big Muff π : https://www.electrosmash.com/big-muff-pi-analysis

A19.33 Appendix

The following Appendix contains the patents that I reference in numerous locations.

A19.33.1 Patents:

- Pat: 5,524,055 *Solid State Circuit for Emulating Tube Compression Effect*
- Pat: 5,619,578 *Multistage Solid State Amplifier That Emulates Tube Distortion*
- Pat: 5,841,875 *Digital Audio Signal Processor with Harmonics Modification*
- Pat: 4,811,401 *Superdistorted Amplifier with Normal Gain*
- Pat: 4,584,700 *Electronic Audio Signal Processor*
- Pat: 5,133,015 *Method and Apparatus for Processing an Audio Signal*

A19.33.2 Patent Applications:

- Pat App: US 2008/0049950 *Nonlinear Processor for Audio Signals*

A19.33.3 App Notes:

- Texas Instruments App Note called *Tone Stack for Guitar Amplifier Reference Design*

A19.33.4 Guitar Amps

I've been playing guitar since my father gave me a ¾ size Yamaha steel-string for my birthday when I was in 7th grade. I won't tell you what year that was, but an unknown band out of Boston had just exploded on to the Billboard charts a few months prior. I've been building, repairing, and modifying guitars and amplifiers since just after high school. In college, I used to occasionally buy broken FX pedals from *Ed's Guitars* in Miami, fix them, and sell them back to him – how I wish I had saved a few of them for myself. And, I now regularly bring my amps to the University of Miami where I allow my students to take them apart and analyze them with an oscillator and scope. The amps I (currently) own, and that have made a contribution to this document in some way are:

- 1974 Silverface Princeton Reverb
- 1987 Gallien-Krueger GK 250ML
- 1996 Mesa-Boogie DC-2
- 2008 Mesa-Boogie Lonestar Special ®
- 2019 Peavey TransTube Bandit 112

Ian Williams

TI Designs – Precision: Verified Design
Tone Stack for Guitar Amplifier Reference Design

TEXAS INSTRUMENTS

TI Designs – Precision

TI Designs – Precision are analog solutions created by TI's analog experts. Verified Designs offer the theory, component selection, simulation, complete PCB schematic & layout, bill of materials, and measured performance of useful circuits. Circuit modifications that help to meet alternate design goals are also discussed.

Circuit Description

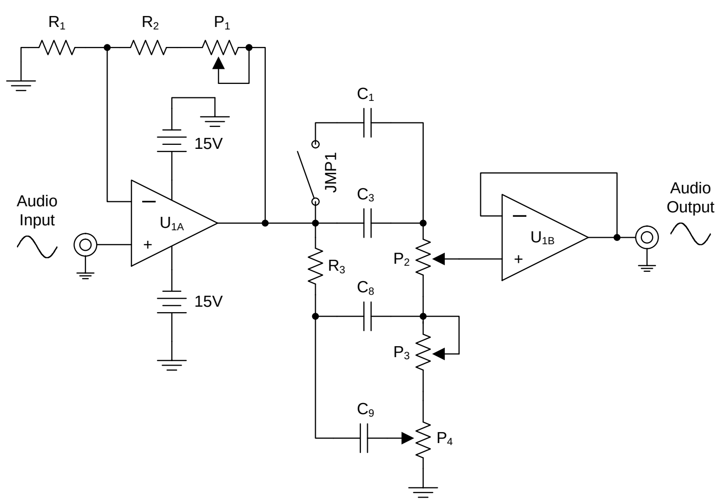
This split-supply, high-performance guitar tone circuit provides control of the bass, mid, and treble frequencies of an electric guitar signal, while also providing gain with minimal distortion and noise. Buffered inputs and outputs preserve the behavior of the system independent of the source and load impedances, and a radio frequency (RF) filter on the circuit front end attenuates noise from outside the audio band.

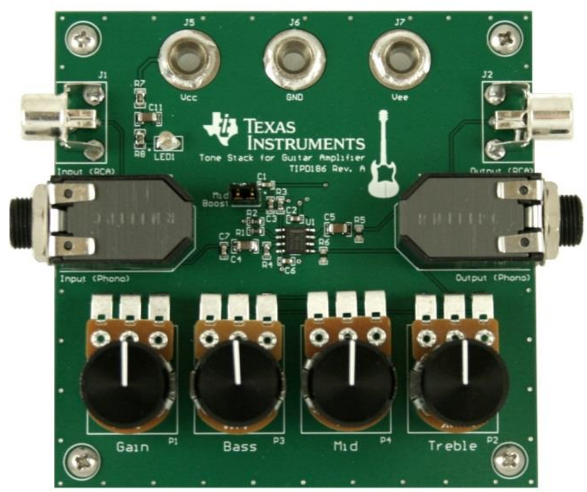
Design Resources

Design Archive	All Design files
TINA-TI™	SPICE Simulator
OPA1642	Product Folder

TI E2E™ Community

Ask the Analog Experts
WEBENCH® Design Center
TI Designs – Precision Library





An IMPORTANT NOTICE at the end of this TI reference design addresses authorized use, intellectual property matters and other important disclaimers and information.

1 Design Summary

The design requirements are as follows:

- Supply voltage: ± 15 V
- Input voltage: 1 V_{RMS}
- Source impedance: 6 kΩ
- Input stage signal gain: 6 dB – 28 dB
- Total harmonic distortion + noise (THD+N) level at 1 kHz: -100 dB (0.001%)
- Treble adjustment range: 10 dB
- Mid adjustment range: 6 dB
- Bass adjustment range: 15 dB

The design goals and performance are summarized in Table 1. Figure 1 depicts the measured transfer function of the design.

Table 1. Comparison of Design Goals, Simulation, and Measured Performance

	Goal	Simulated	Measured
THD+N level at 1 kHz	-100 dB (0.001%)	-102.6 dB (0.00074%)	-105.4 dB (0.00054%)
Treble adjustment range	10 dB	10.4 dB	10.6 dB
Mid adjustment range	6 dB	6.1 dB	8.9 dB
Bass adjustment range	15 dB	18 dB	19.2 dB

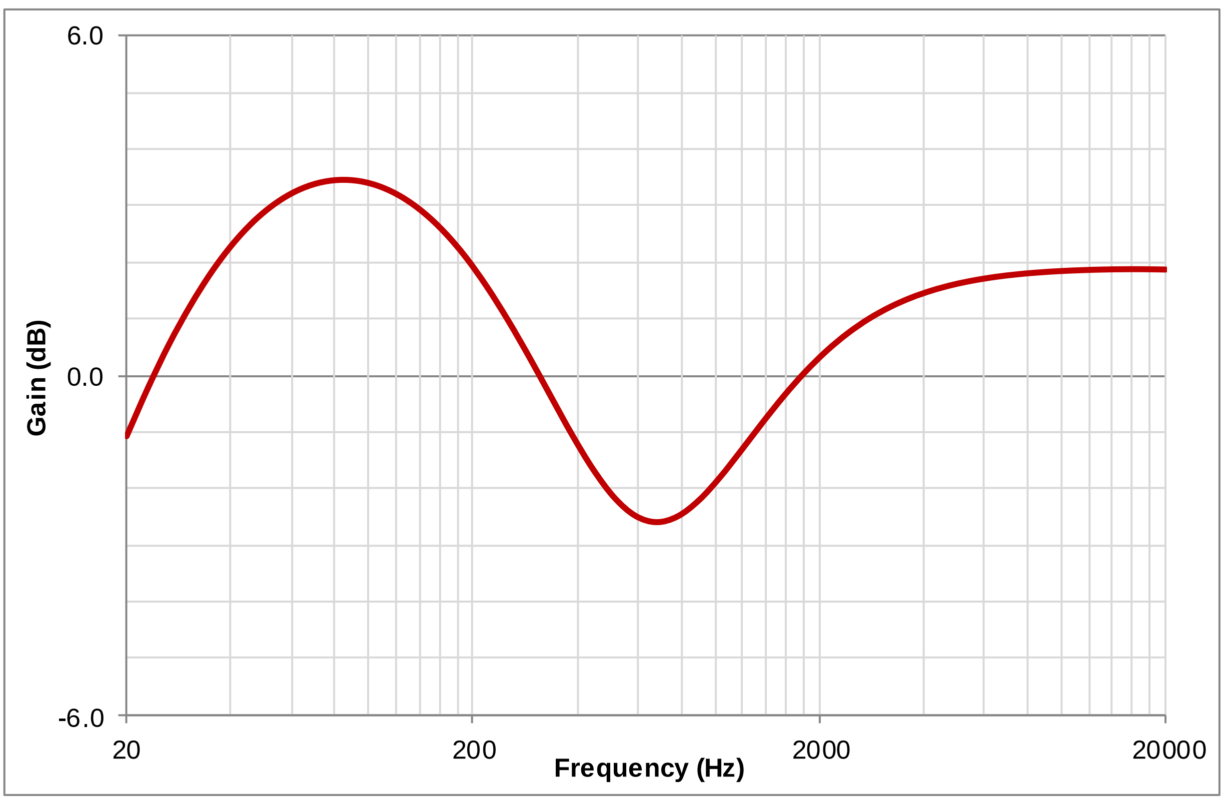


Figure 1: Measured Transfer Function – Bass, Mid, Treble at 50%

2 Theory of Operation

A more complete schematic for this design is shown in Figure 2. The three primary functional blocks of the circuit are the input filter and gain stage, tone stack, and output buffer.

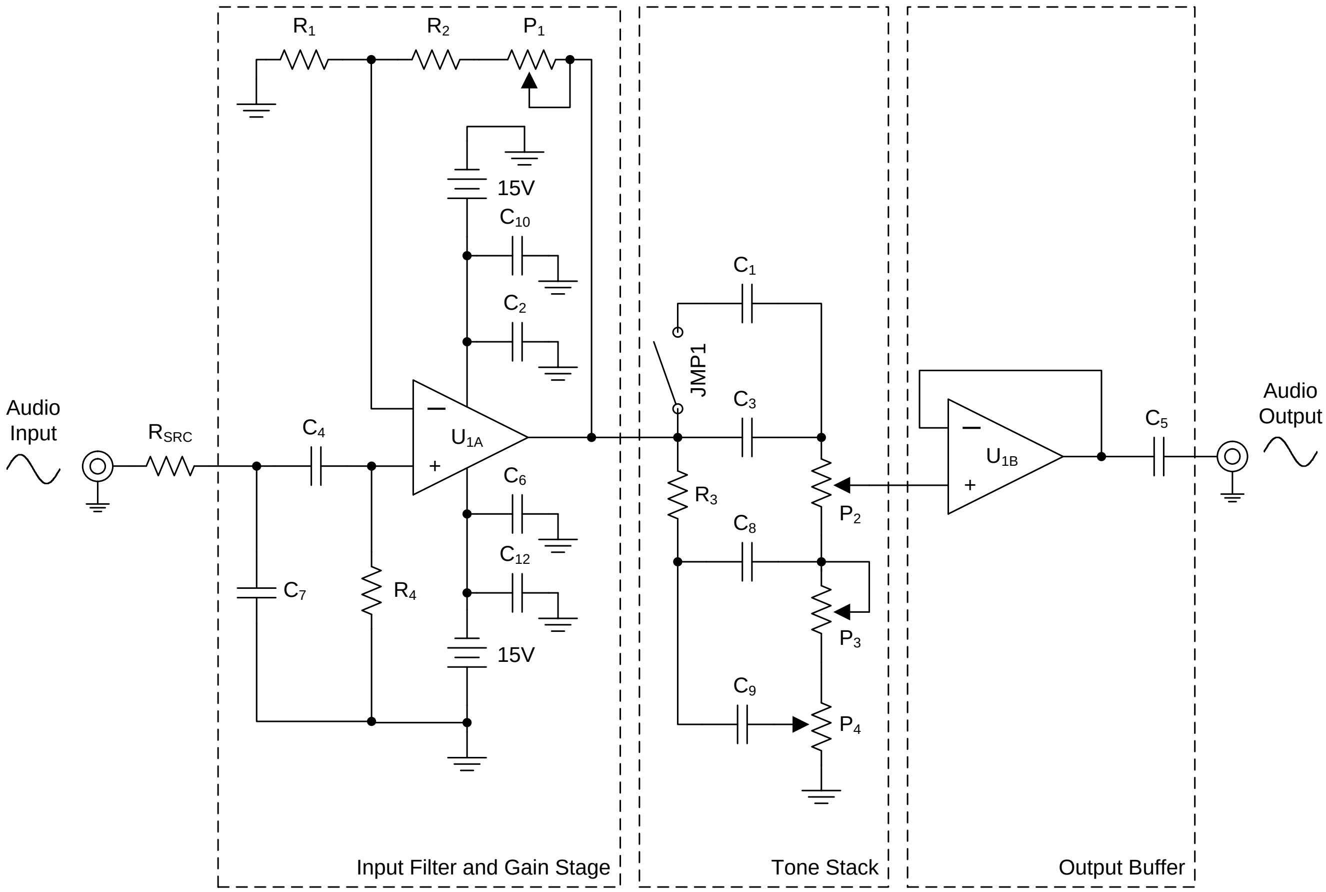


Figure 2: Complete Circuit Schematic

2.1 Input Filter

A passive filter at the input of the circuit serves two purposes: provide significant attenuation at frequencies outside the audio band, and remove any dc voltage from the input signal. The filter is made up of R_{SRC}, the guitar pickup output impedance, capacitors C_7 and C_4, and resistor R_4, as shown in Figure 3.

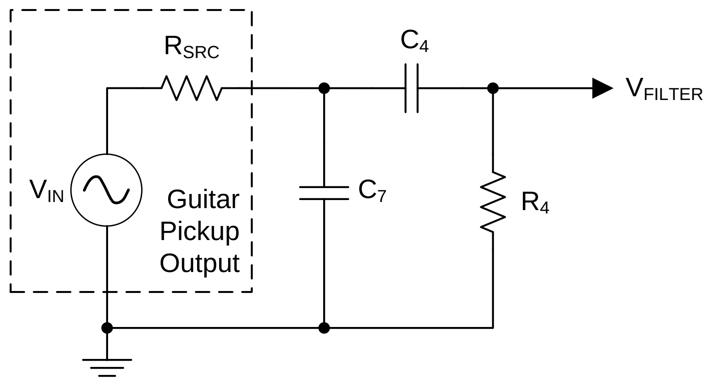


Figure 3: Input Filter Schematic

2.1.1 Low Pass Filter

R_{SRC} and C_7 create a first-order low pass filter. The -3 dB cutoff frequency of the filter is calculated using Equation 1.

$$f_{C\_LPF} = \frac{1}{2\pi * R_{SRC} * C_7} \tag{1}$$

400 kHz is selected as the -3 dB cutoff frequency for the filter. This will effectively attenuate RF noise while preserving the gain and phase behavior at 20 kHz. Since R_{SRC} is specified at 6 kΩ, simply rearrange terms and solve for C_7 in order to achieve the desired cutoff frequency, as shown in Equation 2.

$$C_7 = \frac{1}{2\pi * R_{SRC} * f_{C\_LPF}} = \frac{1}{2\pi * 6k\Omega * 400kHz} = 66.3 \text{ pF} \tag{2}$$

The required value for C_7 is calculated to be 66.3 pF. The nearest standard capacitor value of 68 pF is selected as the actual value. The actual cutoff frequency of the filter is calculated using Equation 3.

$$f_{C\_LPF} = \frac{1}{2\pi * R_{SRC} * C_7} = \frac{1}{2\pi * 6k\Omega * 68pF} = 390 \text{ kHz} \tag{3}$$

2.1.2 High Pass Filter

C_4 and R_4 create a first-order high pass filter. The -3 dB cutoff frequency of the filter is calculated using Equation 4.

$$f_{C\_HPF} = \frac{1}{2\pi * R_4 * C_4} \quad (4)$$

A value of 10 µF is selected for C_4, as it is a common value already used in the circuit for power supply decoupling. R_4 must be significantly higher resistance than R_{SRC} in order to prevent unwanted attenuation from the voltage divider formed by these two resistances. Therefore, an initial value of 499 kΩ is selected for R_4 and the high pass filter cutoff frequency is calculated using Equation 5.

$$f_{C\_HPF} = \frac{1}{2\pi * R_4 * C_4} = \frac{1}{2\pi * 499k\Omega * 10\mu F} = 0.03\ \text{Hz} \quad (5)$$

The -3 dB cutoff frequency of the filter is calculated to be 0.03 Hz. This will effectively ac couple the input signal while preserving the gain and phase behavior at 20 Hz.

2.1.3 Input Filter Transfer Function

The complete transfer function of the input filter is shown in Figure 4.

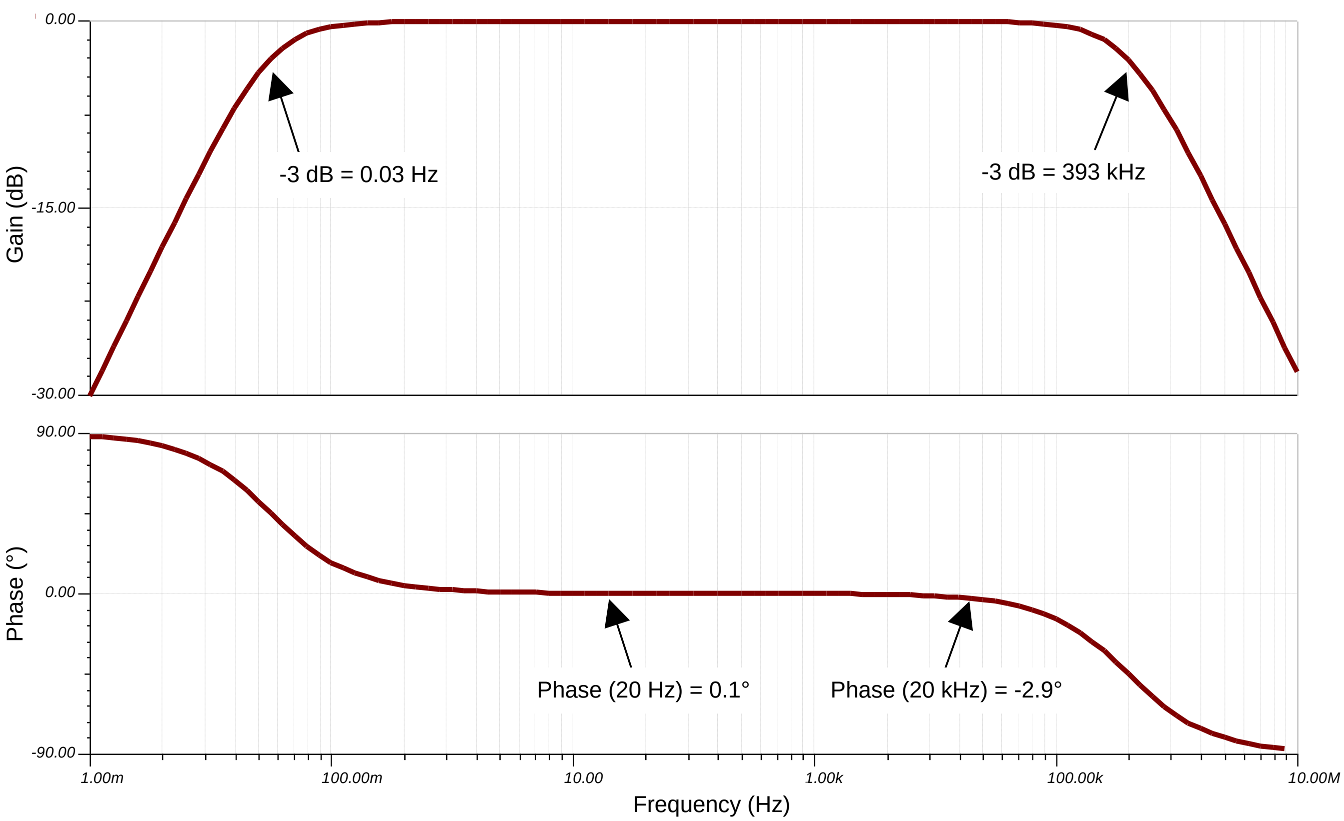


Figure 4: Transfer Function - Input Filter

2.2 Input Gain Stage

The input signal in this design is specified at 1 V_{RMS}, or 1.414 V_{PK}. Since the OPA1642 used in this design can swing its output voltage within 200 mV from each rail and ±15 V power supply rails are provided, an input gain stage is used to amplify the input signal as needed. The gain stage is made up of amplifier U_{1A}, potentiometer P_1, and resistors R_1 and R_2, as shown in Figure 5.

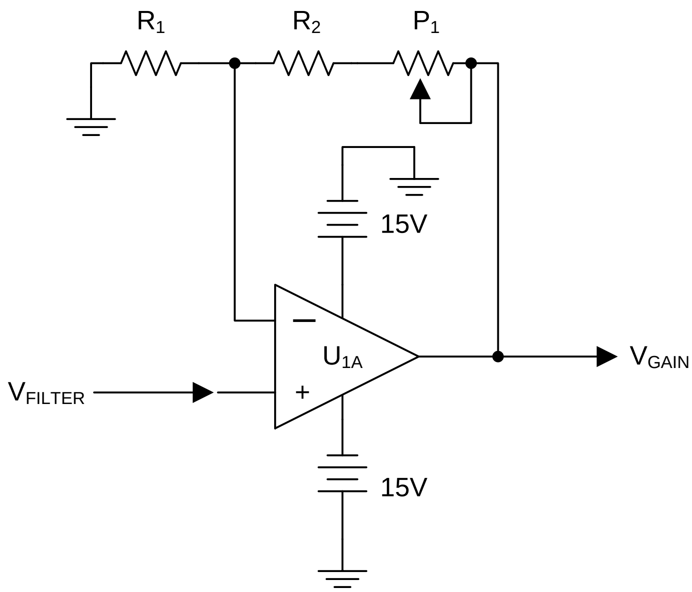


Figure 5: Input Gain Stage Schematic

This straightforward non-inverting gain stage has a transfer function as defined as in Equation 6, where R_{P1} is the equivalent series resistance of potentiometer P_1.

$$V_{GAIN} = \left(1 + \frac{R_2 + R_{P1}}{R_1}\right) * V_{FILTER} \quad (6)$$

At 0% rotation of potentiometer P_1, its equivalent series resistance is 0 Ω. Therefore the transfer function of the gain stage simplifies to Equation 7.

$$V_{GAIN} = \left(1 + \frac{R_2}{R_1}\right) * V_{FILTER} \quad (7)$$

The minimum gain of the circuit is specified at 6 dB, or 2 V/V. To achieve this, the resistances of R_1 and R_2 must be equal. A value of 1 kΩ is selected for R_1 and R_2 in order to ensure low thermal noise. The maximum gain of the circuit is specified at 28 dB, or approximately 25 V/V. This gain occurs at 100% rotation of potentiometer P_1. The required value of R_{P1} is calculated by rearranging the terms of Equation 6 and solving for R_{P1}, as shown in Equation 8.

$$R_{P1} = R_1 * \left(\frac{V_{GAIN}}{V_{FILTER}} - 1\right) - R_2 = 1k\Omega * (25 - 1) - 1k\Omega = 23k\Omega \quad (8)$$

The required value for R_{P1} is calculated to be 23 kΩ. The nearest standard value of 25 kΩ is selected as the actual value.

2.3 Tone Stack

The tone stack is a passive filter network which allows a guitarist to control the frequency response of the amplifier [1]. Many different tone stack implementations exist, but this design uses what is known as the FMV tone stack. Introduced by Fender in 1957 in the 5F6 Bassman, it was later copied by Marshall, Vox, and many other guitar amplifier manufacturers [2]. Because of its ubiquity, the circuit is very well understood with extensive analysis and documentation widely available.

The tone stack is made up of capacitors C_1, C_3, C_8, and C_9, jumper JMP1, resistor R_3, and potentiometers P_2, P_3, and P_4, as shown in Figure 6.

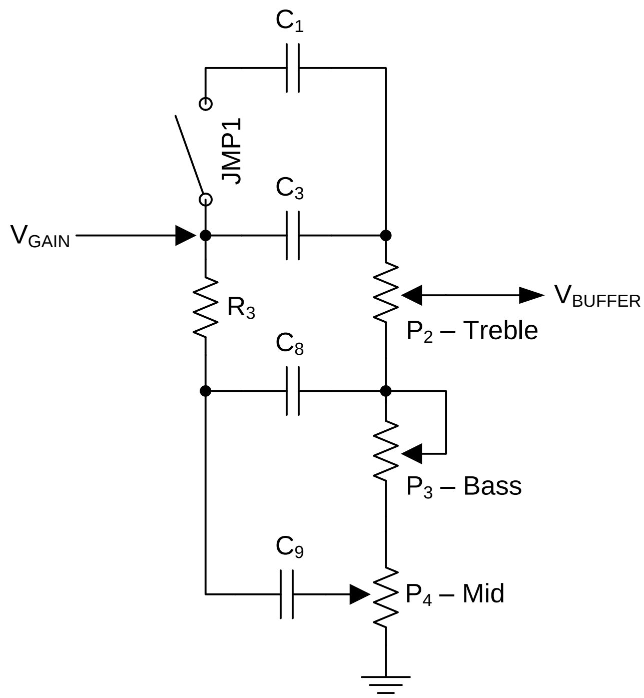


Figure 6: Tone Stack Schematic

Since the tone stack contains multiple filters with many possible states and interactive impedances, it is not trivial to analyze. Rather than perform a complete analysis here, the approach of this document will be to summarize the effect of each potentiometer on the circuit behavior, provide equations which allow the user to customize component values, and then refer to material where the full analysis is available.

2.3.1 Potentiometer Effects

The treble potentiometer P_2 acts as a balance control between the output of a high-pass filter formed by C_3 (in parallel with C_1 if JMP1 is closed) and the series combination of all three potentiometers, and the output of the complex filter created by R_3, C_8, C_9, P_3, and P_4.

The bass potentiometer P_3 sets the lower -3 dB cutoff frequency of a band-pass filter formed by R_3, C_8, P_3, and P_4. It also affects the -3 dB cutoff frequency of the treble control circuit.

The mid potentiometer P_4 controls the attenuation of the band-pass filter formed by R_3, C_9, and P_4. It also acts as a variable attenuator for the tone stack output [3].

2.3.2 Tone Stack Component Values

Calculating the values of R_3 and P_4 first allows the circuit designer to set the attenuation of all frequencies when the bass and treble controls are at 0% and the mid control is at 100%. An initial value of 25 kΩ is selected for P_4 as it is already used for the gain potentiometer P_1. The value of R_3 can then be calculated using Equation 9, where R_{P4} is the maximum resistance of P_4 and A is a positive number representing the desired amount of attenuation in dB. This design targets the frequency response of the Marshall JMP50 amplifier, where a value of 7.3 is used for A.

$$\mathrm{R_3} = \mathrm{R_{P4}} * \left(10^{\frac{A}{20}} - 1\right) = 25\ \mathrm{k\Omega} * \left(10^{\frac{7.3}{20}} - 1\right) = 33\ \mathrm{k\Omega} \quad (9)$$

The required value for R_3 is calculated to be 33 kΩ, which is a standard value.

Once the values of R_3 and P_4 are set, the value of C_9 can be determined. This capacitor defines the upper cutoff frequency off the bass passband, which is a function of C_9 and R_3. C_3 is calculated as shown in Equation 10, where f_1 is the upper cutoff frequency of the bass passband. The Marshall JMP50 uses a cutoff frequency of 219 Hz.

$$\mathrm{C_9} = \frac{1}{2\pi * \mathrm{f_1} * \mathrm{R_3}} = \frac{1}{2\pi * 219\ \mathrm{Hz} * 33\mathrm{k\Omega}} = 22\ \mathrm{nF} \quad (10)$$

The required value for C_9 is calculated to be 22 nF, which is a standard value.

Next, the value of C_8 can be determined in order to complete the bass passband design. C_8 controls the amount of bass attenuation when the bass potentiometer is at 0% (in a short-circuit condition), so the resistance of P_3 is not included in the calculation. A value of 1 MΩ is selected for P_3, consistent with the values used in the Marshall JMP50. This ensures that the lower end of the bass passband is well below the lowest frequencies output by a guitar. C_8 is calculated as shown in Equation 11, where f_2 is the lower cutoff frequency of the bass passband. The Marshall JMP50 uses a cutoff frequency of 62 Hz.

$$\mathrm{C_8} = \frac{1}{2\pi * \mathrm{f_2} * (\mathrm{R_3} + \mathrm{R_{P4}})} - \mathrm{C_9} = \frac{1}{2\pi * 62\ \mathrm{Hz} * (33\ \mathrm{k\Omega} + 25\ \mathrm{k\Omega})} - 22\ \mathrm{nF} = 22\ \mathrm{nF} \quad (11)$$

The required value for C_8 is also calculated to be 22 nF.

Finally, the values of C_3 and P_2 can be selected, which set the cutoff frequency of the treble high pass filter. A value of 250 kΩ is selected for P_2, and C_3 is calculated as shown in Equation 12, where f_3 is the cutoff frequency of the treble high pass filter and R_{P2} is the maximum resistance of P_2. The Marshall JMP50 uses a cutoff frequency of 1.4 kHz.

$$\mathrm{C_3} = \frac{1}{2\pi * \mathrm{f_3} * R_{P2}} = \frac{1}{2\pi * 1.4\ \mathrm{kHz} * 250\ \mathrm{k\Omega}} = 455\ \mathrm{pF} \quad (12)$$

The required value for C_3 is calculated to be 455 pF. The nearest standard value of 470 pF is selected as the actual value.

Closing switch JMP1 connects C_1 in parallel with C_3, adding the two capacitances to the cutoff frequency calculation. If C_1 also has a value of 470 pF, the cutoff frequency will be reduced by a factor of two to approximately 700 Hz [4].

The calculated component values and associated tone stack characteristics are summarized in Table 2.

Table 2. Summary of Tone Stack Component Values and Performance Characteristics

	Components	Values	Behavior
Overall attenuation	P_4	25 kΩ	-7.3 dB
	R_3	33 kΩ	
Bass passband upper cutoff frequency	C_9	22 nF	219 Hz
Bass passband lower cutoff frequency	C_8	22 nF	62 Hz
	P_3	1 MΩ	
Treble high pass cutoff frequency	C_3	470 pF	1.4 kHz
	P_2	250 kΩ	
Mid boost high pass cutoff frequency	C_1	470 pF	700 Hz

2.3.3 Further Reading

If the reader wishes to expand their understanding of the FMV tone stack, a more thorough analysis is available in *Circuit Analysis of a Legendary Tube Amplifier: The Fender Bassman 5F6-A* by Richard Kuehnel. Another useful resource is *Designing Tube Preamps for Guitar and Bass* by Merlin Blencowe, which discusses the FMV tone stack as well as several other topologies along with their advantages and disadvantages.

3 Component Selection

3.1 Amplifier

This tone stack circuit must provide gain and accurate control over the frequency response of the input audio signal while introducing as little distortion or noise as possible. Therefore, the amplifier selected must have very low distortion and noise performance in the audio frequency range, even when high source impedances are present [5]. A wide supply voltage range is also required, as most professional audio circuits use large split supplies in order to avoid output clipping. Low quiescent current and relatively low cost are also desirable qualities which help to maintain an efficient design.

The OPA1642 is an excellent choice for this high-performance audio application, with total harmonic distortion + noise (THD+N) of only -126 dB (0.00005%) and input voltage noise density of 5.1 nV/√Hz. The amplifier can utilize power supply voltages up to ±18 V while consuming only 1.8 mA of quiescent current per channel, and its reasonable price point ensures that the total solution cost remains competitive.

3.2 Passive Component Selection

3.2.1 Resistor Selection

The type of resistors used in an ultra-low distortion audio circuit can have a significant impact on the circuit's overall performance. Real resistors have a certain amount of nonlinearity, which results in unwanted contributions to distortion and noise [6]. The most common sources of resistor nonlinearity are temperature coefficient of resistance (TC_R), which describes how the resistance changes as a function of temperature, and voltage coefficient of resistance (VC_R), which describes how the resistance changes as a function of applied voltage. Both VC_R and TC_R are related to the resistor's self-heating – as the voltage across the resistor increases, the current through the resistor increases and its temperature rises.

Two of the most common types of surface mount resistors are thick film and thin film. Thin film resistors typically perform better than thick film resistors, but thin film resistors also typically cost several times as much. When high-level audio signals are involved, the lower VC_R and TC_R of thin film resistors can become critical to achieving ultra-low distortion performance.

This design involves audio signals with maximum amplitude of 15 V_{PK}, or 10.6 V_{RMS}. Despite the significant voltage, current through the signal path resistors remains low and meaningful self-heating does not occur. Therefore, all signal path resistors on the board are thick film, ±1% tolerance, 0.1 watt devices in a 0603 package.

3.2.2 Capacitor Selection

Like resistors, capacitors also have a voltage coefficient (VC_C), which describes how the capacitance changes as a function of applied voltage. This change in capacitance results in unwanted distortion [7], so any capacitors in the audio signal path which can be subjected to significant voltages should have low VC_C.

The critical signal path capacitors in this design are C_7 in the input filter and C_1, C_3, C_8, and C_9 in the tone stack. For these components NP0-type capacitors are used. All other capacitors for ac coupling and power supply bypass are of type X7R.

3.2.3 Potentiometer Selection

Other than the maximum resistance and taper characteristics, the potentiometers selected for this design are not critical. Single-turn rotary potentiometers with ±20% tolerance and right-angle PCB mount termination were selected for this application.

4 Simulation

The TINA-TI™ schematic shown in Figure 7 includes the circuit values obtained in the design process. The source impedance of V_{IN}, the input signal, is included as a discrete resistance R_{SRC}. A load resistance of 100 kΩ is added to simulate the input resistance of the audio analyzer which is used for real-world measurements.

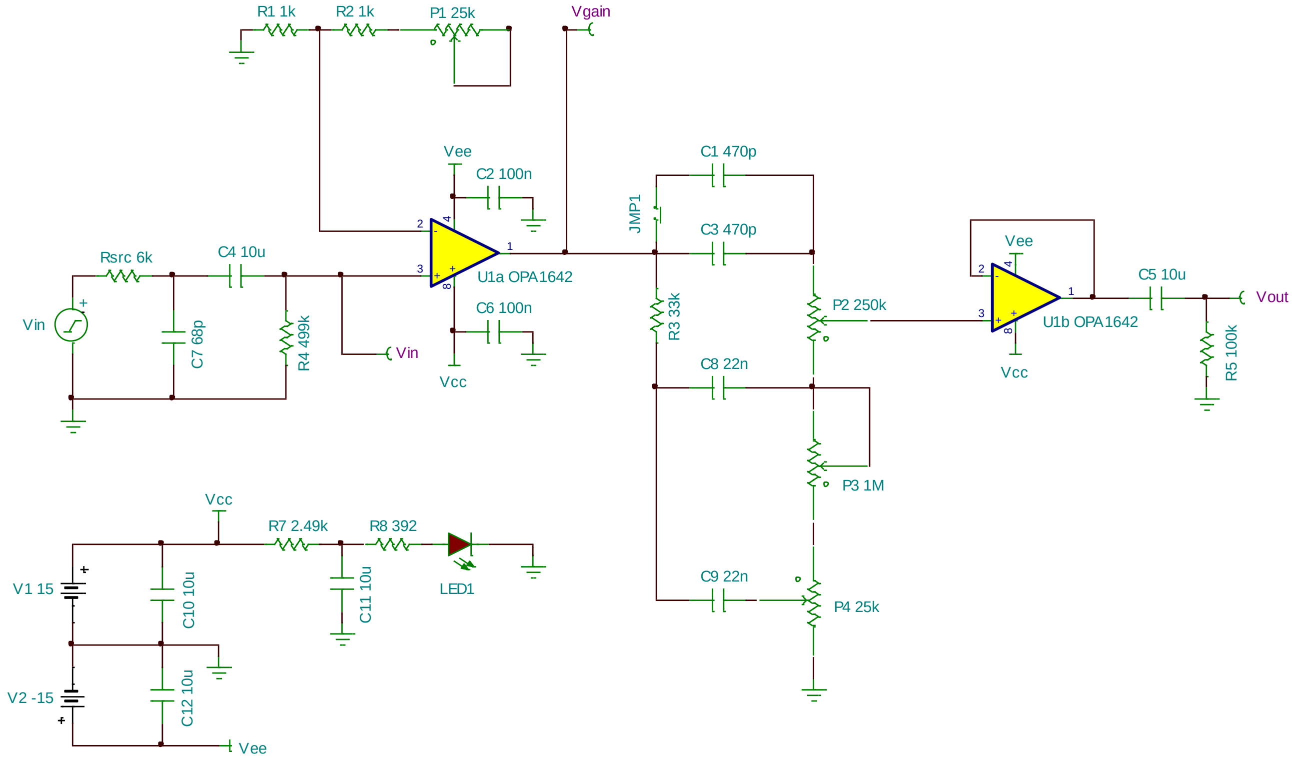


Figure 7: TINA-TI™ Schematic

4.1 Gain Characteristic

The result of the simulated gain characteristic as a function of gain potentiometer P_1 rotation is shown in Figure 8.

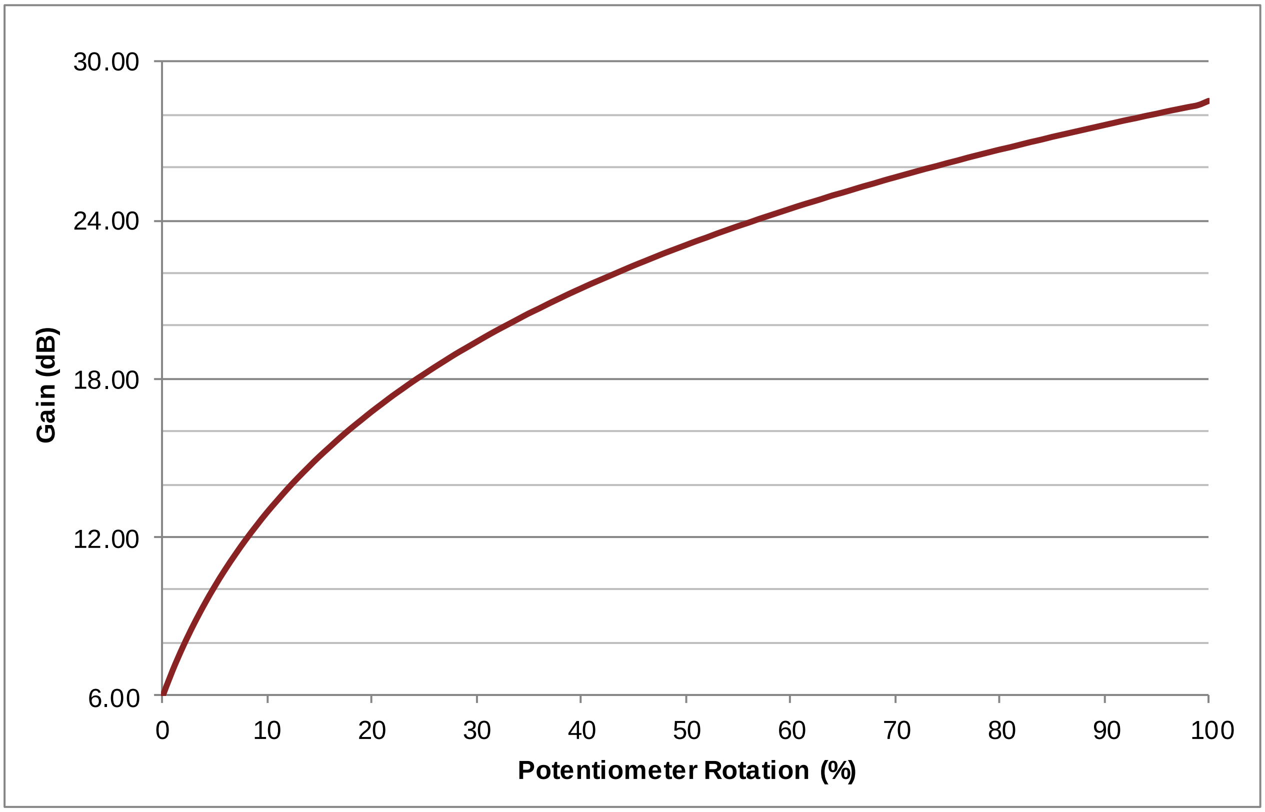


Figure 8: Simulated Gain Characteristic

4.2 Frequency Response

4.2.1 Input Filter and Gain Stage

The result of the simulated ac analysis of the input filter and gain stage when gain = 6 dB is shown in Figure 9.

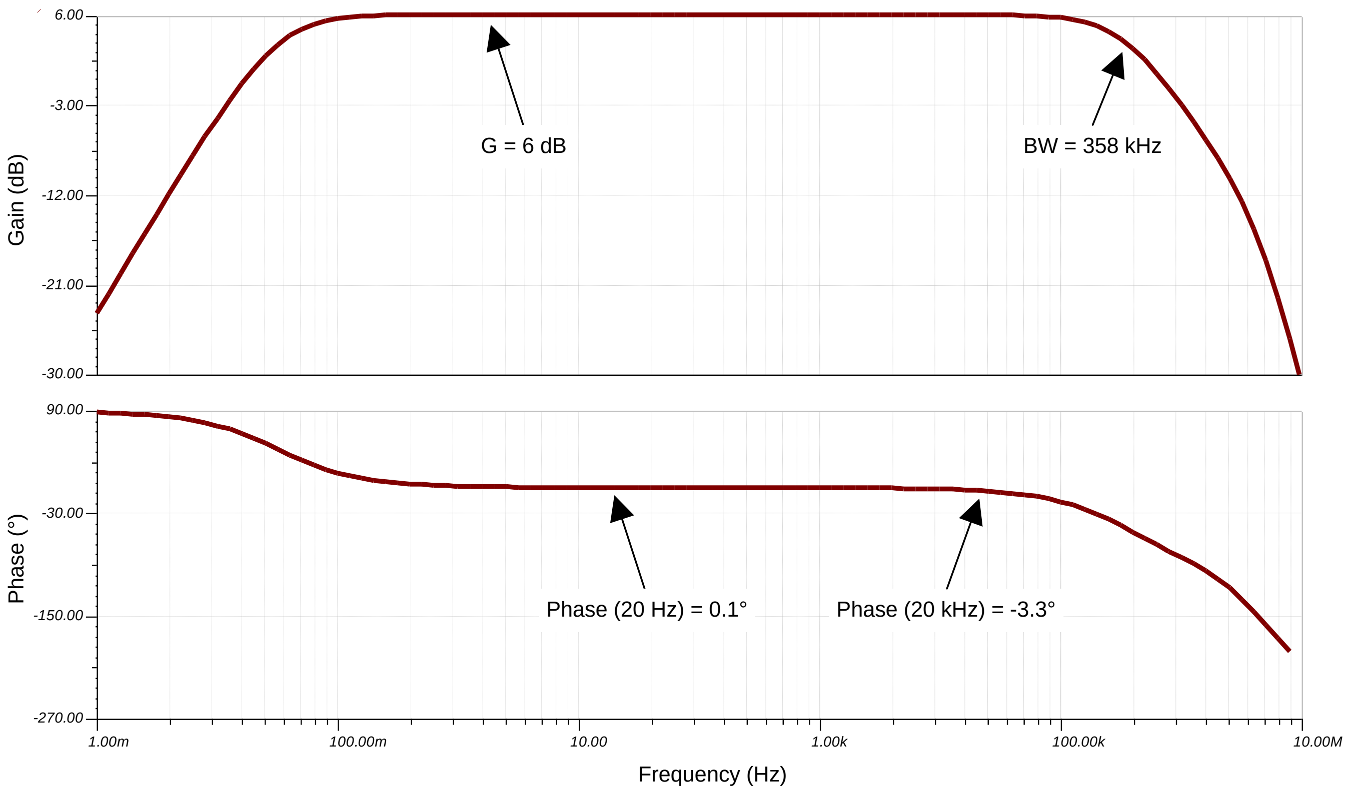


Figure 9: Simulated AC Analysis – Input Filter and Gain Stage

The gain of the simulation throughout the audio band was measured to be 6 dB. The -3 dB bandwidth was 358 kHz. The phase of the simulation was 0.1° at 20 Hz and -3.3° at 20 kHz.

4.2.2 *Tone Stack*

4.2.2.1 Treble Control

The result of the simulated ac analysis of the complete circuit as the treble potentiometer P_2 is rotated, when gain = 6 dB, the mid and bass potentiometers are set to 50%, and mid boost is off, is shown in Figure 10.

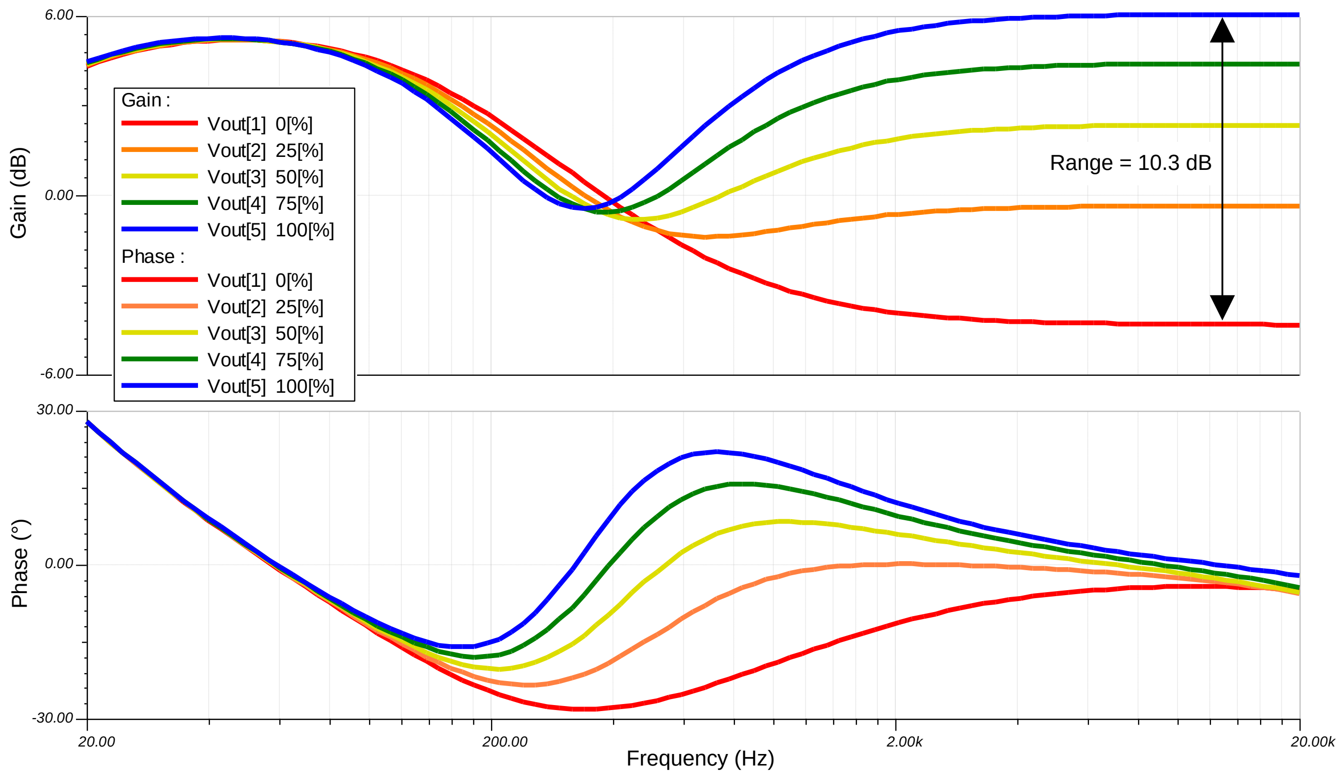


Figure 10: Simulated AC Analysis – Treble Control

In this condition, the treble gain varies from -4.3 dB at 0% potentiometer rotation to +6.0 dB at 100% potentiometer rotation. This gives an adjustment range of 10.3 dB, which meets the design requirement of 10 dB.

4.2.2.2 Mid Control

The result of the simulated ac analysis of the complete circuit as the mid potentiometer P_4 is rotated, when gain = 6 dB, the treble and bass potentiometers are set to 50%, and mid boost is off, is shown in Figure 11.

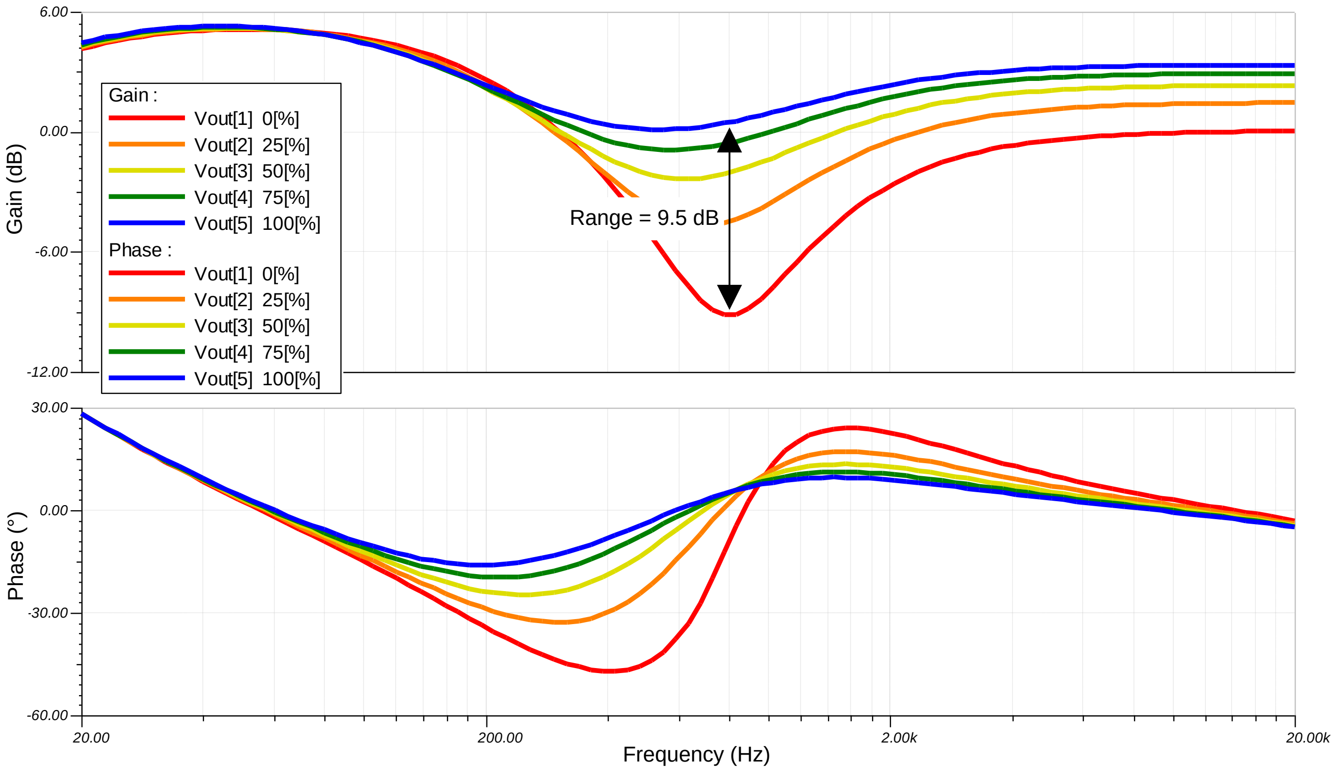


Figure 11: Simulated AC Analysis – Mid Control

In this condition, the mid gain varies from -9.1 dB at 0% potentiometer rotation to +0.4 dB at 100% potentiometer rotation. This gives an adjustment range of 9.5 dB, which meets the design requirement of 6 dB.

4.2.2.3 Mid Boost

The result of the simulated ac analysis of the complete circuit as the mid boost jumper is connected and disconnected, when gain = 6dB and all potentiometers are set to 50%, is shown in Figure 12.

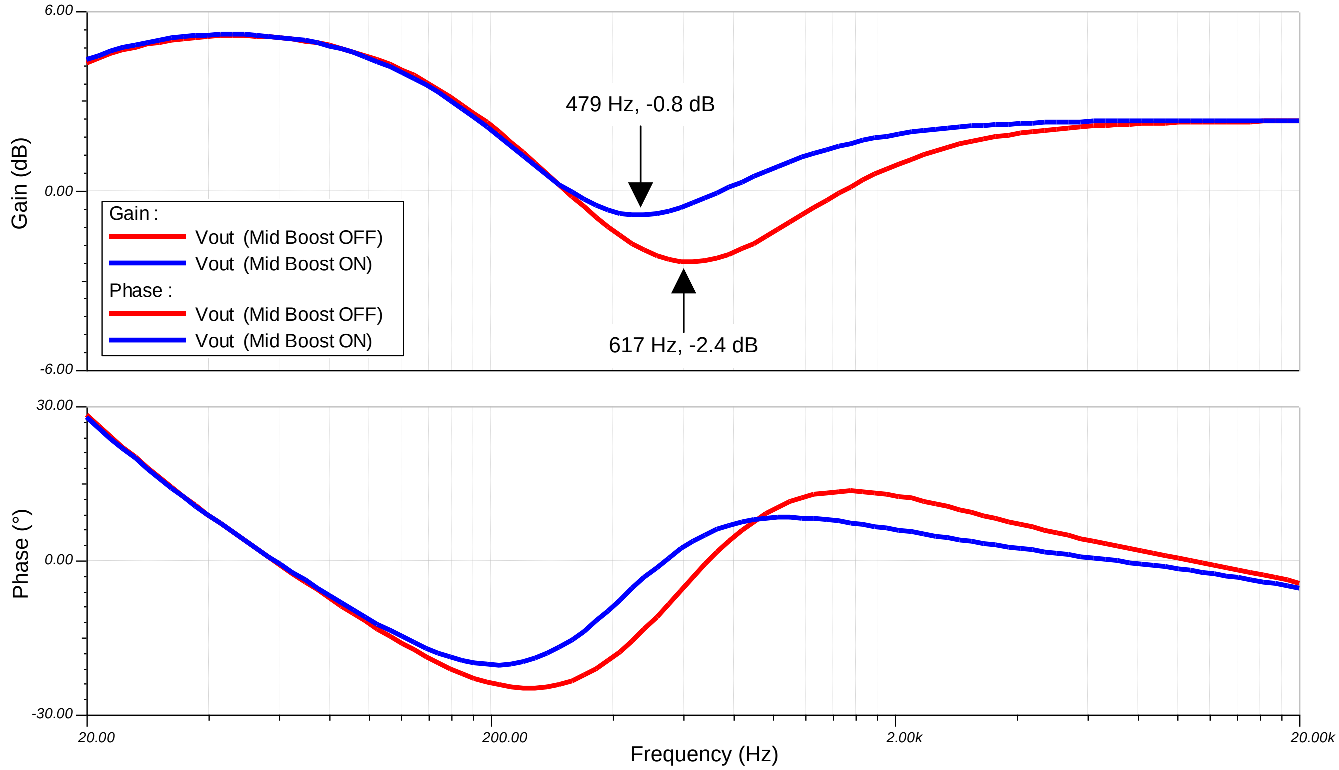


Figure 12: Simulated AC Analysis – Mid Boost

Activating the mid boost lowered the mid cut frequency from 617 Hz to 479 Hz and boosted the gain at the cut frequency from -2.4 dB to -0.8 dB.

4.2.2.4 Bass Control

The result of the simulated ac analysis of the complete circuit as the bass potentiometer P_3 is rotated, when gain = 6 dB, the treble and mid potentiometers are set to 50%, and mid boost is off, is shown in Figure 13.

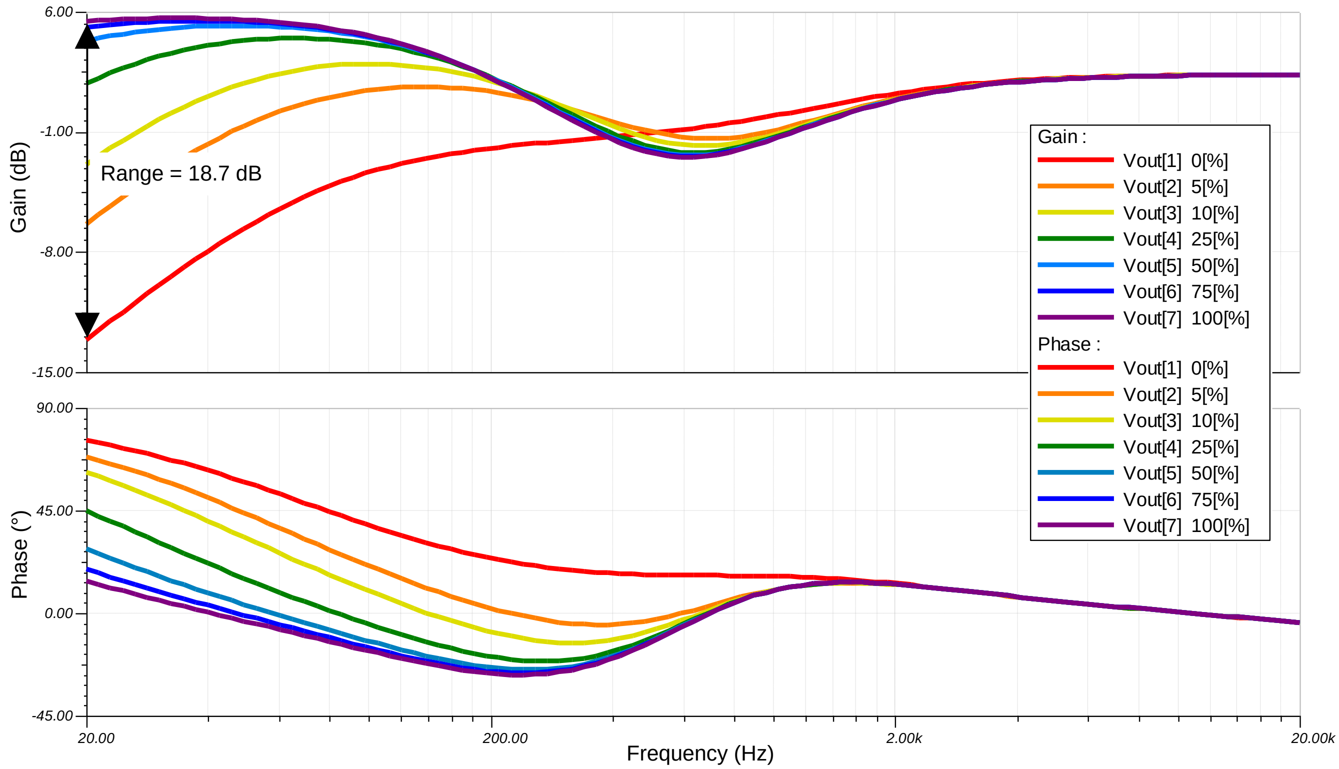


Figure 13: Simulated AC Analysis – Bass Control

In this condition, the bass gain varies from -13.2 dB at 0% potentiometer rotation to +5.5 dB at 100% potentiometer rotation. This gives an adjustment range of 18.7 dB, which meets the design requirement of 15 dB.

4.3 THD+N Performance

Unfortunately, TI's op amp macromodels do not currently support proper THD+N analysis. However, the THD+N ratio of a circuit (when noise is dominant) can be predicted from the total noise analysis by using Equation 13, where V_N is the total voltage noise in V_{RMS} over a specified bandwidth and V_F is the fundamental signal amplitude in V_{RMS}.

$$\mathrm{THD+N(\%)} = \sqrt{\frac{{V_N}^2}{{V_F}^2}} * 100 \qquad (13)$$

The result of the simulated total noise analysis at gain = 6 dB, all tone potentiometers set to 100%, and mid boost on is shown in Figure 14. The audio analyzer which is used for real-world measurements will be set to a measurement bandwidth of 90 kHz, so this simulated noise analysis is performed to 90 kHz.

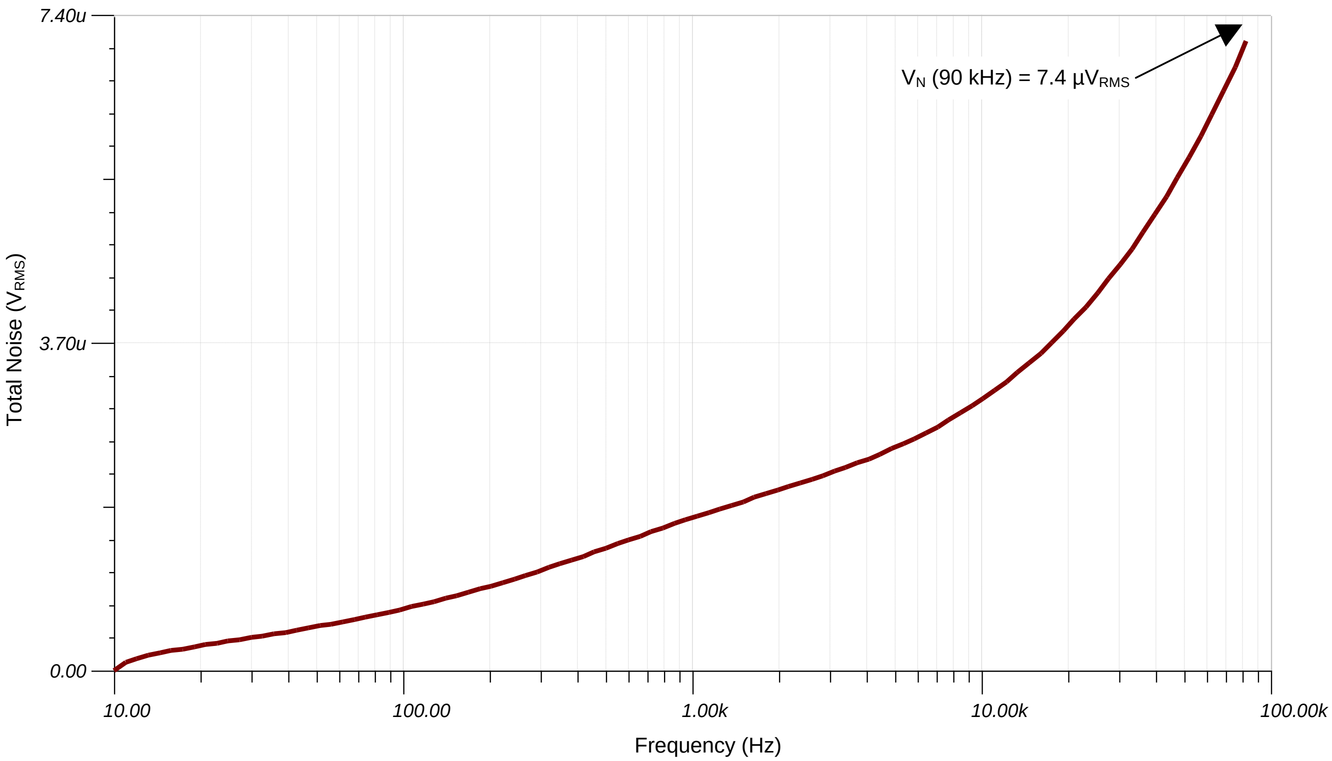


Figure 14: Simulated Total Noise Analysis

The total noise at 90 kHz was found to be 7.4 µV_{RMS}. Given our input signal amplitude of 1 V_{RMS}, the predicted THD+N ratio is calculated using Equation 14.

$$\mathrm{THD+N(\%)} = \sqrt{\frac{{V_N}^2}{{V_F}^2}} * 100 = \sqrt{\frac{(7.4\ \mu V_{RMS})^2}{(1\ V_{RMS})^2}} * 100 = 0.00074\% = -102.6\ \mathrm{dB} \qquad (14)$$

The simulated THD+N ratio was found to be -102.6 dB (0.00074%), which meets the design requirement of -100 dB. However, this does not account for the possibility of harmonic distortion due to output clipping which can occur at higher gain settings.

4.4 Resistor Nonlinearity

As mentioned in Section 3.2.1, resistor nonlinearity due to TC_R and VC_R can have a negative effect on distortion performance. In order to determine if thin film resistors are required, the current through feedback resistors R_1 and R_2 is simulated during the worst-case condition when V_{IN} = 20 Hz and gain = 6 dB. The result of the transient analysis is shown in Figure 15.

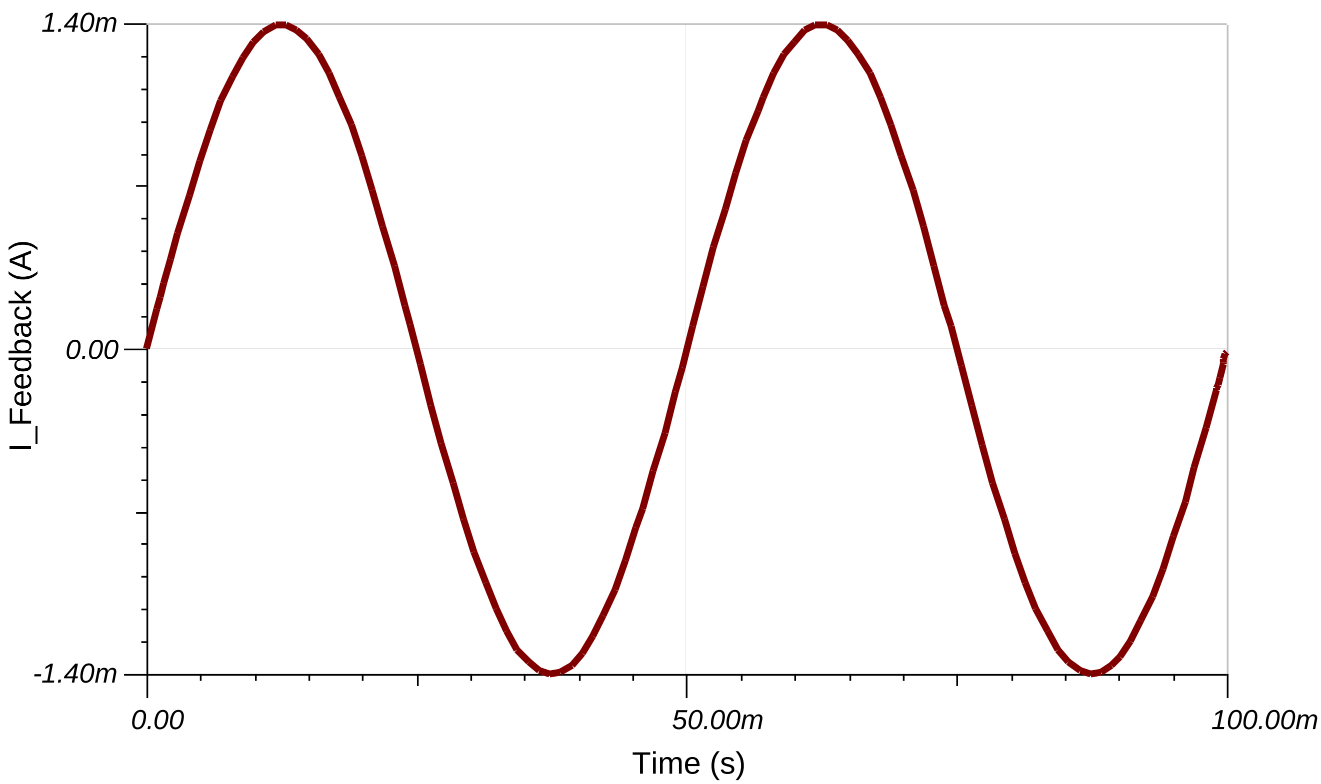


Figure 15: Simulated Transient Analysis – Feedback Resistor Current

In this condition, the maximum current through the feedback network is approximately 1.4 mA. We calculate the maximum power dissipation through the feedback network using Equation 15.

$$P_{DIS}(W) = I^2 * R = 1.4\ mA^2 * 1k\Omega = 2\ mW \quad (15)$$

The maximum power dissipation of approximately 2 mW is well below the resistors' power handling of 100 mW, so TC_R and VC_R will not be an issue and thin film resistors are not required to achieve low-distortion performance.

4.5 Simulated Results Summary

Table 3 summarizes the simulated performance of the design.

Table 3. Comparison of Design Goals and Simulated Performance

	Goal	Simulated
THD+N ratio at 1 kHz	-100 dB (0.001%)	-102.6 dB (0.00074%)
Treble adjustment range	10 dB	10.3 dB
Mid adjustment range	6 dB	9.5 dB
Bass adjustment range	15 dB	18.7 dB

5 PCB Design

The PCB schematic and bill of materials can be found in the Appendix.

5.1 PCB Layout

The PCB used in this design is a 3.4" by 3.4" square. This generous size allows for efficient routing of critical components and the use of larger RCA, ¼", and banana jacks, as well as the four required potentiometers. The high-level approach to this layout was to place nearly all components on the top layer, with the op amp in the center of the board, input connections on the left, output connections on the right, and gain and tone control potentiometers on the bottom. The power supply bulk capacitors were placed on the bottom layer close to the banana jacks. The two low-frequency tone control capacitors were also placed on the bottom layer close to their associated potentiometers.

Standard precision analog PCB layout practices were used in order to achieve the best possible performance. All passive components in the analog signal path are placed and routed very tightly in order to minimize parasitics, and all decoupling capacitors are located very close to their associated power pins. Solid copper planes on both layers provide an excellent low-impedance path for return currents to ground, and stitching vias are used where necessary.

Connections to the split power supply are made at J_5, J_6, and J_7. Connections to the audio inputs and outputs are made at J_1, J_2, J_3, and J_4. RCA connectors J_1 and J_2 are used to easily connect to the test equipment when measuring system performance, while ¼" connectors J_3 and J_4 are used to connect standard guitar cables.

The PCB layout for both layers is shown in Figure 16.

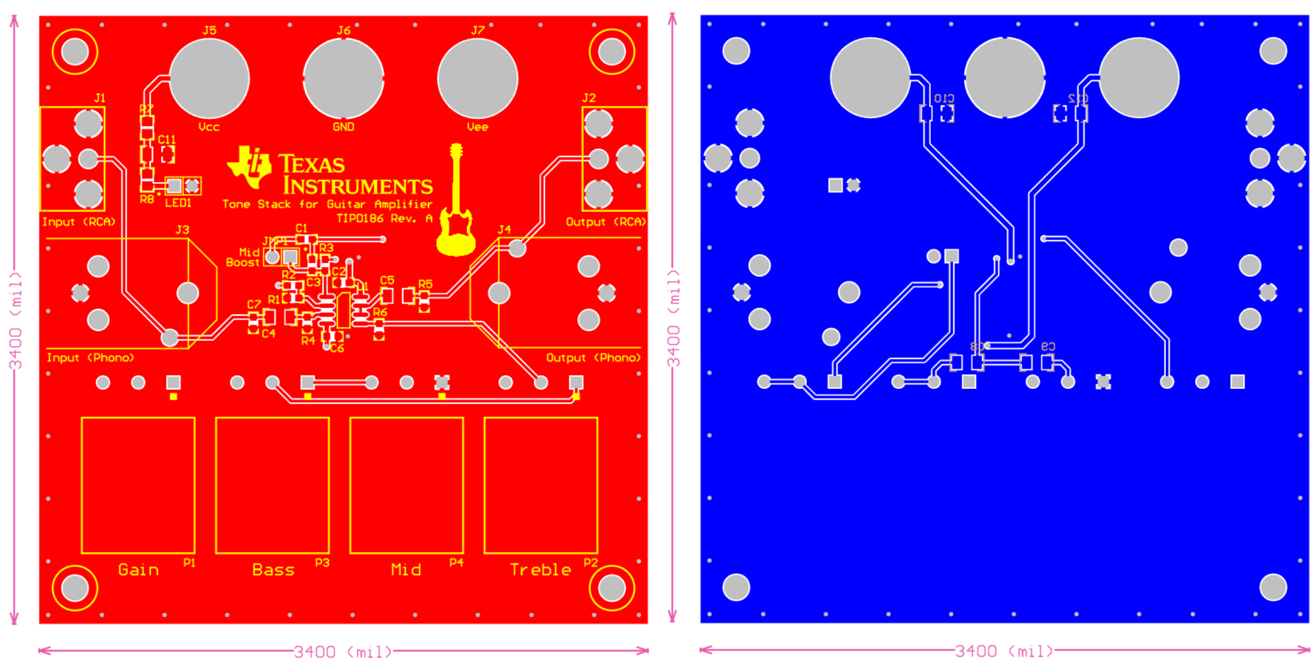


Figure 16: PCB Layout

6 Verification & Measured Performance

6.1 Bench Test Hardware Setup

The tone stack circuit defined by this reference design is intended for use within a complete guitar amplifier system. However, the circuit is also a standalone functional block whose real-world performance can be characterized. The convenient input, output and power connectors on the PCB allow the circuit to be easily tested on a bench using standard lab equipment. The test setup used consists of the components listed below. Figure 17 shows the bench test setup (computer not shown).

1. High performance audio analyzer: Provides the audio input and measures the audio output of the system.

2. Bode analyzer: Measures the gain and phase response of the system over frequency.

3. Personal computer (PC): Communicates with and controls the audio analyzer and Bode analyzer through a digital interface. Software provided by the hardware manufacturers allows the user to specify signal characteristics and perform measurements.

4. Triple output power supply: Provides ±15 V power supply rails to the system.

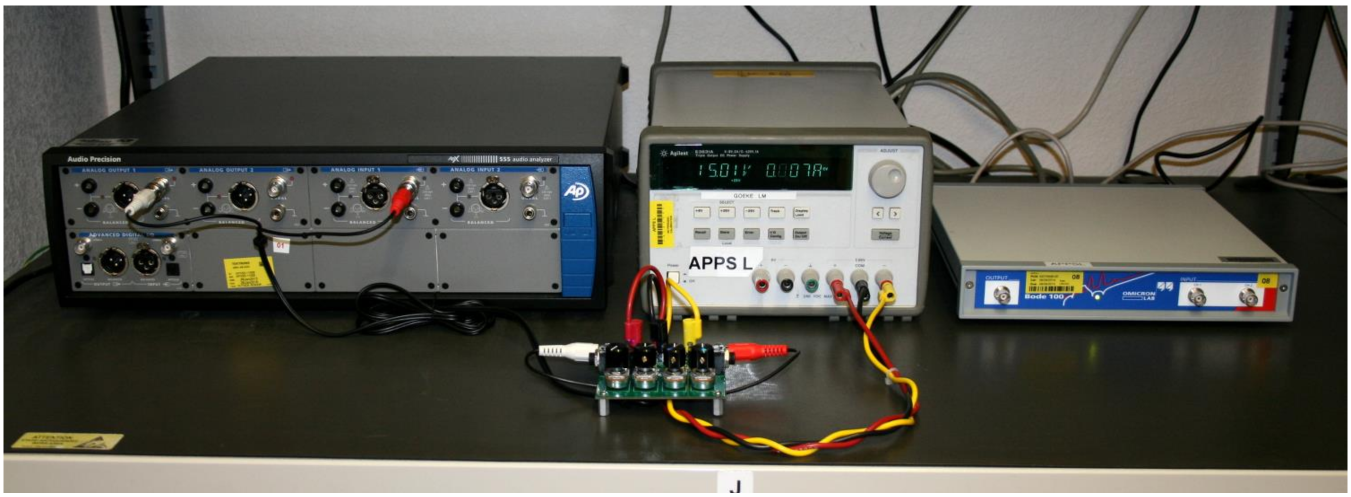


Figure 17: Bench Test Hardware Setup

6.2 Gain Characteristic

The result of the measured gain characteristic as a function of gain potentiometer P_1 rotation is shown in Figure 18. Gain was measured at 0%, 25%, 50%, 75%, and 100% rotation.

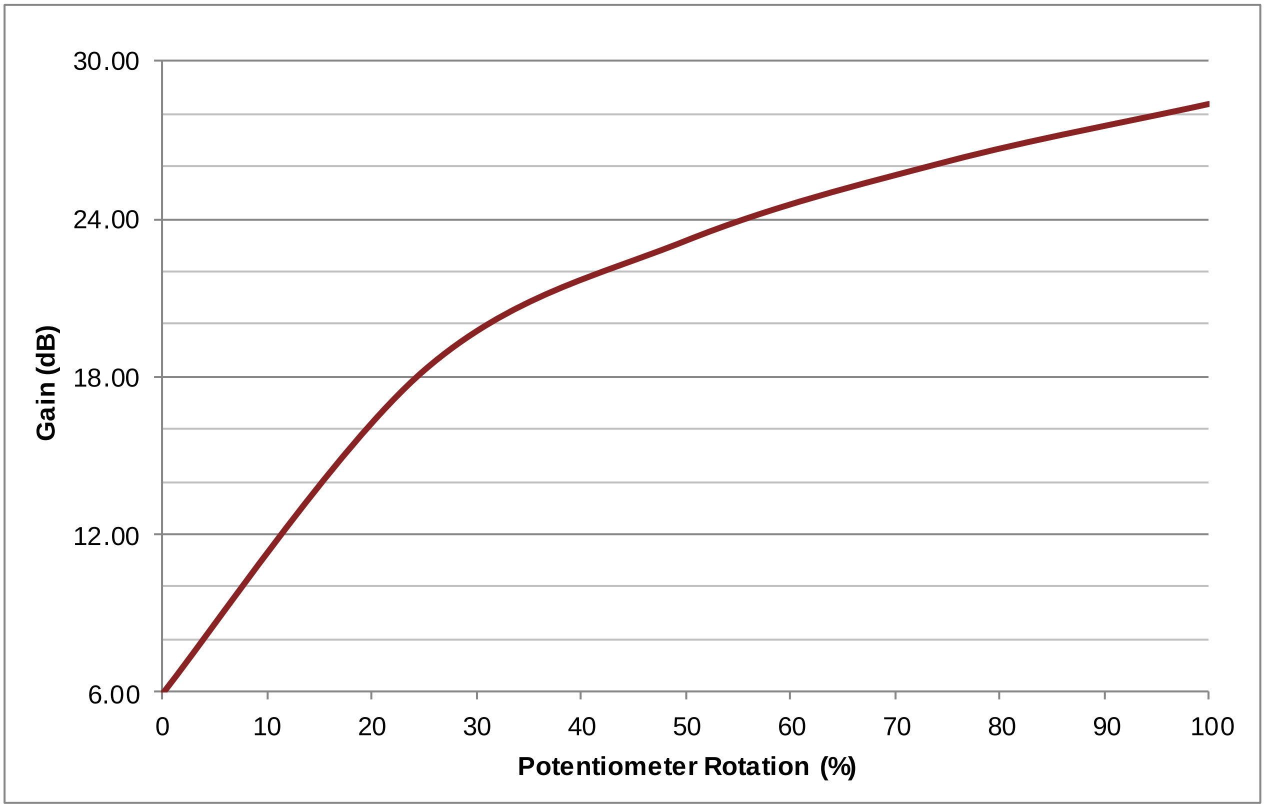


Figure 18: Measured Gain Characteristic

6.3 Frequency Response

6.3.1 Input Filter and Gain Stage

The result of the measured ac analysis of the input filter and gain stage when gain = 6 dB is shown in Figure 19.

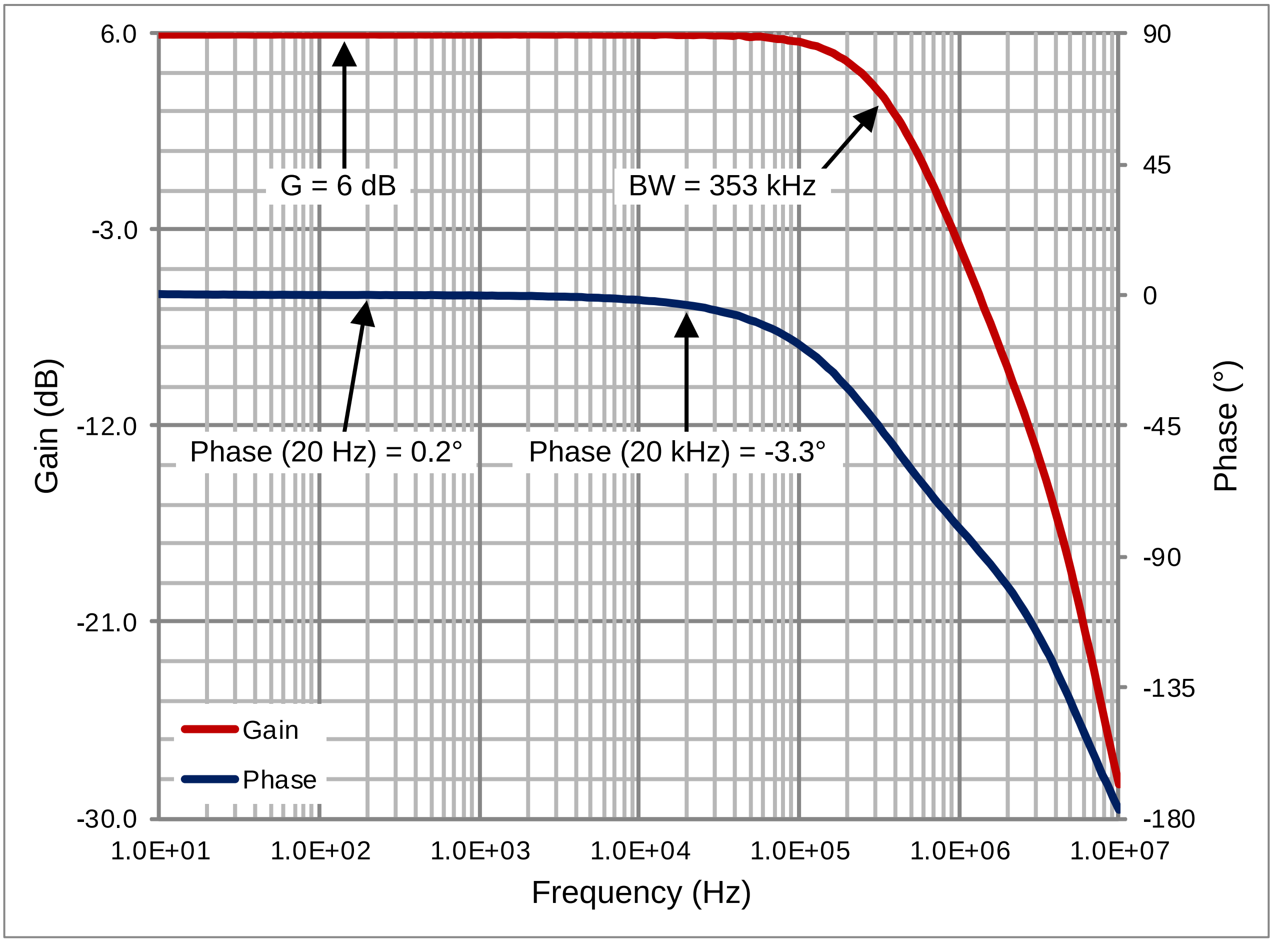


Figure 19: Measured AC Analysis – Input Filter and Gain Stage

The gain of the measurement throughout the audio band was measured to be 6 dB. The measured -3 dB bandwidth was 353 kHz, which correlates very well to the simulated -3 dB bandwidth of 358 kHz. The phase of the circuit was measured to be 0.2° at 20 Hz and -3.3° at 20 kHz, which is nearly the exact result found in simulation.

6.3.2 *Tone Stack*

6.3.2.1 Treble Control

The result of the measured ac gain analysis of the complete circuit as the treble potentiometer P_2 is rotated, when gain = 6 dB, the mid and bass potentiometers are set to 50%, and mid boost is on, is shown in Figure 20.

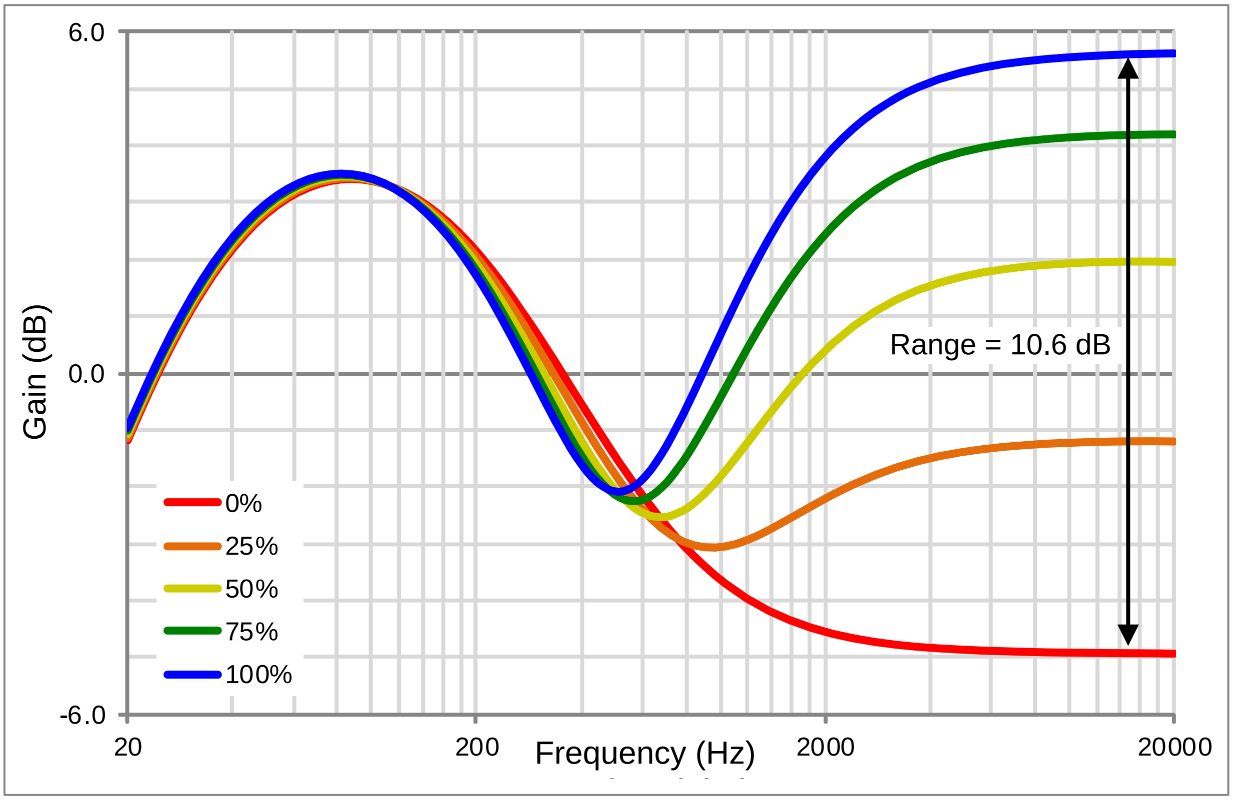


Figure 20: Measured AC Gain Analysis – Treble Control

In this condition, the treble gain varies from -5.0 dB at 0% potentiometer rotation to +5.6 dB at 100% potentiometer rotation. This gives an adjustment range of 10.6 dB, which meets the design requirement of 10 dB.

The result of the measured ac phase analysis under the same conditions is shown in section A.3.

6.3.2.2 Mid Control

The result of the measured ac gain analysis of the complete circuit as the mid potentiometer P_4 is rotated, when gain = 6 dB, the treble and bass potentiometers are set to 50%, and mid boost is on, is shown in Figure 21.

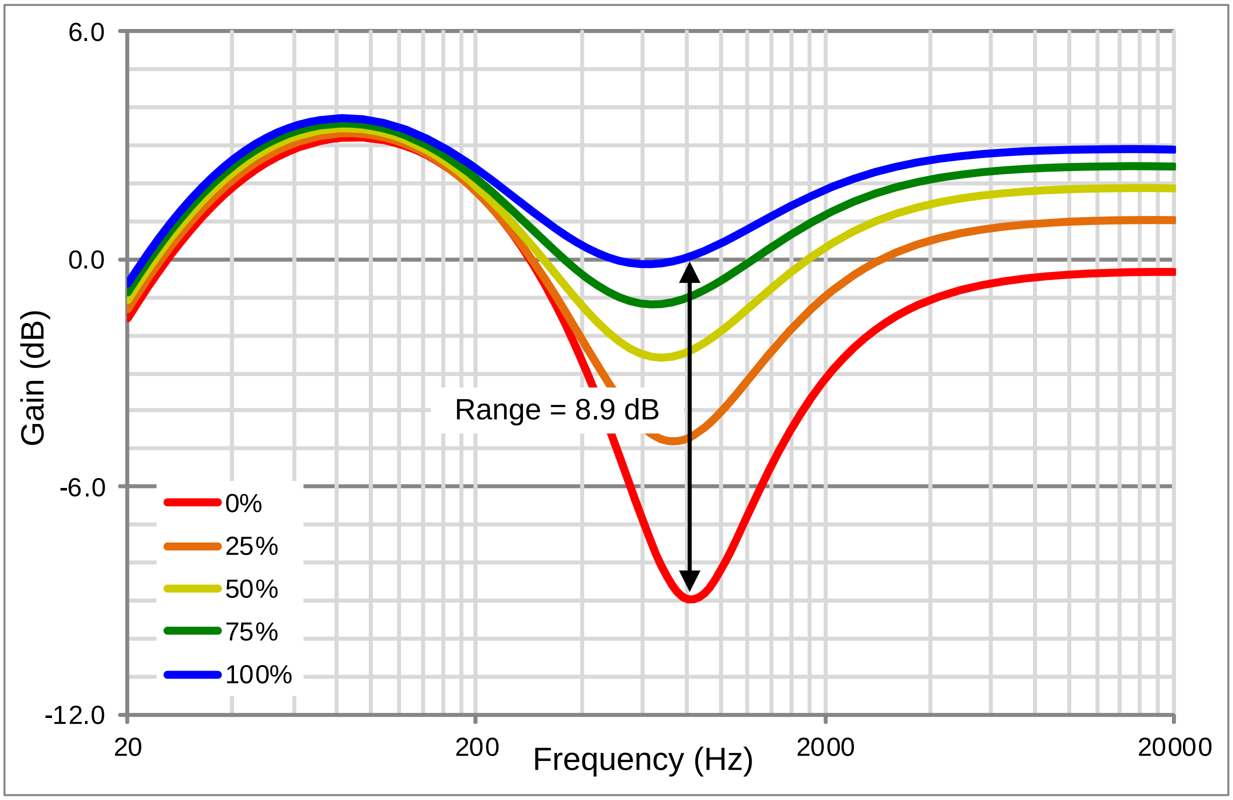


Figure 21: Measured AC Gain Analysis – Mid Control

In this condition, the mid gain varies from -8.9 dB at 0% potentiometer rotation to 0 dB at 100% potentiometer rotation. This gives an adjustment range of 8.9 dB, which meets the design requirement of 6 dB.

The result of the measured ac phase analysis under the same conditions is shown in section A.3.

6.3.2.3 Mid Boost

The result of the measured ac gain analysis of the complete circuit as the mid boost jumper is connected and disconnected, when gain = 6dB and all potentiometers are set to 50%, is shown in Figure 22.

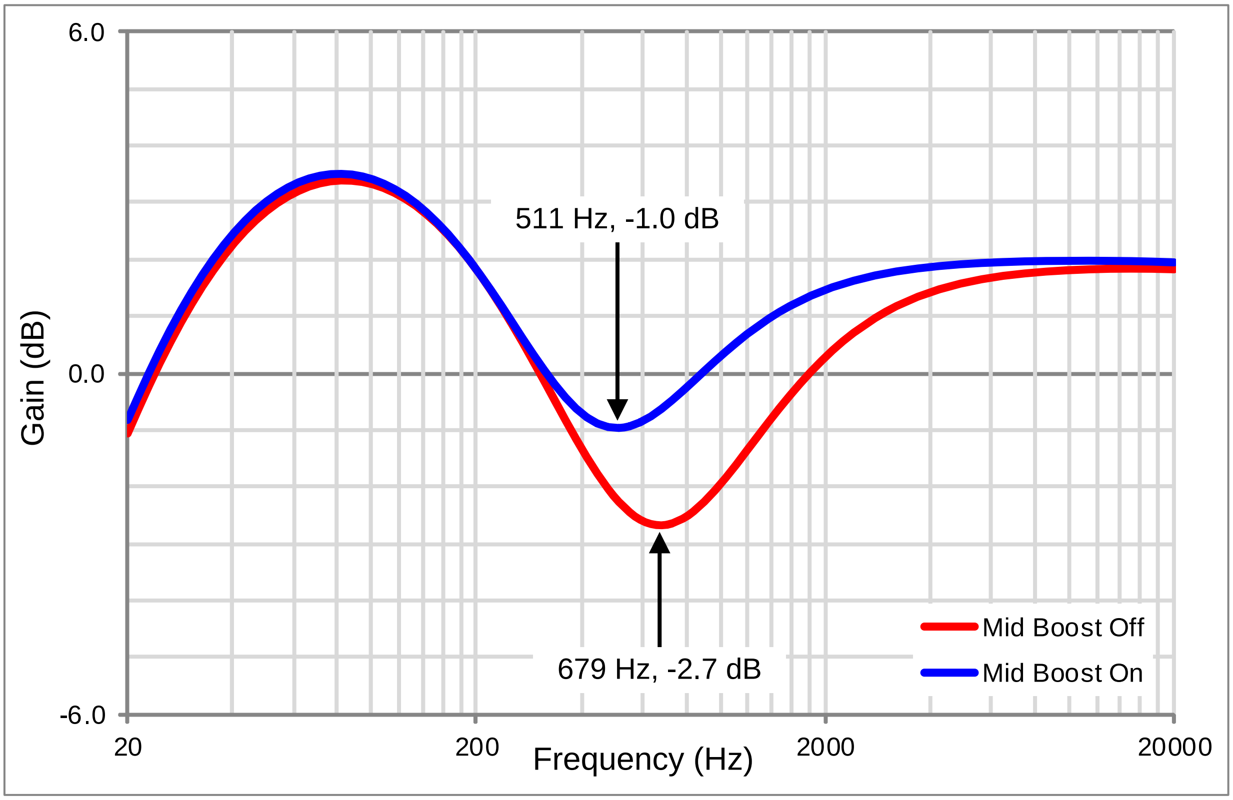


Figure 22: Measured AC Gain Analysis – Mid Boost

Activating the mid boost lowered the mid cut frequency from 679 Hz to 511 Hz and boosted the gain at the cut frequency from -2.7 dB to -1.0 dB.

The result of the measured ac phase analysis under the same conditions is shown in section A.3.

6.3.2.4 Bass Control

The result of the measured ac gain analysis of the complete circuit as the bass potentiometer P_3 is rotated, when gain = 6 dB, the treble and mid potentiometers are set to 50%, and mid boost is on, is shown in Figure 23.

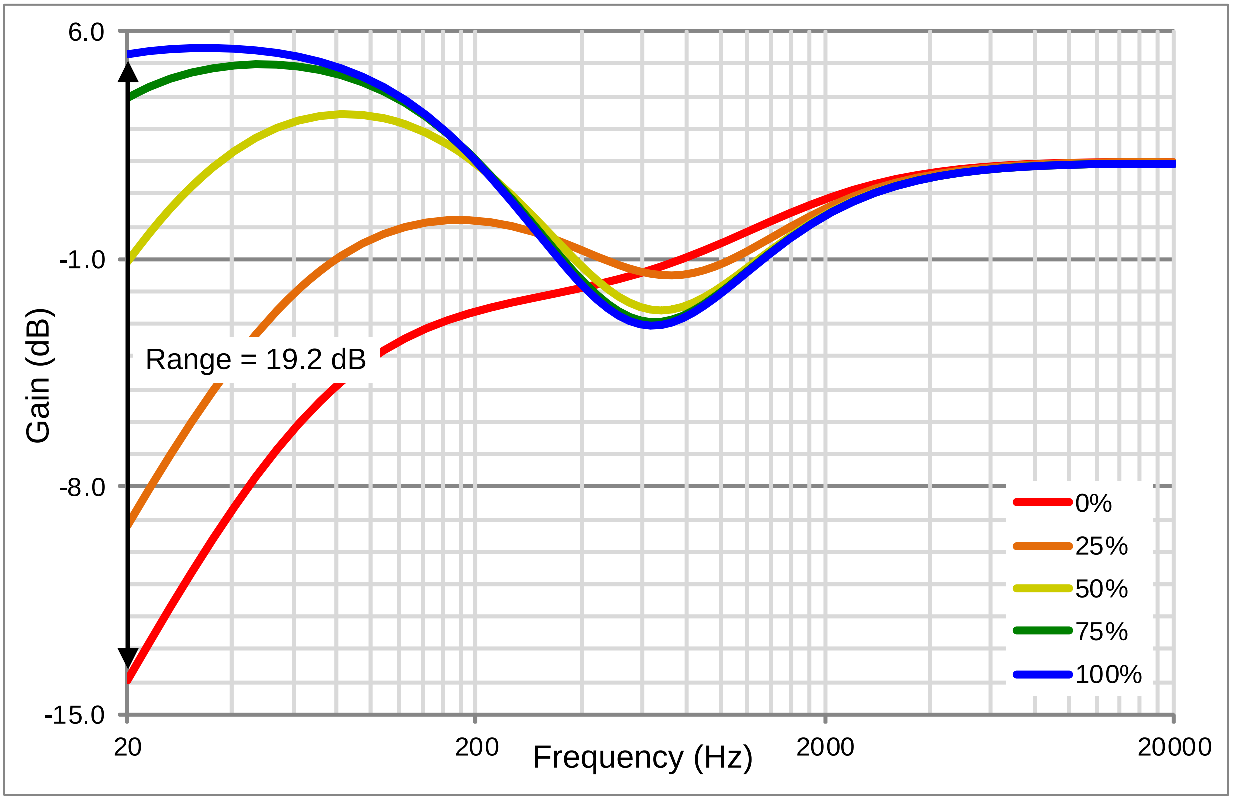


Figure 23: Measured AC Gain Analysis – Bass Control

In this condition, the bass gain varies from -13.9 dB at 0% potentiometer rotation to +5.3 dB at 100% potentiometer rotation. This gives an adjustment range of 19.2 dB, which meets the design requirement of 15 dB.

The result of the measured ac phase analysis under the same conditions is shown in section A.3.

6.4 THD+N Performance

The result of the THD+N measurement over frequency with gain = 6 dB, all potentiometers set to 100%, and mid boost off is shown in Figure 24. The audio analyzer is set to a measurement bandwidth of 90 kHz and no additional filtering or weighting is applied.

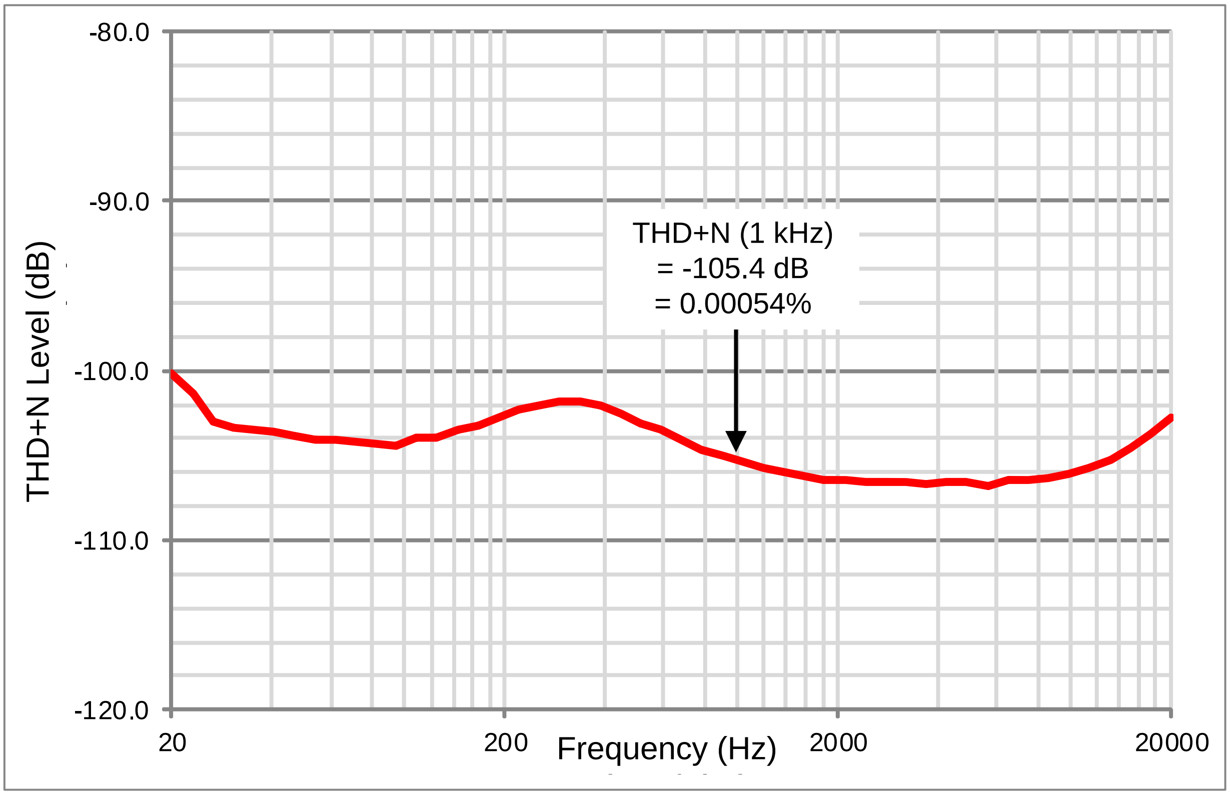


Figure 24: Measured THD+N Level vs. Frequency

This THD+N level at 1 kHz is measured to be -105.4 dB (0.00054%), which meets the design requirement of -100 dB. The THD+N levels at 20 Hz and 20 kHz are measured to be -100.1 dB (0.00099%) and -102.8 dB (0.00072%), respectively.

The Fast Fourier transform (FFT) measurement with V_{IN} at 1 kHz, gain = 6 dB, all potentiometers set to 100%, and mid boost off is shown in Figure 25. The FFT measurement is set to 192k points and 4x averaging.

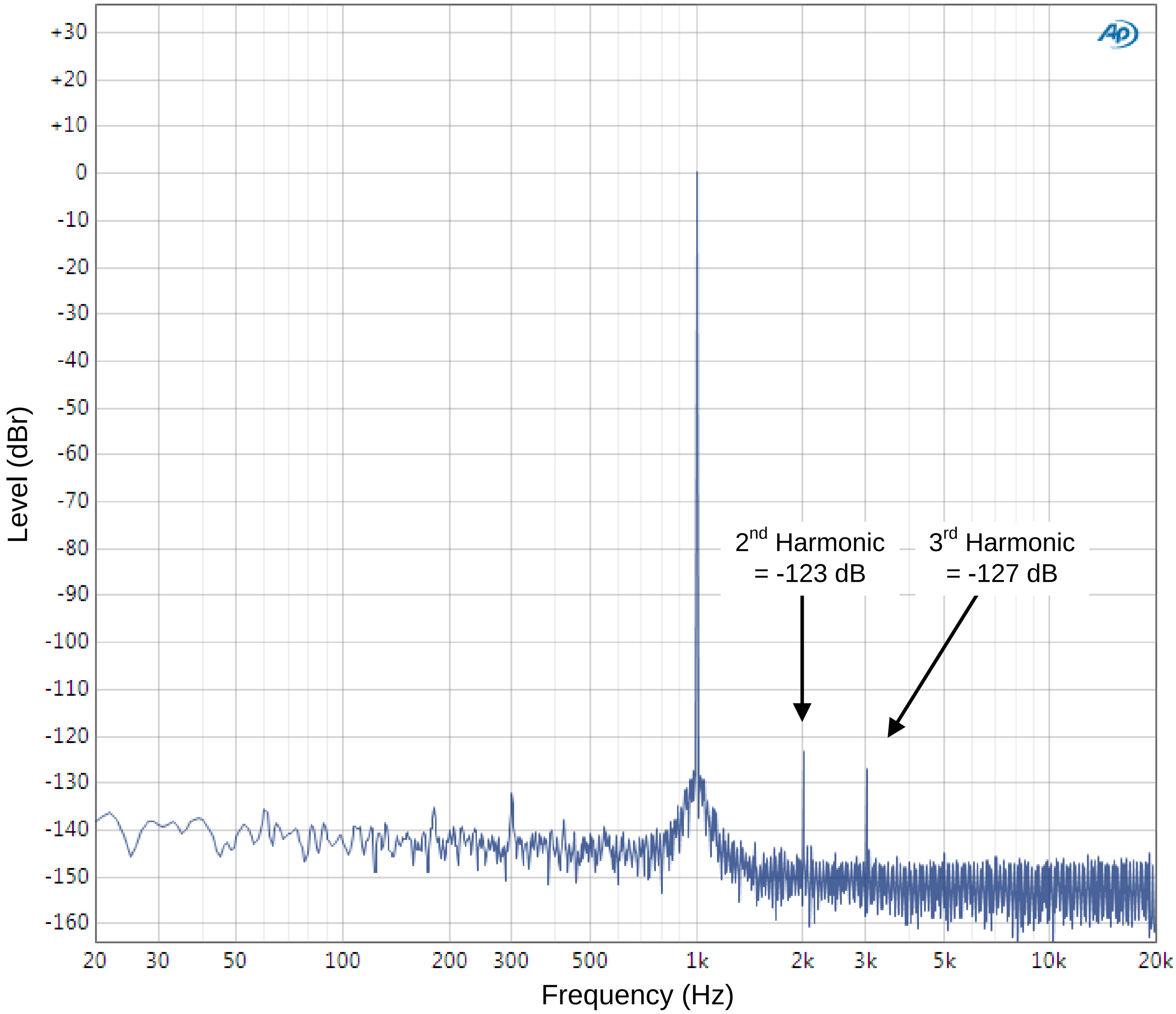


Figure 25: Measured Fast Fourier Transform (FFT)

The y-axis is referenced to the fundamental frequency output level of 1.58 V_{RMS}. The second harmonic is measured at 123 dB below the fundamental, while the third harmonic is measured at 127 dB below the fundamental.

6.5 Measured Results Summary

Table 4 summarizes the measured performance of the design.

Table 4. Comparison of Design Goals and Measured Performance

	Goal	Measured
THD+N ratio at 1 kHz	-100 dB (0.001%)	-105.4 dB (0.00054%)
Treble adjustment range	10 dB	10.6 dB
Mid adjustment range	6 dB	8.9 dB
Bass adjustment range	15 dB	19.2 dB

7 Audio Recordings

While frequency response curves and FFTs can be useful in measuring the performance of a circuit, in audio applications many times “hearing is believing.” The following audio recordings capture the tonal differences between circuit settings as the author plays an E major chord on a Gibson SG through its neck pickup.

Listen online here.

7.1 Audio Recording Downloads

1. All controls 100% vs. all controls 0%: Download
2. Bass 100% vs. bass 0% (mid and treble 50%): Download
3. Mid 100% vs. mid 0% (bass and treble 50%): Download
4. Treble 100% vs. treble 0% (bass and mid 50%): Download
5. Mid boost off vs. mid boost on (all controls 50%): Download

8 Modifications

The components selected for this design were based on the design goals outlined at the beginning of the design process.

This design specifies an input impedance of 6 kΩ. While this is a reasonable specification for passive electric guitar pickups, the actual value will vary across electric guitar and pickup manufacturers. It may be necessary to adjust the values of R_4, C_4 and C_7 in the input filter to achieve the desired cutoff frequencies.

If modifications to the frequency response of the tone stack are desired, the component values of the FMV tone stack may easily be modified using the equations given in section 2.3.2. Duncan’s Tone Stack Calculator is a free software tool which may also be used to model the response of different tone stack topologies, component values and potentiometer settings [8].

A JFET-input amplifier was selected for this application because of the high impedances present in the circuit. The extremely low input bias current (I_B) and input current noise (I_n) of FET-input devices prevent large offset and noise voltages from developing and degrading audio performance.

Among the FET-input audio amplifiers offered by Texas Instruments, the OPA1642 was selected for this application because of its extremely stable input common-mode capacitance which preserves excellent distortion performance even with high source impedances. The OPA1652 is another FET-input audio amplifier with excellent THD+N performance, low noise, and low cost; however its in-circuit distortion performance may be reduced compared to the OPA1642. Table 5 summarizes the key specs between these two devices.

Table 5. Brief Comparison of Audio Operational Amplifiers

Operational Amplifier	THD+N Level at 1kHz	e_N at 1 kHz	I_Q / Channel	Input Type	Approx. Cost / Channel
OPA1642	-126 dB (0.00005%)	5.1 nV/√Hz	1.8 mA	JFET	$0.70 / 1ku
OPA1652	-126 dB (0.00005%)	4.5 nV/√Hz	2.5 mA	CMOS	$0.33 / 1ku

It is often desirable for a gain or volume control to have a response which is linear-in-dB with respect to the rotation of the controlling potentiometer. This results in a very natural change in perceived volume as the user rotates the volume knob.

Because the gain control in this circuit is a potentiometer with a linear taper, the characteristic is not linear-in-dB. If linear-in-dB behavior is desired a potentiometer with an audio taper may be used, or the gain control circuit can be modified to include a Baxandall active volume control as described in TIPD136 [9].

9 About the Author

Ian Williams (ian@ti.com) is an applications engineer in the Precision Analog – Linear team at Texas Instruments where he supports industrial products and applications. Ian graduated from the University of Texas, Dallas, where he earned a Bachelor of Science in Electrical Engineering with a concentration in Microelectronics.

10 Acknowledgements & References

10.1 Acknowledgements

The author wishes to acknowledge John Caldwell for his assistance in the completion of this design.

10.2 References

1. *R. Kuehnel, Circuit Analysis of a Legendary Tube Amplifier: The Fender Bassman 5F6-A, 3rd Edition. Amp Books LLC, 2009.*
2. *M. Blencowe, Designing Tube Preamps for Guitar and Bass, 2nd Edition. Wem Publishing, 2012.*
3. *D. Sorlien and S. Keller. (2009). Introduction to Tube Amplifier Theory Featuring the AX84 P1-eXtreme Amplifier v10.02.15. Available: http://ax84.com/static/p1x/p1-ex-theory.pdf*
4. *R. Kuehnel. (accessed March 2015), Tone Stacks: Fender Bassman 5F6-A vs. Marshall Model 1987. Available: http://www.ampbooks.com/mobile/classic-circuits/bassman-tonestack-intro/*
5. *J. Caldwell (2014). Distortion and Source Impedance in JFET-input Op Amps. Available: http://www.ti.com/lit/an/slyt595/slyt595.pdf*
6. *B. Hofer. (2013, December). Designing for Ultra-Low THD+N (Part 2). Audioxpress, 18-23.*
7. *J. Caldwell. (2013, June 16). Signal distortion from high-K ceramic capacitors. Available: http://www.edn.com/design/analog/4416466/Signal-distortion-from-high-K-ceramic-capacitors*
8. *D. Munro (2008). Duncan's Tone Stack Calculator. Available: http://www.duncanamps.com/tsc/index.html*
9. *I. Williams (2013, December). Active Volume Control for Professional Audio. Available: http://www.ti.com/lit/ug/tidu034/tidu034.pdf*

Appendix A.

A.1 Electrical Schematic

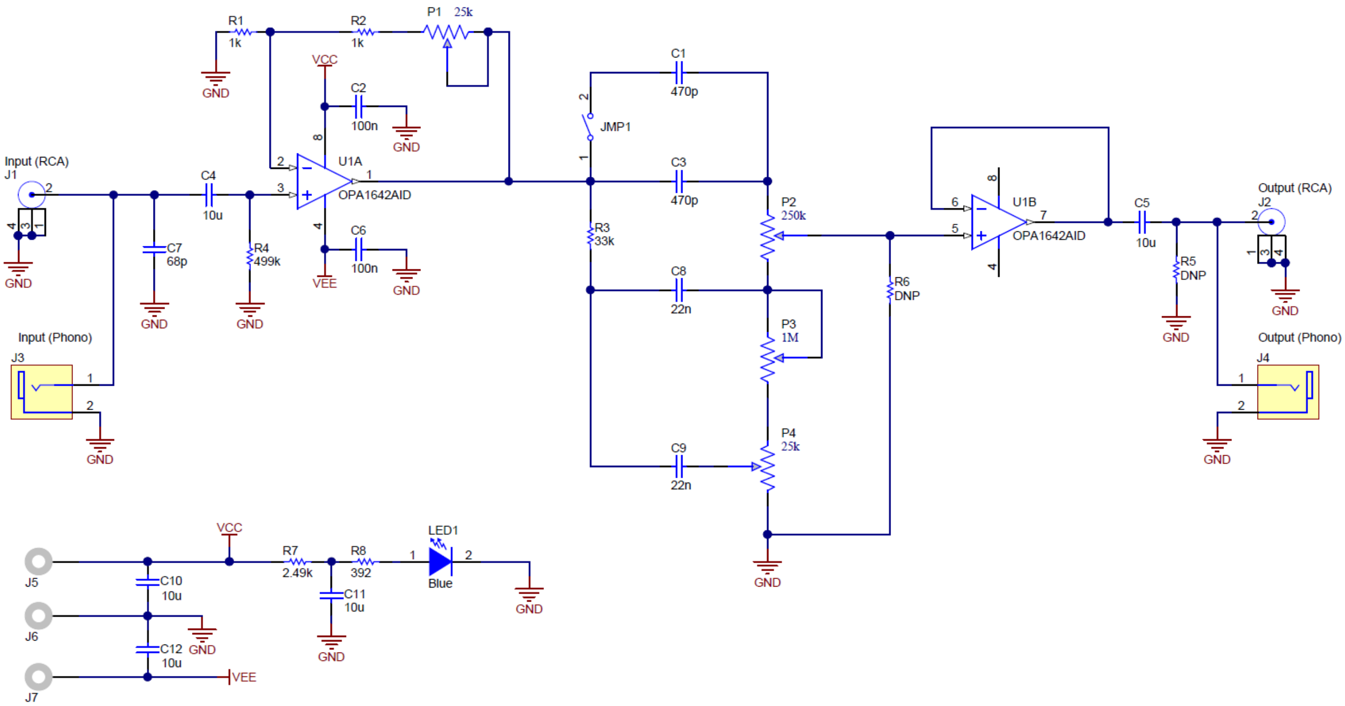


Figure A-1: Electrical Schematic

A.2 Bill of Materials

Item #	Quantity	Value	Designator	Description	Manufacturer	Part Number	Supplier	Supplier Part Number
1	2	470p	C1, C3	CAP CER 470PF 50V 5% NP0 0603	Samsung Electro-Mechanics America	CL10C471JB8NCNC	Digi-Key	1276-2302-1-ND
2	2	100n	C2, C6	CAP CER 0.1UF 50V 10% X7R 0603	Samsung Electro-Mechanics America	CL10B104KB8SFNC	Digi-Key	1276-1936-1-ND
3	5	10u	C4, C5, C10-C12	CAP CER 10UF 35V 10% X7R 1206	Samsung Electro-Mechanics America	CL31B106KLHNNNE	Digi-Key	1276-3103-1-ND
4	1	68p	C7	CAP CER 68PF 50V 5% NP0 0603	Samsung Electro-Mechanics America	CL10C680JB8NCNC	Digi-Key	1276-2324-1-ND
5	2	22n	C8, C9	CAP CER 0.022UF 50V 5% NP0 1206	TDK Corporation	C3216C0G1H223J060AA	Digi-Key	445-7690-1-ND
6	2	1k	R1, R2	RES SMD 1K OHM 1% 1/10W 0603	Panasonic Electronic Components	ERJ-3EKF1001V	Digi-Key	P1.00KHCT-ND
7	1	33k	R3	RES SMD 33K OHM 1% 1/10W 0603	Panasonic Electronic Components	ERJ-3EKF3302V	Digi-Key	P33.0KHCT-ND
8	1	499k	R4	RES SMD 499K OHM 1% 1/10W 0603	Panasonic Electronic Components	ERJ-3EKF4993V	Digi-Key	P499KHCT-ND
9	1	2.49k	R7	RES SMD 2.49K OHM 1% 1/8W 0805	Panasonic Electronic Components	ERJ-6ENF2491V	Digi-Key	P2.49KCCT-ND
10	1	392	R8	RES SMD 392 OHM 1% 1/8W 0805	Panasonic Electronic Components	ERJ-6ENF3920V	Digi-Key	P392CCT-ND
11	2	25k	P1, P4	25K LINEAR POT RIGHT ANGLE PC MOUNT	Alpha	RV16AF-41-15R1-B25K	Mammoth Electronics	210-100-B-25K
12	1	250k	P2	250K LINEAR POT RIGHT ANGLE PC MOUNT	Alpha	RV16AF-41-15R1-B250K	Mammoth Electronics	210-100-B-250K
13	1	1M	P3	1M LOG POT RIGHT ANGLE PC MOUNT	Alpha	RV16AF-41-15R1-A1M	Mammoth Electronics	210-100-A-1M
14	3	KNOB	N/A	BLK ALUM KNURLED KNOB 15MM X 13MM	Unknown	4SKA-15X13KBK	Mammoth Electronics	4SKA-15X13KBK
15	1	LED	LED1	LED BLUE CLEAR 3MM 470NM	KingBright	WP710A10QBC/D	Digi-Key	754-1596-ND
16	1	OPA1642	U1	OPA1642 SOUND-PLUS OP AMP	Texas Instruments	OPA1642AID	N/A	N/A
17	3	BANANA	J5-J7	CONN JACK BANANA UNINS PANEL MOU	Emerson Network Power	108-0740-001	Digi-Key	J147-ND
18	2	RCA	J1, J2	CONN RCA JACK METAL R/A BLK PCB	CUI Inc	RCJ-011	Digi-Key	CP-1400-ND
19	2	PHONO PLUG	J3, J4	CONN JACK PHONE 1/4" 2POS RA OPN	Switchcraft Inc.	RN111PC	Digi-Key	SC1121-ND
20	1	JUMPER	JMP1	CONN HEADER 2POS .100" SGL GOLD	Samtec	TSW-102-07-G-S	Digi-Key	SAM1029-02-ND
21	1	JUMPER SHUNT	JMP1	JUMPER SHORTING UNITS	TE Connectivity	881545-2	Digi-Key	A26242-ND
22	4	SCREW	Screws	SCREW MACHINE PHILLIPS 4-40X3/8	B&F Fastener Supply	PMS 440 0038 PH	Digi-Key	H781-ND
23	4	STANDOFF	H1-H4	STANDOFF HEX 4-40THR ALUM .500"L	Keystone Electronics	2203	Digi-Key	2203K-ND
24	1	DNP	R5, R6	DO NOT POPULATE	n/a	N/A	N/A	N/A

Figure A-2: Bill of Materials

A.3 Phase Response Measurements

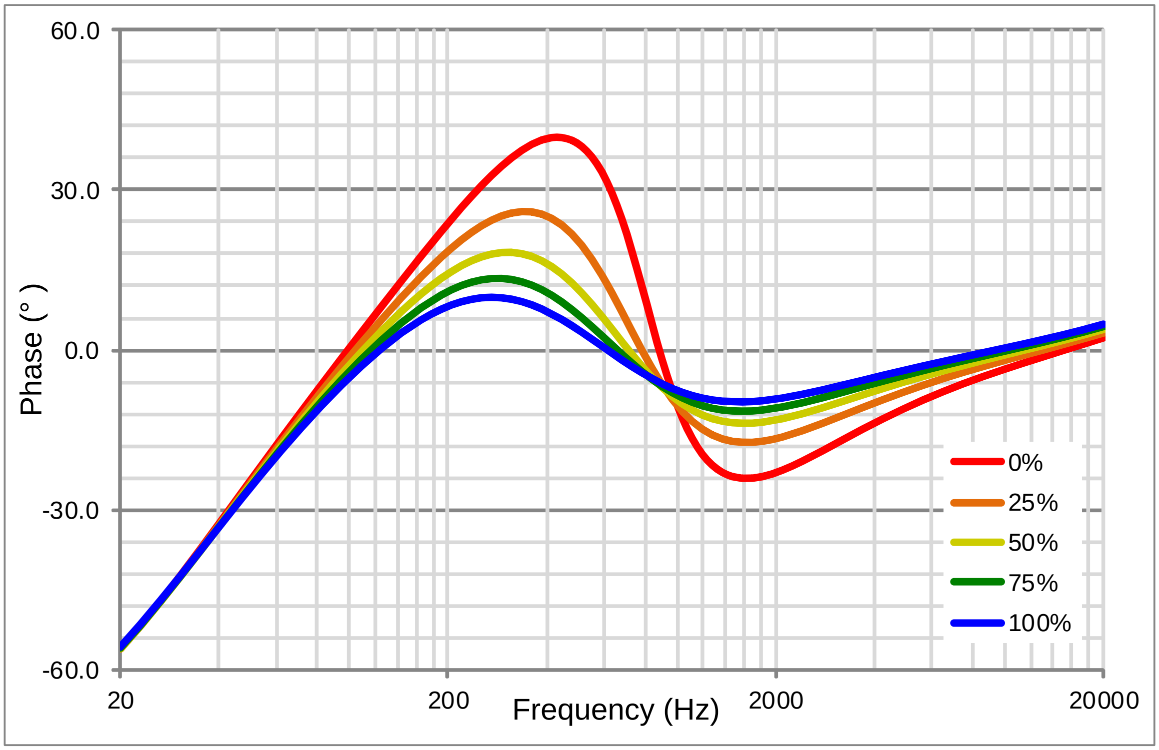


Figure A-3: Measured AC Phase Analysis – Treble Control

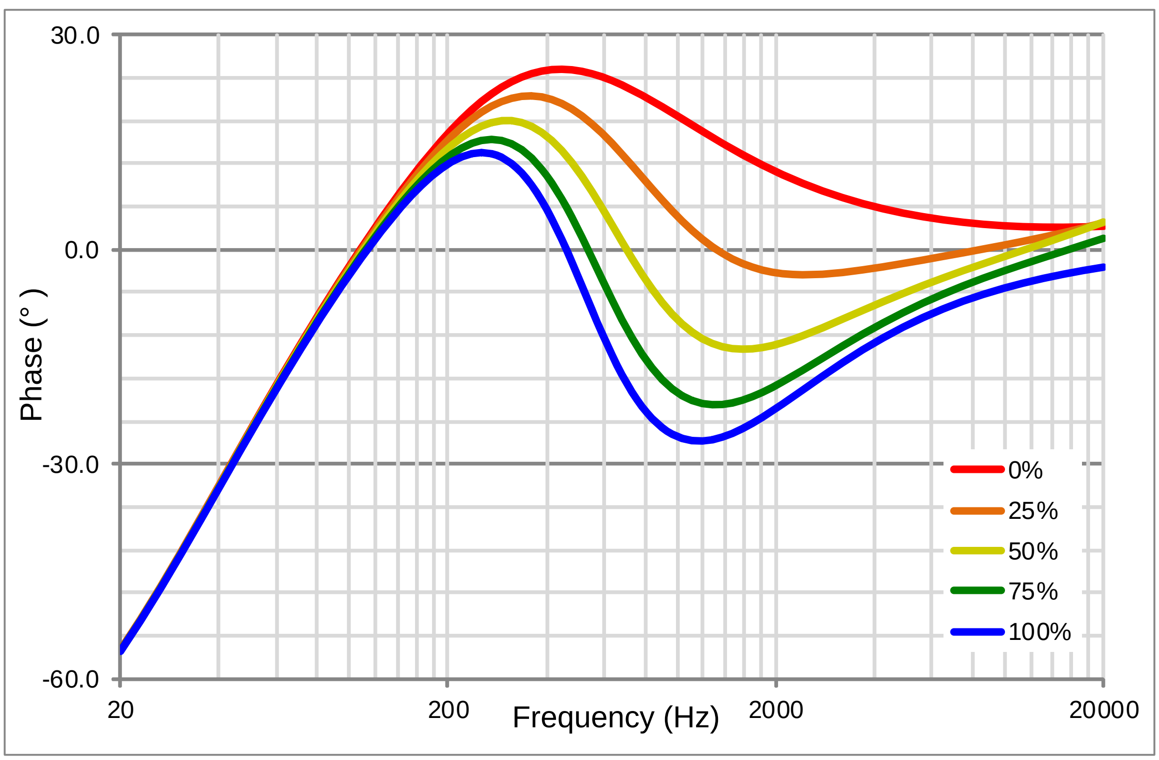


Figure A-4: Measured AC Phase Analysis – Mid Control

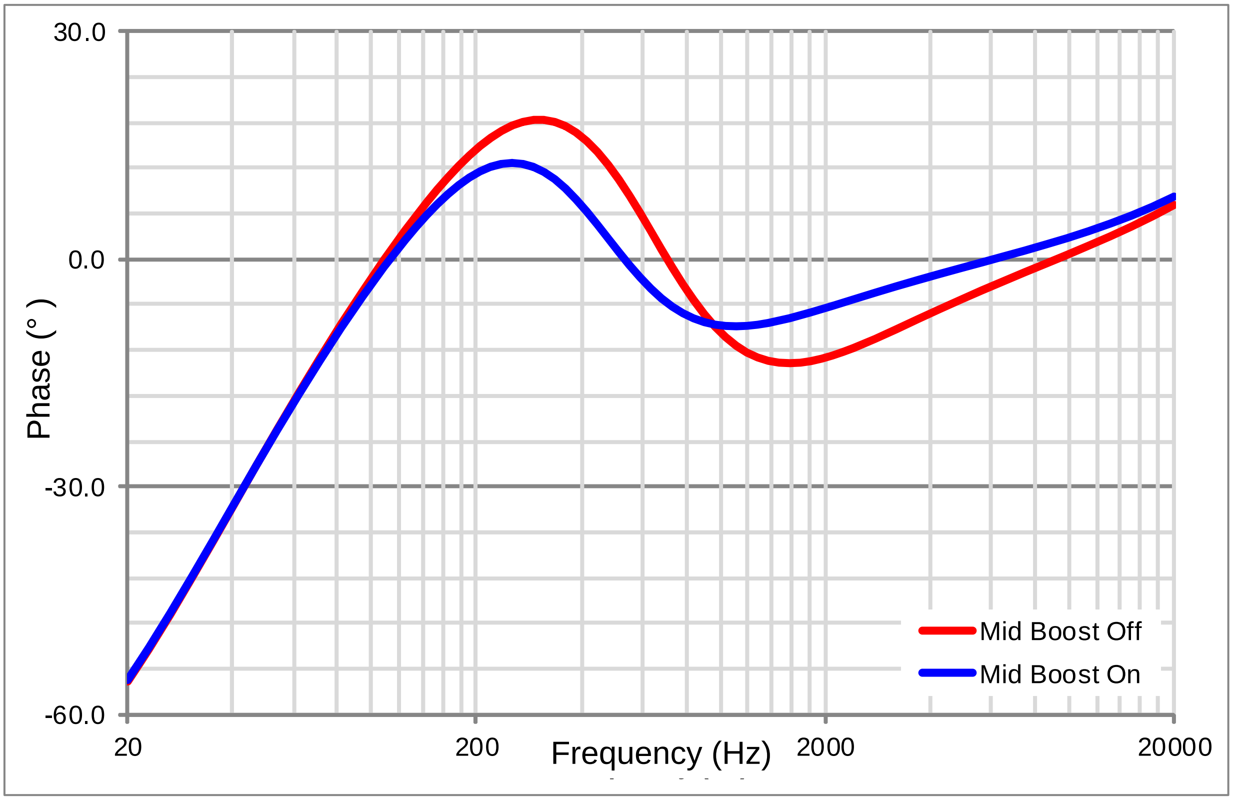


Figure A-5: Measured AC Phase Analysis – Mid Boost

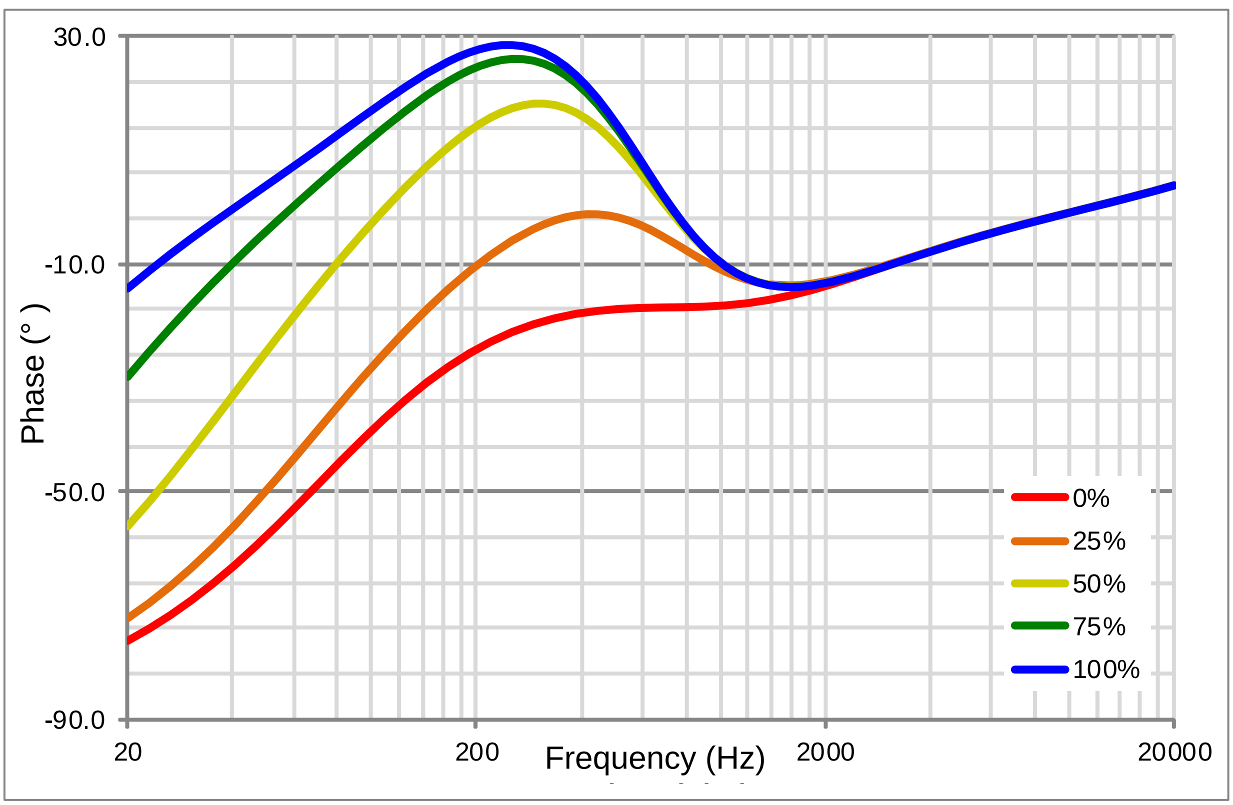


Figure A-6: Measured AC Phase Analysis – Bass Control

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US005524055A

United States Patent [19]

Sondermeyer

[11] Patent Number: 5,524,055

[45] Date of Patent: Jun. 4, 1996

[54] **SOLID STATE CIRCUIT FOR EMULATING TUBE COMPRESSION EFFECT**

[75] Inventor: **Jack C. Sondermeyer**, Meridian, Miss.

[73] Assignee: **Peavey Electronics Corporation**, Meridian, Miss.

[21] Appl. No.: **182,493**

[22] Filed: **Jan. 18, 1994**

[51] **Int. Cl.**[6] **H03G 3/08**

[52] **U.S. Cl.** **381/61**

[58] **Field of Search** 381/61; 330/51, 330/255, 262

[56] **References Cited**

U.S. PATENT DOCUMENTS

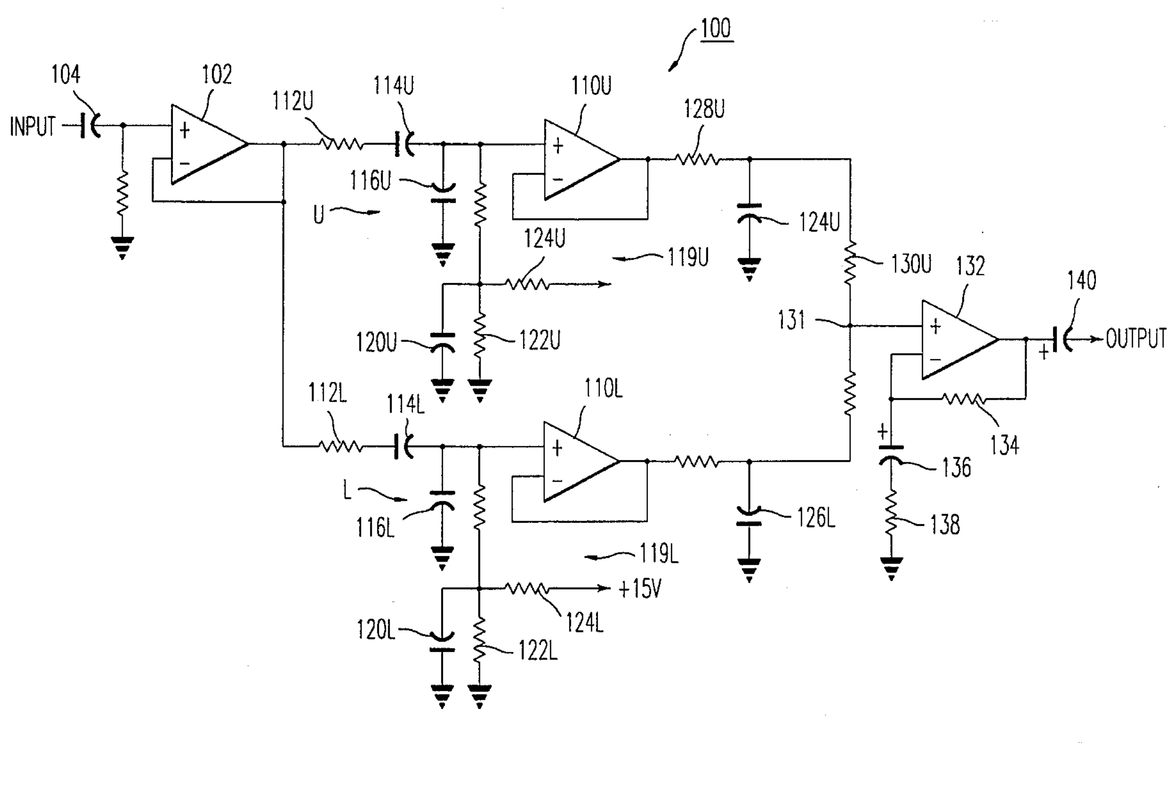
4,405,832	9/1983	Sondermeyer	179/1 D
4,439,742	3/1984	Sondermeyer	330/262
4,809,336	2/1989	Pritchard	381/61
4,937,847	6/1990	Kruger	378/62
4,987,381	1/1991	Butler	330/255
4,995,084	2/1991	Pritchard	379/94
5,012,199	4/1991	McKale	330/51
5,133,014	7/1992	Pritchard	381/61

Primary Examiner—Stephen Brinich
Attorney, Agent, or Firm—Watson Cole Stevens Davis

[57] **ABSTRACT**

A solid state amplifier for emulating the compression associated with an overbiased class-B push-pull tube amplifier at high input signal levels due to the flow of current into the grid of the output tubes resulting in a desirable output clipping characteristic with crossover distortion is disclosed. The invention includes at least one pair of class-B connected solid state devices, each having an input circuit and an output circuit. The output circuits are connected for mixing. A biasing element in the input circuit of each paired solid state device establishes a clipping level offset at the input circuit and at the output circuit of each device. A clipping element in the input circuit and the output circuit clips the offset at the input circuit and clips the offset at the output circuit of each respective solid state device. A charging element overbiases the offset in the input circuit whenever the input signal is greater than the input clipping element. The overbiasing causes crossover distortion for emulating the desirable compression associated with a tube amplifier.

26 Claims, 3 Drawing Sheets



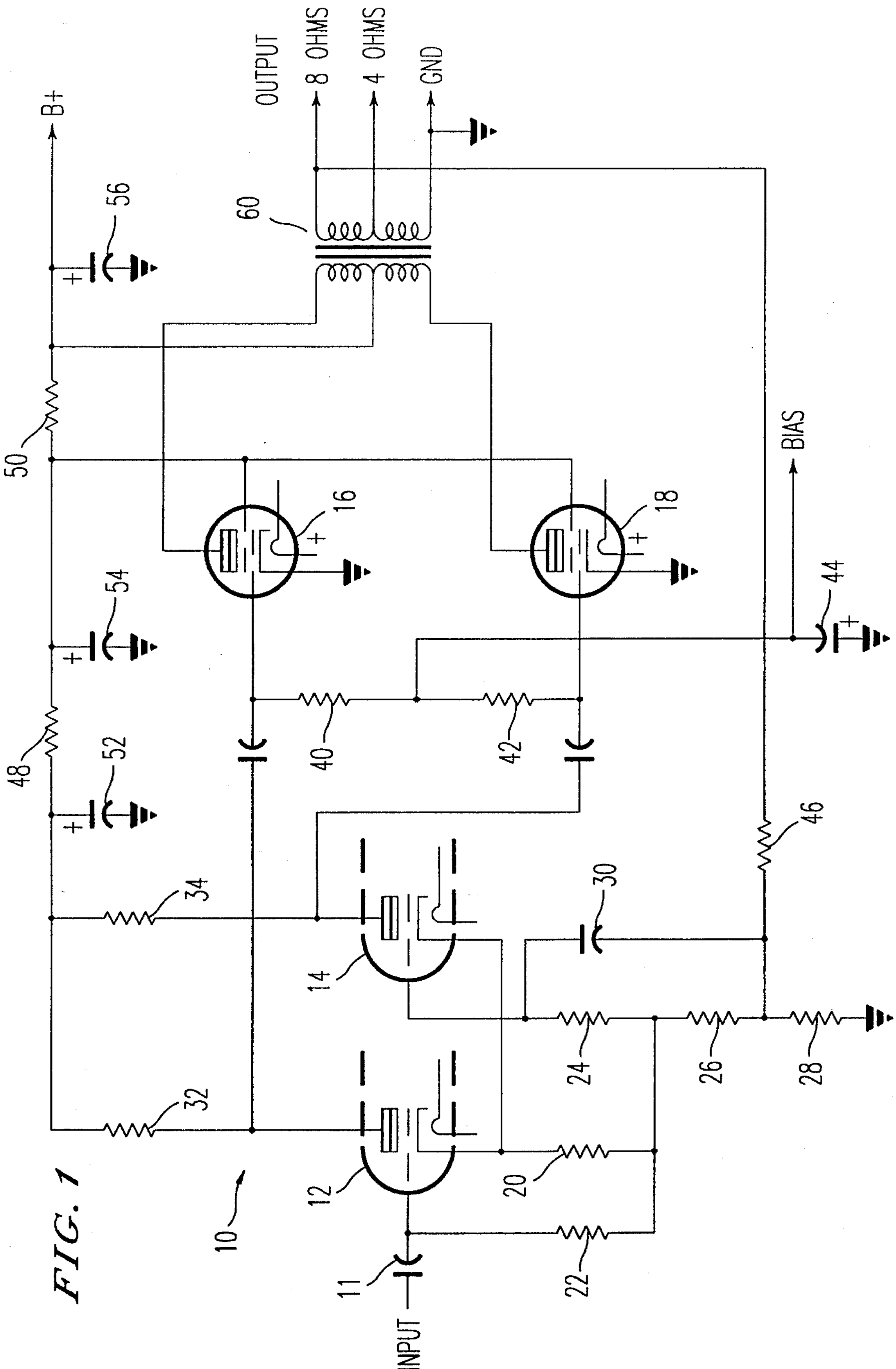
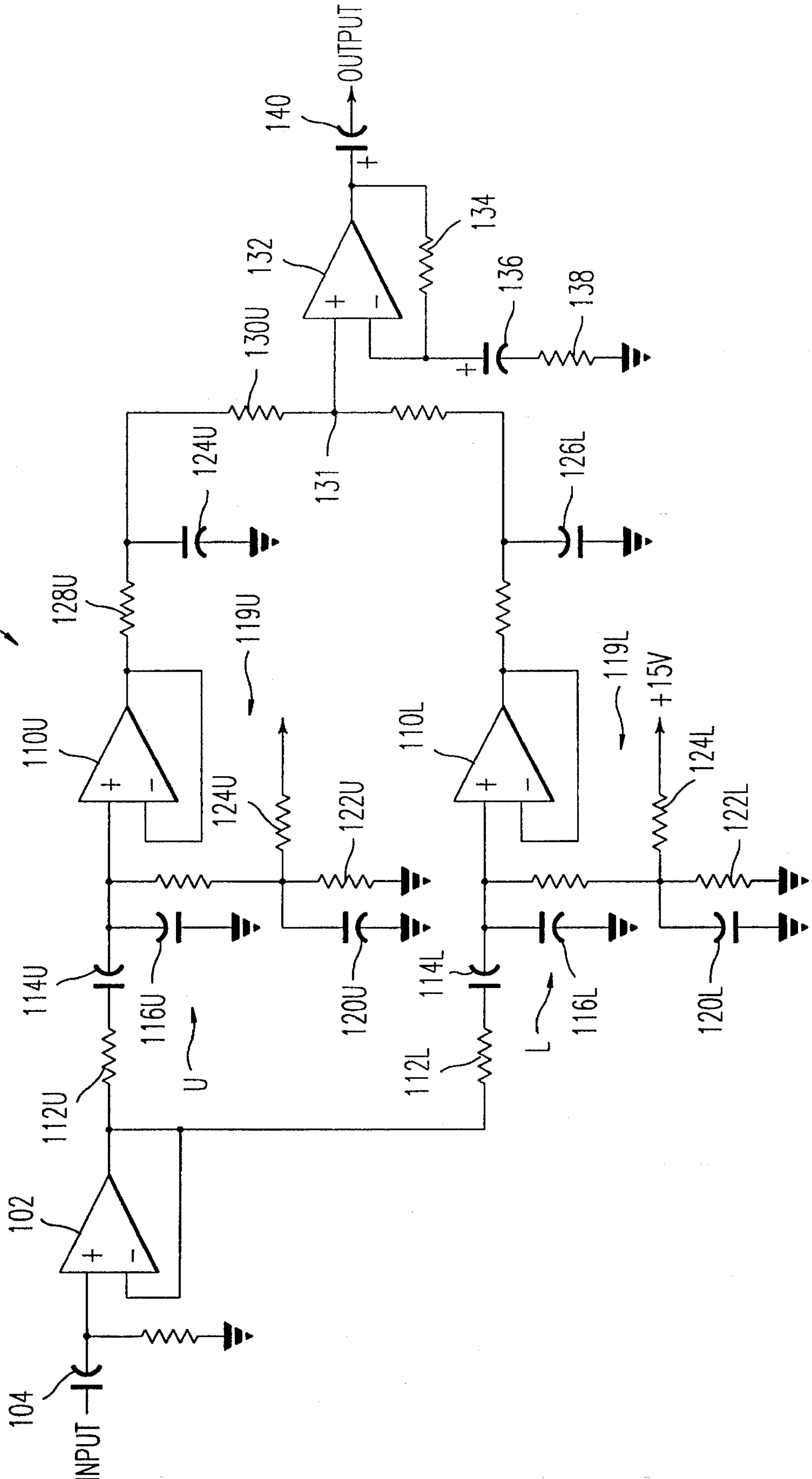

B+
OUTPUT
8 OHMS
4 OHMS
GND
BIAS
INPUT
10
11
12
14
16
18
20
22
24
26
28
30
32
34
40
42
44
46
48
50
52
54
56
60

FIG. 1

FIG. 2



100
INPUT
104
102
112U
114U
116U
U
120U
122U
124U
110U
128U
119U
124U
130U
131
132
134
136
138
140
OUTPUT
112L
114L
116L
L
120L
122L
124L
+15V
119L
110L
126L

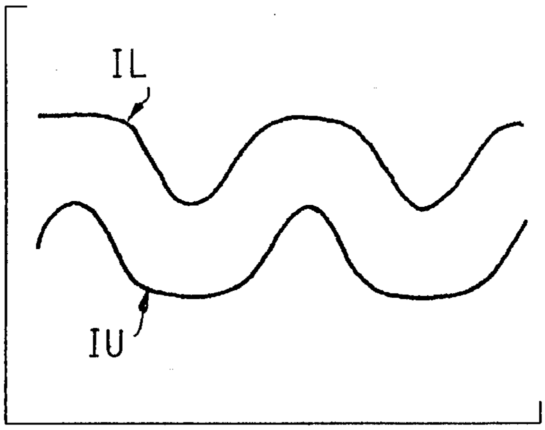


FIG. 3A

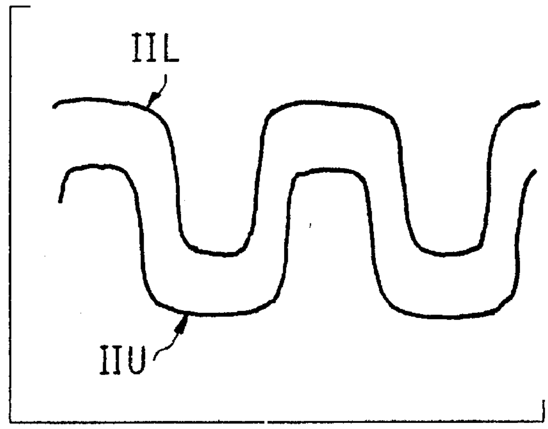


FIG. 3D

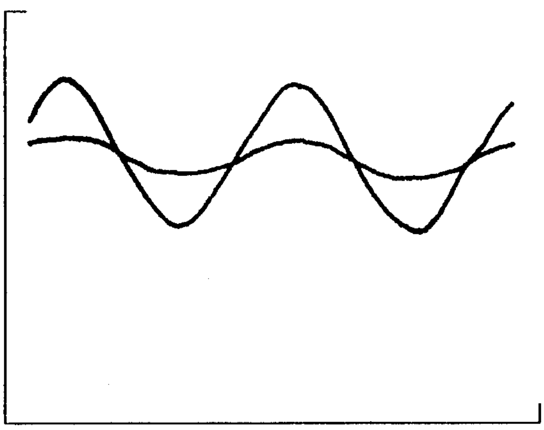


FIG. 3B

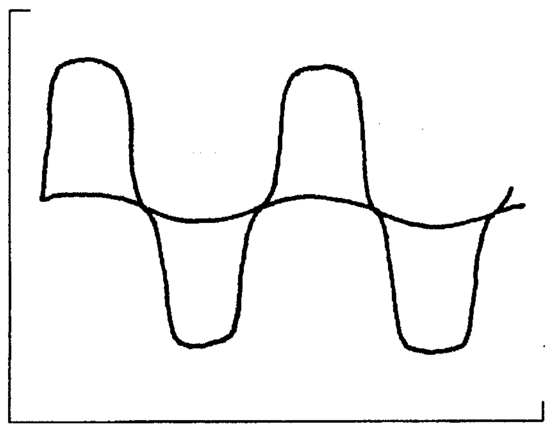


FIG. 3E

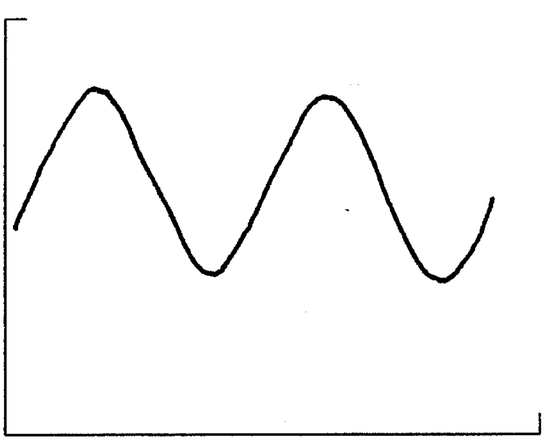


FIG. 3C

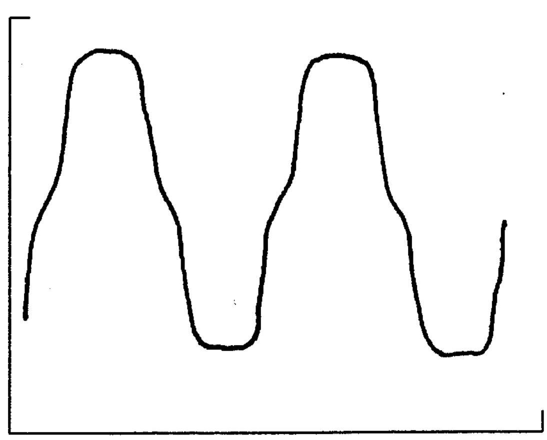


FIG. 3F

SOLID STATE CIRCUIT FOR EMULATING TUBE COMPRESSION EFFECT

BACKGROUND OF THE INVENTION

The invention relates to replacement of tubes in power amplifiers with solid state devices. In particular, the invention is directed to a solid state circuit that duplicates tube power amplifier compression.

Tube compression occurs whenever the tube power amplifier is driven into hard clipping. Normally, a solid state amplifier driven into hard clipping creates harsh odd-order harmonic distortion (square waves). In contrast, a tube amplifier compresses the signal so that the level decreases and it does not sound as harsh and strident. As a result, the sound is more subdued, but still has what the players call "punch". Thus, compression is a musical function that gives a tube power amplifier an edge over conventional solid state power amplifiers according to most heavy metal and bass guitar players, particularly at clipping conditions.

The foregoing is a non-technical description of a phenomenon called increased crossover distortion. This function happens in all tube power amplifier designs whenever the output tube grid is driven positive with respect to the cathode causing it to become simply a forward biased diode.

In a typical push-pull configuration, using two class-B biased tubes, the diode in each push-pull output stage causes the average bias level to increase at high signal levels and forces the class-B biased tubes to become over biased. Such condition causes the output signal to have severe crossover distortion, a condition where the signal zero crossing is delayed significantly.

A typical tube power amplifier **10** which has been used on many popular models, is shown in FIG. **1**. Typical circuit operation is described below followed by a description of overload (or tube compression) conditions.

In FIG. **1**, input signals are coupled via coupling capacitor **11** to the grid of vacuum tube **12** (e.g., 12AX7), which with tube **14** is half of what is called a long tailed phase inverter circuit. In this circuit, the cathodes of tubes **12** and **14** are connected together, as shown. Thus, tube **12** operates in a grounded cathode mode; while tube **14** operates in a grounded grid mode with respect to the input grid of tube **12**. Accordingly, equal but out-of-phase signals appear at the plates of **12** and **14**. The purpose of the phase inverter is to supply two out-of-phase signals to class-B biased push-pull output tubes **16** and **18**.

Cathode resistor **20** sets the bias for each tube **12** and **14**. Grid resistors **22** and **24** are the respective grid bias resistors. Resistor **26** is a common cathode resistor. Resistor **28** is used to introduce feedback from the output to reduce overall distortion. The grid of tube **14** is shunted to ground (in this case, the low impedance feedback point) via capacitor **30**, as is necessary for grounded grid operation. Load resistors **32** and **34** are the respective plate loads for tubes **12** and **14**. The plate signals are coupled to the output tubes **16** and **18** via capacitors **36** and **38**.

Each output tube grid is connected to a negative bias source (e.g., –55 V) via bias resistors **40** and **42**. This –55 V sources is generated externally from this circuit and is filtered adequately by capacitor **44**. Negative 55 volts is chosen as the appropriate value to bias the output tubes **16** and **18** (e.g., 6L6GC) into good class-B operation with minimal crossover distortion at low signal levels.

Completing the circuit, resistor **46** is a feedback resistor; resistors **48** and **50** are power supply decoupling resistors; capacitors **52**, **54** and **56** are filter capacitors for the various supply sources in the B+circuit. Finally, transformer **60** is a conventional tube push-pull output transformer, in this case with output taps for 8 and 4 ohms. The power amplifier **10** delivers approximately 50 WRMS to the matching load value.

At all signal levels below output clipping (the output waveform being clean and free of distortion), the signal levels at the grid of each output tube **16** and **18** is well below 55 volts peak swing, and the average DC bias level at each output tube grid is –55 VDC. However, at clipping and beyond, the signal levels at each output tube grid will exceed 55 volts peak swing. Thus, the grid will be biased positive with respect to the cathode at each positive peak signal swing. Whenever the grid is driven positive with respect to the cathode, it becomes a simple forward biased diode. With the positive peak swing clipped, the average negative DC bias voltage level at the grid of each output tube **16** and **18** is increased in proportion to the overload input value above the clipping value. Thus, the output tubes **16** and **18** become over biased beyond class-B and at severe output clipping significant crossover distortion is generated as well. Consequently, at overload, the output signal of tube amplifier **10** will be clipped at the peaks. However, it will not be as "dirty" as a typical solid state power amplifier operating under the same conditions, because a large portion of the overloaded output waveform is forced or compressed into the severe crossover distortion region. To a musician, such a waveform is much more musical in nature and "cleaner" (i.e., less harsh) than a solid state amplifier at overload. Due to the compression (i.e., distortion near the zero crossover), the actual peak output clipping is reduced and is far more tolerable than that of the solid state amplifier. This phenomenon is thus, tube power amplifier compression.

SUMMARY OF THE INVENTION

The present invention is directed to a solid state amplifier for emulating the compression associated with an overbiased class-B push-pull tube amplifier at high input signal levels due to the flow of current into the grid of the output tubes resulting in a desirable output clipping characteristic with crossover distortion. The invention includes at least one pair of class-B connected solid state devices. Each device has an input circuit and an output circuit. The output circuits are connected for mixing. Biasing means in the input circuit of each paired solid state device establishes a clipping level offset at the input circuit and at the output circuit of each device. Clipping means in the input circuit and in the output circuit clips the offset at the input circuit and the offset at the output circuit of each respective solid state device. Charging means overbiases the offset in the input circuit whenever the input signal is greater than said input clipping means, said overbiasing causing crossover distortion for emulating the desirable compression associated with a tube amplifier.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. **1** is a schematic diagram of a known class-B tube amplifier circuit;

FIG. **2** is a schematic diagram of a solid state amplifier which emulates tube compression in accordance with the present invention;

FIGS. **3A–3F** are waveform diagrams illustrating tlube amplifier compression emulation of the present invention.

DESCRIPTION OF THE INVENTION

A solid state emulator **100** of the invention is shown in FIG. **2**. Input signal is coupled to an operational amplifier (OP AMP) **102** via coupling capacitor **104** with resistor **106** providing a reference to ground. The output of amplifier **102** drives upper and lower circuits U and L including class-B biased, push-pull connected emulator operational amplifiers **110**U and **110**L. Each **0**P AMP circuit **110**U and **110**L is a unity gain stage that duplicates one of the output tubes **16** and **18** in the push-pull tube power amplifier **10** shown in FIG. **1**. The OPAMP emulator circuits **110**U and **110**L are identical except for the diode directions discussed hereinafter. Thus, the reference numbers and designations U and L will be used only where necessary. The upper circuit U is discussed below followed by a discussion of the differences in the lower circuit L.

In the upper circuit U, the output of amplifier **102** is coupled to amplifier **110** via resistor **112** and capacitor **114**. A diode **116** is coupled to ground at the input of amplifier **110**. A resistor **118** is coupled to an upper bias circuit **119** comprising the parallel combination of diode **120** and resistor **122** to ground, in series with resistor **124** to the −15 volt supply. The output of amplifier **110** is applied to diode **126** via resistor **128**. The signal at diode **126** (i.e., the output of the upper circuit U) is mixed with signal from the lower circuit L via resistors **130**U and **130**L. The mixed outputs are then amplified by output amplifier **132** which is a non-inverting gain stage with a feedback resistor **134**, a ground circuit including capacitor **136** and series resistor **138**, and output coupling capacitor **140**. In order to provide a greater offset voltage the diodes **120** and **126** may be multiple diodes in series (not shown).

In the exemplary embodiment illustrated, the upper bias circuit **119** creates −0.6 volts at the cathode of diode **120**), and this bias is applied to the input of amplifier **110** via resistor **118**. This −0.6 volt input bias offsets the output of amplifier **110** at the same amount. Further, this offset is applied to diode **126** through resistor **128**. Thus, output circuit diode **126** is biased into slight forward conduction at idle. The lower emulator circuit L is identical to the upper circuit U except that the direction of diodes **116**L, **120**L and **126**L are reversed or complimentary to the diodes **116**U, **120**U and **126**U. All other elements are the same.

A low level input signal, e.g., a 1 volt peak sine wave, is coupled in the upper circuit U via resistor **112** and capacitor **114** to the input circuit of amplifier **110**U. The input is offset −0.6 VDC. The applied signal has a negative peak value of −1.6 volts and a positive peak value of +0.4 volts. Diode **116**, whose cathode is at ground, is reversed biased at the negative peak swing, and is forward biased at the positive peak swing. However, diode **116** does not conduct in the forward direction because the peak swing is only +0.4 volts and diode conduction begins at +0.6 volts. The same signal swing occurs at the output of amplifier **110** because it has a unity gain. The output signal is then applied to diode **126** in the output circuit, which as noted above, is already biased at idle into a slight forward conduction. Hence, diode **126** clips the negative swing because it is forward biased for this swing, and it allows the positive swing to pass, because it is biased below 0.6 volts forward and is in effect ultimately reverse biased. The resulting waveform is shown in FIG. **3**A as curve IU. The waveform is a clean half sine wave in the positive direction and a clipped half sine wave in the negative direction.

The lower emulator circuit L using lower amplifier **110**L is identical except all the diodes are reversed and lower the bias circuit **119**L consisting of diode **120**L, and resistors **118**L, **122**L and **124**L therein creates +0.6 volts at the anode of diode lower **126**L (0.6 volts being the typical forward drop of the diode). In the lower circuit L the bias is applied to the input of lower amplifier **110**L via resistor **118**L. This +0.6 volt input bias then also offsets the output of amplifier **110**L by the same amount. Further, this offset is applied to diode **126** through resistor **128**. Thus, diode **126** is biased into slight forward conduction at idle. A 1 volt peak sine wave applied to this lower emulator circuit L is thus opposite the upper emulator circuit U. As a result, a clean half sine wave is produced in the negative direction and a clipped half sine wave is produced in the positive direction. This waveform is shown in FIG. **3**A as curve IL. The two emulated waveforms IL and IU are mixed together at node **131** creating a relatively clean sine wave as shown in FIG. **3**C. To appreciate how these combine FIG. **3**B shows IL and IU superimposed.

At high level signals in the upper circuit U, e.g., at a 3 volt peak sine wave, the input signal is coupled via resistor **112** and capacitor **114** to the input of upper amplifier **110**. The input is offset −0.6 VDC. If diode **116** were not present, the applied signal would have a negative peak value of −3.6 volts and a positive peak value of +2.4 volts. However, with diode **116** present and with its cathode grounded, it is reversed biased at the negative peak swing, and forward biased at the positive peak swing. Thus, diode **116** conducts in the forward direction because the peak swing is greater than +0.6 volts. Accordingly, diode **116** limits the peak swing to +0.6 volts and clips the positive waveform somewhat. Capacitor **114** charges in the negative direction to allow the 3 volt peak sine wave to pass with a positive peak value of 0.6 volts and a negative peak value of approximately −4.6 volts. At this condition, the average bias is −1.6 VDC rather than −0.6 VDC. Hence, the upper emulator circuit U is over-biased for these signal conditions.

As noted above, the same signal swing occurs at the output of amplifier **110** as is at the input, because the amplifier is a unity gain stage. This signal is then applied to output diode **126**, which is already biased at zero crossing into a heavy forward conduction due to the over-biased conditions. Hence, diode **126** clips the negative swing, (because it is forward biased for this swing) and it clips that portion of the positive swing for which it is over-biased. Diode **126** then allows the remaining positive swing to pass because it is biased below 0.6 volts forward and then is ultimately reverse biased. The resulting signal is thus asymmetrical, having spent more time in the negative swing than the positive swing. This waveform is shown in FIG. **3**D as curve II U. The signal is a partial clipped half sine wave in the positive direction and a fully clipped half sine wave in the negative direction with significant asymmetry.

The lower emulator circuit L using lower amplifier **110** is identical except all the diodes are reversed. Thus, it should be clear that a 3 volt peak sine wave applied to the lower emulator circuit L will be opposite the upper one. A partially clipped half sine wave in the negative direction and a fully clipped half sine wave in the positive direction with significant asymmetry results. This waveform is shown in FIG. **3**D as curve IIL. Mixing these two emulated waveforms together at node **131** creates a clipped sine wave with considerable crossover distortion as shown in FIG. **3**F. To appreciate how these combine, FIG. **3**E shows the signals IIU and IIL superimposed.

It is useful to point out the components in the circuits of FIGS. **1** and **2** that perform the same functions or act in the same manner:

1: Resistors **32** and **34** (Tube) and Resistors **112**U and **112**L (SS) are source resistors for the clipping function.

2: **36**, **38** (Tube) and **114**U, **114**L (SS) are the coupling capacitors that charge to overbias.

3: **40**, **42** (Tube) and **118**L, **118**U (SS) are the bias source resistors.

4: Grid of **16**, grid of **18** (Tube) and diodes **116**U, **116**L (SS) provide the input clipping mechanism.

5: **16**, **18** in push/pull (Tube) and diodes **126**L, **126**U (SS) correspond as follows, in the tube amplifier, each output tube supplies one polarity signal swing to the output. In the solid state amplifier, the diodes remove the unwanted polarity output swing. In the tube amplifier, the input signal is split into two out-of-phase signals to drive identical output tubes in push-pull via the output transformer. In the solid state amplifier, identical input signals are applied to two emulators which are polarity reversed, and the output signals are summed.

Finally, tube compression has a certain attack and decay which is how fast the compression happens and how long it takes to stop. The solid state emulator **100** acts in a similar manner. Additionally, depending upon input waveform, different overbias conditions can occur on each signal half cycle in the tube amplifier. Similarly, the solid state emulator **100** can also overbias in a similar manner.

While there have been described what are at present considered to be the preferred embodiments of the present invention, it will be apparent to those skilled in the art that various changes and modifications may be made therein without departing from the invention, and it is intended in the appended claims to cover such changes and modifications as fall within the spirit and scope of the invention.

What is claimed is:

1. A solid state amplifier for emulating the compression effect associated with an overbiased class-B push-pull tube amplifier at high input signal levels due to a flow of current into the grid of the output tubes resulting in a desirable output clipping characteristic with crossover distortion comprising:

at least one pair of class-B connected solid state devices, each including an input circuit and output circuit, the output circuit and said at least one pair being connected for mixing;

biasing means in the input circuit of each paired solid state device for establishing an offset level at the input circuit and the output circuit of each device;

clipping means in the input circuit and in the output circuit each having a respective clipping level relative to the offset level for clipping signals at the input circuit and clipping signals at the the output circuit of each solid state device;

charging means for overbiasing the input circuit whenever the input signal is greater than said input clipping level, said overbiasing causing crossover distortion, emulating the desirable compression effect associated with a tube amplifier.

2. The amplifier of claim **1** wherein the biasing means in the input circuit of said pair of solid state devices comprise complimentary diodes.

3. The amplifier of claim **1** wherein the biasing means in the input circuit of each solid state device includes a diode and a resistor network coupled to the input circuit.

4. The amplifier of claim **1** wherein the clipping means in the input circuit and the output circuit comprise complimentary connected diodes.

5. The amplifier of claim **1** wherein the solid state devices comprise operational amplifiers.

6. The amplifier of claim **1** wherein the charging means comprises a resistor capacitor network in the input circuit of each solid state device.

7. The amplifier of claim **1** further comprising input amplifier means commonly coupled to the input circuit of each solid state device.

8. The amplifier of claim **1** further comprising output amplifier means commonly coupled for receiving the mixed output of the output circuits.

9. The amplifier of claim **1** wherein the offset level at the input equals at least one diode voltage drop and the offset in the output circuit equals said at least one diode voltage drop.

10. The amplifier of claim **1** wherein the gain of each solid state device is unity.

11. The amplifier of claim **1** wherein the clipping means in the input clips signals greater than the clipping level plus the offset level.

12. The amplifier of claim **1** wherein the clipping means in the output circuit clips unused opposite half cycles of the input signals.

13. A solid state amplifier comprising:

a pair of solid state devices, each having an input circuit and an output circuit, the output circuits being connected for mixing;

offset means in the input circuit of each solid state device for establishing an offset level at the input circuit and at the output circuit of each device;

clipping means having a clipping level relative to the offset level in the input circuit for clipping input signals relatively greater than the clipping level plus the offset level;

charging means for overbiasing the input circuit whenever the input signal is greater than said offset level and said clipping level, said overbiasing causing crossover distortion.

14. The amplifier of claim **13** wherein the offset means comprises complimentary connected diodes, one each in the input circuit of each solid state device.

15. The amplifier of claim **13** wherein the offset means in the input circuit of each solid state device includes a diode coupled to the input circuit.

16. The amplifier of claim **15** wherein the offset means further comprises a resistor in series with the solid state device.

17. The amplifier of claim **13** wherein the clipping means in the input circuit comprises complimentary connected diodes.

18. The amplifier of claim **13** wherein the charging means comprises a resistor and capacitor network in the input circuit of each solid state device.

19. The amplifier of claim **13** further comprising input amplifier means commonly coupled to the input circuit of each solid state device.

20. The amplifier of claim **13** further comprising output amplifier means commonly coupled for receiving the mixed outputs of the output circuits.

21. The amplifier of claim **13** wherein the offset at the input equals at least one diode voltage drop.

22. The amplifier of claim **13** wherein the gain of each solid state device is unity.

23. The amplifier of claim **13** wherein the offset means establishes an offset in the output circuit.

24. The amplifier of claim **22** wherein the offset in the output circuit equals at least one diode voltage drop.

25. The amplifier of claim **13** further including means at the output of each solid state device for clipping unused opposite half cycles of the input signals.

26. The amplifier of claim **25** wherein the means at the output of each solid state device comprises a diode forward biased with respect to the input for clipping opposite half cycles of the input signal to each solid state device.

* * * * *

UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF CORRECTION

PATENT NO. : 5,524,055 Page 1 of 3

DATED : June 4, 1996

INVENTOR(S) : Jack C. Sondermeyer

It is certified that error appears in the above-indentified patent and that said Letters Patent is hereby corrected as shown below:

The title page, showing an illustrative figure, should be deleted and substitute therefor the attached title page.

Delete Drawing Figure 2,and substitute therefor the Drawing Sheet, consistign of Fig. 2, as shown on the attached pages.

Signed and Sealed this

Eighth Day of October, 1996

Attest:

Bruce Lehman

BRUCE LEHMAN

Attesting Officer

Commissioner of Patents and Trademarks

United States Patent [19]
Sondermeyer

[11] **Patent Number:** **5,524,055**
[45] **Date of Patent:** **Jun. 4, 1996**

[54] **SOLID STATE CIRCUIT FOR EMULATING TUBE COMPRESSION EFFECT**

[75] Inventor: **Jack C. Sondermeyer**, Meridian, Miss.

[73] Assignee: **Peavey Electronics Corporation**, Meridian, Miss.

[21] Appl. No.: **182,493**

[22] Filed: **Jan. 18, 1994**

[51] **Int. Cl.**^{6} **H03G 3/08**
[52] **U.S. Cl.** **381/61**
[58] **Field of Search** 381/61; 330/51, 330/255, 262

[56] **References Cited**

U.S. PATENT DOCUMENTS

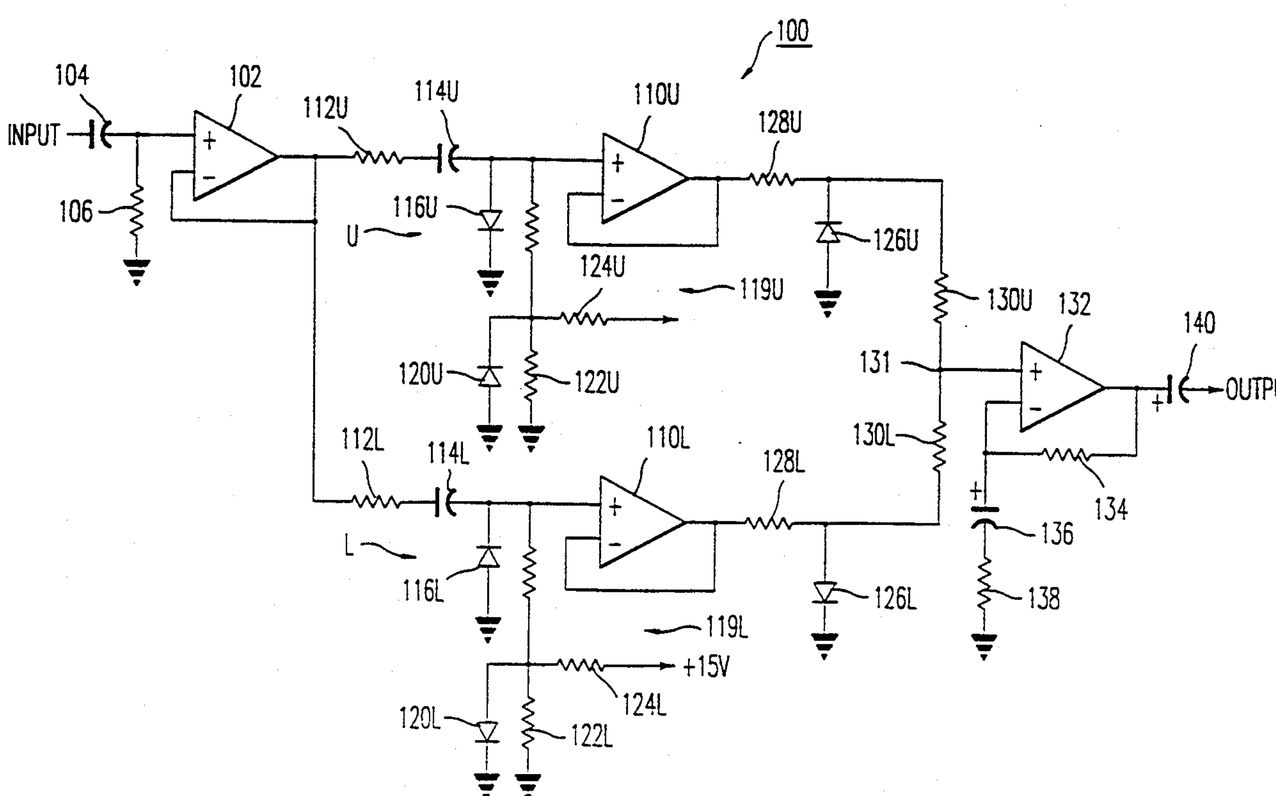
4,405,832	9/1983	Sondermeyer	179/1 D
4,439,742	3/1984	Sondermeyer	330/262
4,809,336	2/1989	Pritchard	381/61
4,937,847	6/1990	Kruger	378/62
4,987,381	1/1991	Butler	330/255
4,995,084	2/1991	Pritchard	379/94
5,012,199	4/1991	McKale	330/51
5,133,014	7/1992	Pritchard	381/61

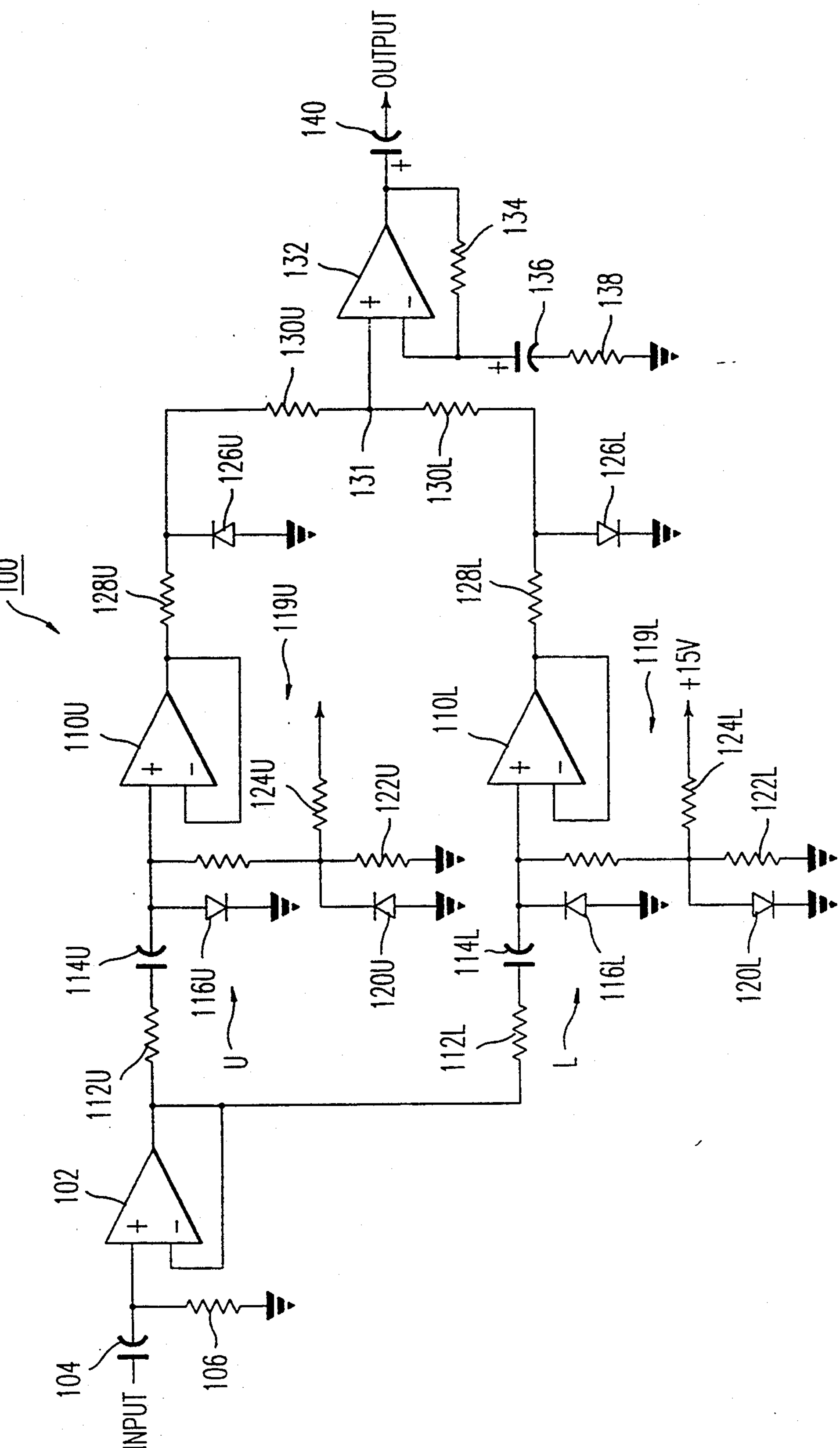
Primary Examiner—Stephen Brinich
Attorney, Agent, or Firm—Watson Cole Stevens Davis

[57] **ABSTRACT**

A solid state amplifier for emulating the compression associated with an overbiased class-B push-pull tube amplifier at high input signal levels due to the flow of current into the grid of the output tubes resulting in a desirable output clipping characteristic with crossover distortion is disclosed. The invention includes at least one pair of class-B connected solid state devices, each having an input circuit and an output circuit. The output circuits are connected for mixing. A biasing element in the input circuit of each paired solid state device establishes a clipping level offset at the input circuit and at the output circuit of each device. A clipping element in the input circuit and the output circuit clips the offset at the input circuit and clips the offset at the output circuit of each respective solid state device. A charging element overbiases the offset in the input circuit whenever the input signal is greater than the input clipping element. The overbiasing causes crossover distortion for emulating the desirable compression associated with a tube amplifier.

26 Claims, 3 Drawing Sheets

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106
102
112U
114U
116U
U
110U
124U
122U
120U
119U
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130L
126L
128L
119L
+15V
110L
124L
122L
120L
116L
114L
112L
L
132
134
136
138
140
OUTPUT

FIG. 2

US005619578A

United States Patent [19]

Sondermeyer et al.

[11] **Patent Number: 5,619,578**

[45] **Date of Patent: Apr. 8, 1997**

[54] **MULTI-STAGE SOLID STATE AMPLIFIER THAT EMULATES TUBE DISTORTION**

[75] Inventors: **Jack C. Sondermeyer**; **James W. Brown, Sr.**, Meridian, both of Miss.

[73] Assignee: **Peavey Electronics Corporation**, Meridian, Miss.

[21] Appl. No.: **299,104**

[22] Filed: **Sep. 2, 1994**

Related U.S. Application Data

[63] Continuation-in-part of Ser. No. 179,546, Jan. 10, 1994.

[51] **Int. Cl.**^{6} **H03G 3/00**

[52] **U.S. Cl.** **381/61**; 381/62

[58] **Field of Search** 381/61, 62–65, 381/96, 120, 118

[56] **References Cited**

U.S. PATENT DOCUMENTS

5,131,044 7/1992 Pritchard 381/61

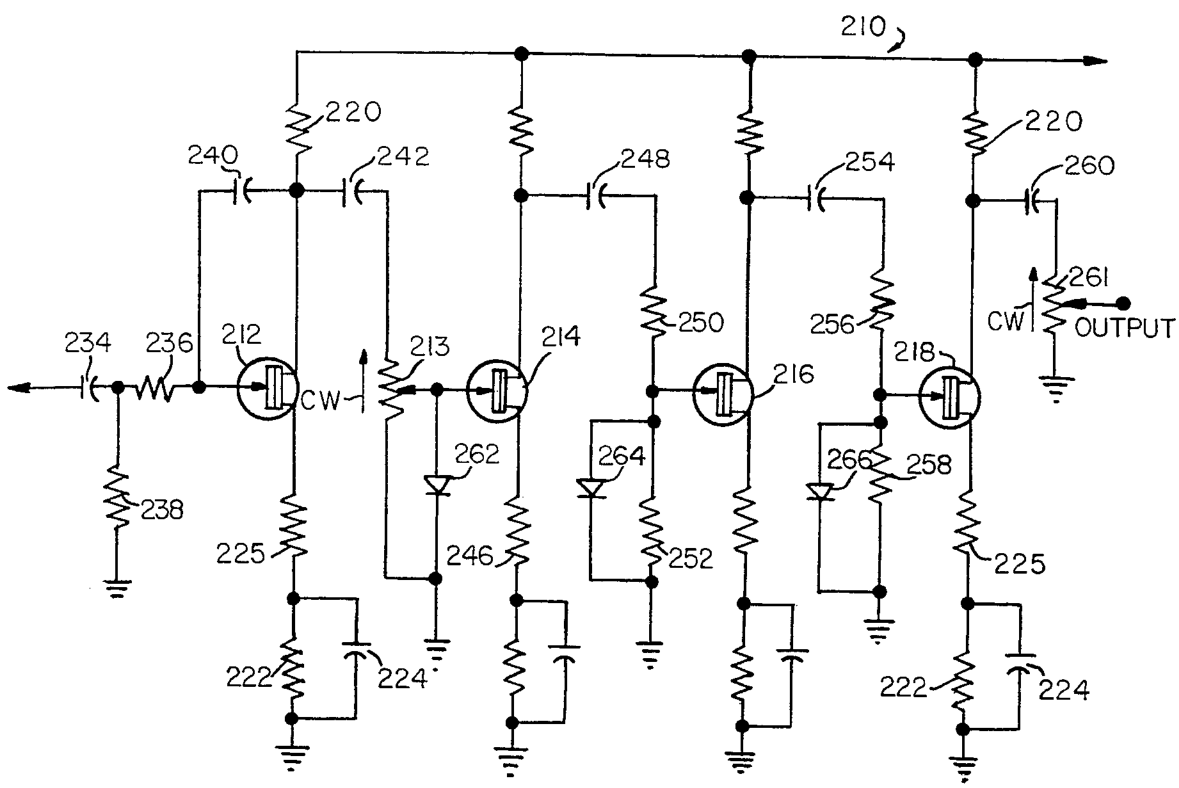
Primary Examiner—Stephen Brinich

Attorney, Agent, or Firm—Watson Cole Stevens Davis, P.L.L.C.

[57] **ABSTRACT**

A multi-stage solid state amplifier emulates the distortion associated with grid current flow in a multi-stage tube amplifier by means of a clipping device in the circuit between each series connected stage. In a particular embodiment, each stage includes a field effect transistor (FET) and the clipping device is a diode. In another embodiment, each stage includes a Darlington connected pair of transistors. An input diode and a multilevel biasing circuit emulates a tube circuit input.

27 Claims, 4 Drawing Sheets



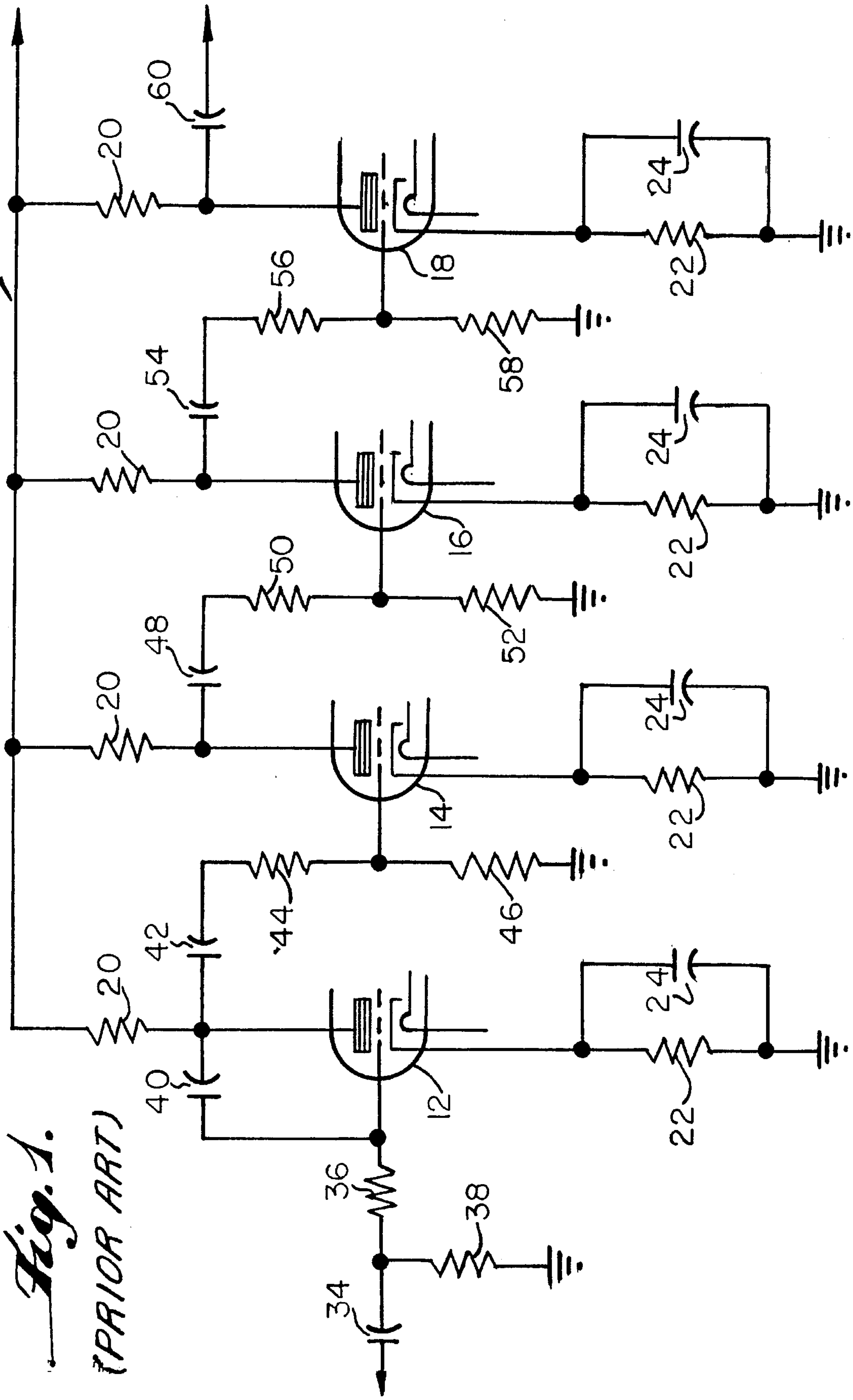


Fig. 1.
(PRIOR ART)

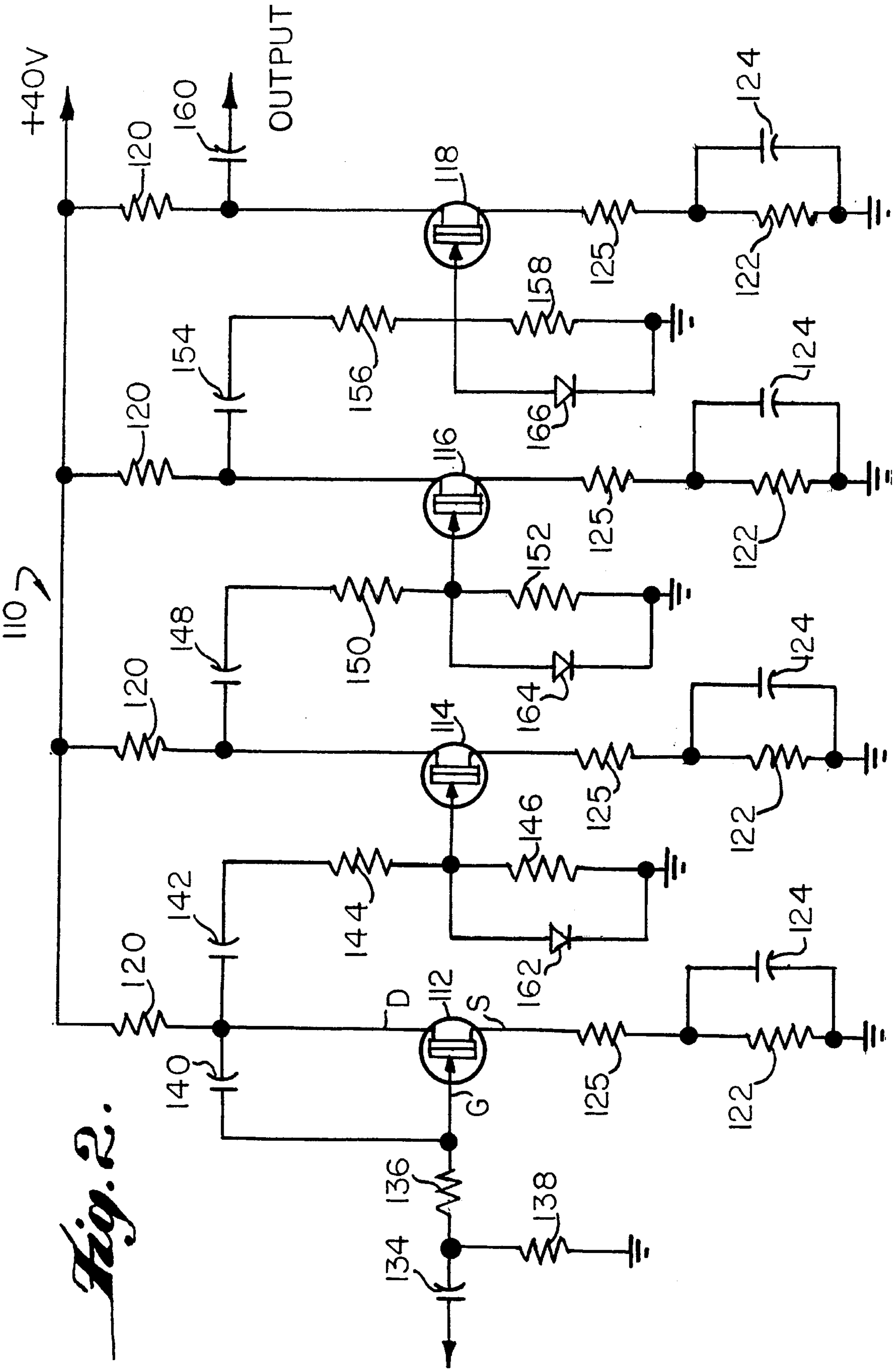
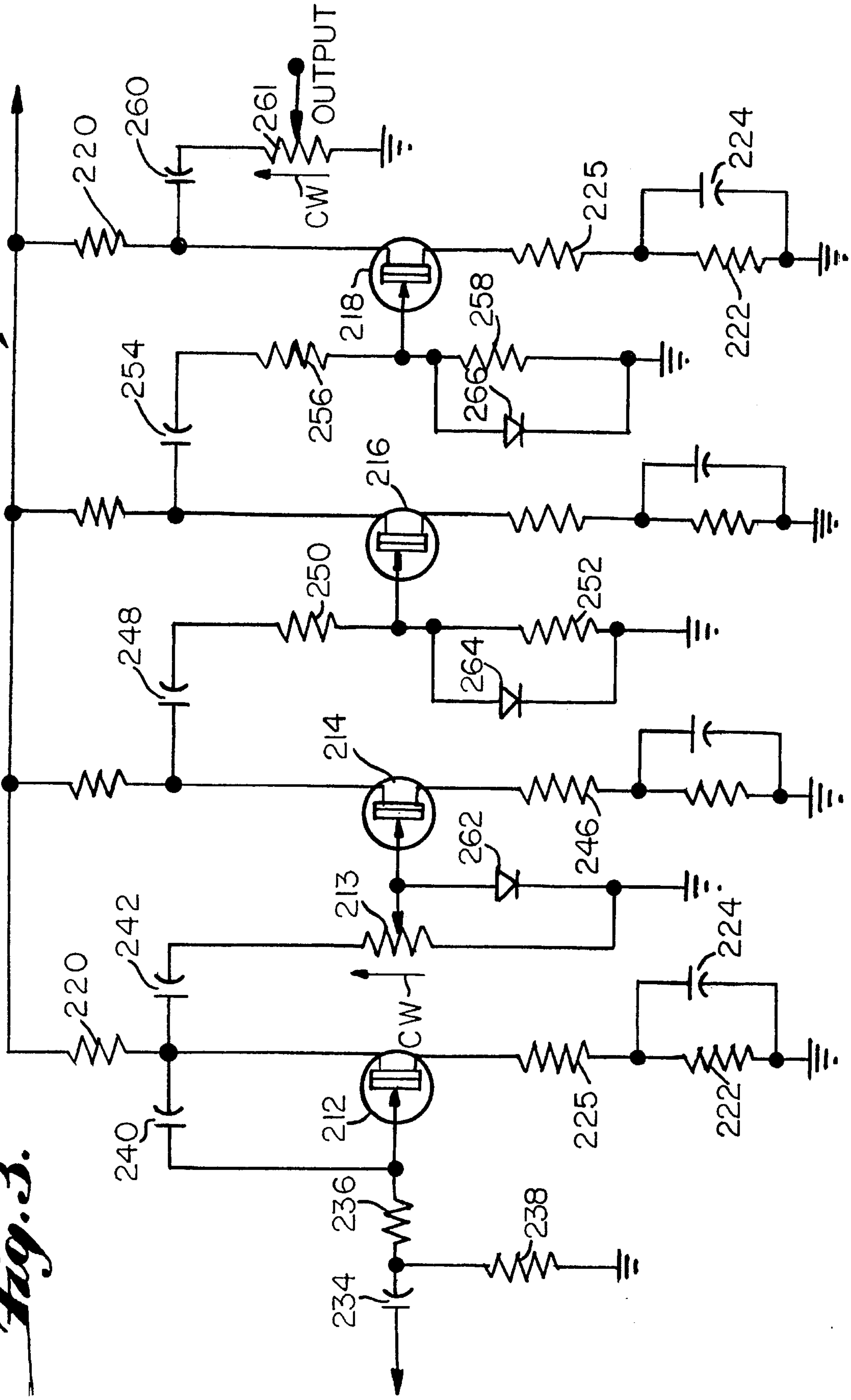

+40V
OUTPUT
160
120
118
124
125
122
154
156
158
166
116
110
148
150
152
164
114
142
144
146
162
112
D
S
G
140
136
138
134

Fig. 2.


OUTPUT
CW
CW
210
212
213
214
216
218
220
222
224
225
234
236
238
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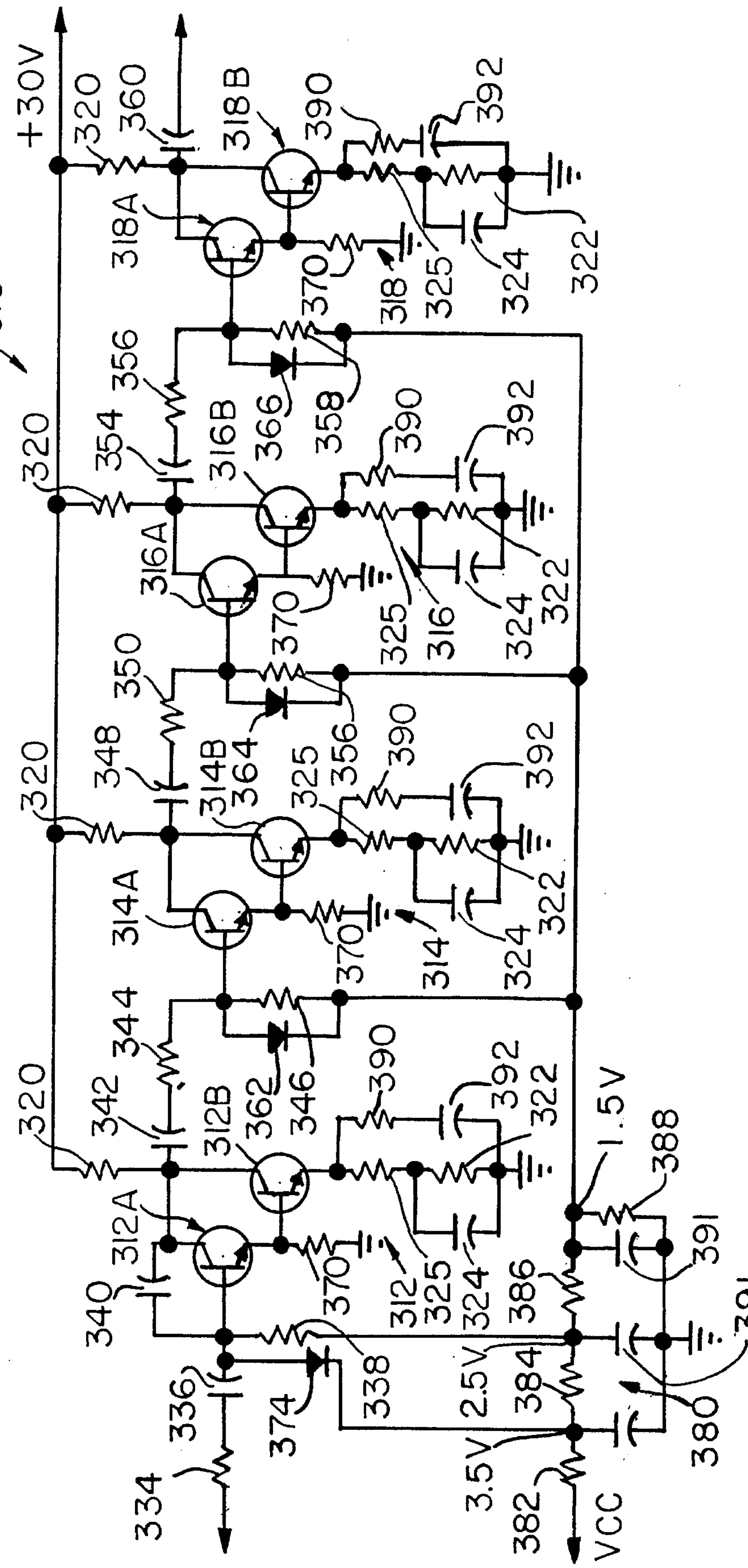
246
248
250
252
254
256
258
260
261
262
264
266

Fig. 3.

Fig. 4.


+30V
310
320
360
318B
390
392
318A
370
318
325
324
322
356
354
316B
366
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316A
316
350
348
314B
364
314A
314
344
342
312B
362
346
312A
340
312
336
334
374
338
386
384
382
380
388
391
1.5V
2.5V
3.5V
VCC

MULTI-STAGE SOLID STATE AMPLIFIER THAT EMULATES TUBE DISTORTION

RELATED APPLICATION

This application is a Continuation-In-Part of U.S. patent application Ser. No. 08/179,546 filed Jan. 10, 1994.

BACKGROUND OF THE INVENTION

The invention pertains to amplifiers for musical instruments. In particular, the invention pertains to a solid state multi-stage amplifier which has distortion so that it sounds like a tube amplifier when overdriven.

Tube amplifiers are often preferred by musical artists because tubes produce a distorted output sound which is familiar and thought to be most pleasing. Solid state amplifiers are often preferred because they tend to be lighter, and are often less expensive to produce, are more durable and consume less power. It is difficult to make a solid state amplifier produce a distorted sound like a tube amplifier. Also, the supply of tubes available for use in amplifiers has become scarce and more expensive.

A known tube amplifier **10** is shown in FIG. **1**. The amplifier is described with exemplary values of the various elements being noted for characterizing the operation of the device. The amplifier **10** illustrated in FIG. **1** is a pre-amplifier comprising four identical tube sections **12**, **14**, **16** and **18** (e.g., four 12AX7 tube sections), each tube section has a corresponding plate resistor **20** (100K ohm) and a cathode resistor **22** (1.5K ohm). Each cathode resistor **22** is bypassed with a capacitor **24** (2.2 uF). With these plate and cathode resistor values, a typical 12AX7 amplifier tube section will idle at approximately 1 mA of plate current, approximately 1.5 volts at the cathode and about +200 volts at the plate from a +300 volt source. A positive grid swing in excess of 1.5 volts peak will cause the grid to conduct. A normal guitar input is coupled to the grid of the first tube stage **12** by a coupling capacitor **34** and a grid resistor **36**. A resistor **38** is coupled to the node between the capacitor **34** and grid resistor **38** and provides a ground reference for the input to tube **12**. Feedback capacitor **40** (10 PF) is coupled between the plate and grid of tube **12** and provides some control of high frequency roll-off, known as the Miller effect, which helps to keep the amplifier stable at open input conditions. The signal from the plate of amplifier **12** is coupled to input of amplifier stage **14** via capacitor **42** and grid resistor **44**. Resistor **46** provides a ground reference for stage **14**. Resistors **44** and **46** act as a voltage divider. The signal from stage **14** is likewise coupled to stage **16** via capacitor **48** and grid resistor **50**, with resistor **52** providing a ground reference for the input and voltage division. Finally, the signal from stage **16** is coupled to the stage **18** via capacitor **54**, grid resistor **56** and reference resistor **58** to ground also with voltage division. The output of stage **18** is coupled to the output of the pre-amplifier **10** by output capacitor **60**.

The coupling capacitor values **42**, **48** and **54**, as well as the values of the divider resistances **44/46**, **50/52** and **56/58** are chosen in a known manner to provide good distorted sound. Typically, with a guitar level input signal applied, the first stage **12** is clean and free of distortion, although with some high level guitars, even this stage clips at times. The first stage output signal level is high enough to cause input clipping at the second stage **14** because the grid of the second stage **14** is driven positive with respect to the cathode and conducts for a substantial portion of the input cycle.

Input clipping at stage **14** results in an average negative voltage on the grid, causing the operating point of the second stage **14** to shift dramatically resulting in a significant amount of second harmonic distortion. The signal at the plate of the second stage **14** resembles a square wave with about two-thirds of the period spent in the positive half cycle. The plate of the second stage **14** has a high enough signal level to cause significant input clipping at the third stage **16**. Here too, the grid swings positive with respect to the cathode. Thus, input clipping causes the operating point of the third stage **16** to shift. This is repeated yet one more time, resulting in input clipping and operating point shift of the fourth stage **18**. The output at the plate of the fourth stage **18** has gone through several different levels of clipping at the input and output and several operating point shifts and is thus rich in harmonics. All of this essentially results in a characteristic sound which is referred to as good tube sound.

In the exemplary pre-amplifier **10** illustrated in FIG. **1**, the available peak plate swing in the positive direction for any stage is about 100 volts (i.e., about one-half the plate voltage). Further, each grid conducts at a positive peak swing of about 1.5 volts. The ratio of 100 to 1.5 or 66.7 is a high number, and its value is important to shift the operating point of each stage enough to generate the appropriate amount of second harmonic distortion. The values of divider resistors **44/46**, **50/52** and **56/58** are also critical, and are carefully chosen to set just the right amount of input clipping and resulting second harmonic distortion to produce a pleasing sound.

Of note here are two key ingredients in the so called distorted tube sound. First, the tube characteristics themselves with the 100 volt output capability and only a 1.5 volt input clipping capability or level is unique and required for successful generation of the second harmonic distortion, and that so called tube sound. Secondly, the multiplicity of stages is necessary for a sustained distortion sound as the guitar output level drops after being plucked by the musician. Although more or fewer stages may be employed, at least three, and preferably four stages are required to achieve the desired distortion sound sought by most musicians.

There is, therefore, a need for a solid state amplifier which is capable of replacing the various tube stages in a multi-stage pre-amplifier, and which may be overdriven to emulate the tube sound produced by the such known tube amplifiers.

SUMMARY OF THE INVENTION

The invention is based upon the discovery that the distortion associated with the flow of current in the grid of a tube amplifier operated at high input levels is duplicated in a multi-stage solid stage amplifier by means of a clipping device in the coupling circuit between stages.

The invention comprises a multistage solid state amplifier for emulating the distortion associated with a tube pre-amplifier when overdriven. The invention includes a plurality of solid state tandem or series connected amplifier stages having an input circuit and an output circuit. Each downstream stage has its input circuit coupled to the output circuit of an upstream stage. A clipping device such as a diode is coupled in the input circuit between the stages.

In a particular embodiment, each solid state amplifier stage includes a field effect transistor (FET) having its output terminal coupled to the input of the next downstream stage and the diode is located in the input circuit so as to duplicate the desirable input clipping characteristics of a tube amplifier, wherein the ratio of the output capability to the input clipping level between stages is sufficient to result in adequate second harmonic distortion.

In another embodiment, each solid state amplifier stage comprises a transistor. An alternative arrangement employs a Darlington transistor. When separate transistors are Darlington connected, an internal base resistor to ground may be employed to improve turn-off characteristics of each stage.

In yet another embodiment, an input diode is employed with a multilevel bias circuit to duplicate the input characteristics of a tube amplifier.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. **1** is a schematic illustration of a known vacuum tube amplifier which exhibits a desirable distorted output;

FIG. **2** is a schematic illustration of a multi-stage solid state amplifier according to the invention, employing FET devices, which emulates the distortion and sound produced by known tube amplifiers;

FIG. **3** is a schematic illustration of a particular embodiment of a multi-stage solid stage application according to the invention; and

FIG. **4** is a schematic illustration of another embodiment of the invention similar to the arrangement of FIG. **2** wherein each FET is replaced by a Darlington connected pair of transistors and a bias circuit with multiple biasing levels along with an input diode.

DESCRIPTION OF THE INVENTION

The present invention as illustrated in FIG. **2**, is directed to a solid state multi-stage amplifier **110** which uses components arranged in a manner similar-to the known device of FIG. **1**. In FIG. **2**, the components are numbered with reference numerals which correspond to the reference numerals of FIG. **1** in a **100** series, and the tube sections of the known device have been replaced with four solid state amplifier stages **112**, **114**, **116** and **118**. The devices illustrated are field effect transistors (FET) sometimes referred to as J-FET devices, which are supplied by a 40 volt supply. Each amplifier **112–118** has a source S, a drain D and a gate G terminal as illustrated. The drain D corresponds to the output of the device, and the gate G corresponds to the input of the device. Each stage **112–118** includes a drain resistor **120** similar in value to the plate resistor in FIG. **1** (100K ohm). In addition, each stage employs a bias circuit including a resistor **122** (33K ohm) and a bypass capacitor **124** (2.2 uF) in parallel. The source S is coupled through the source resistor **125** and the bias circuit to ground in a self biasing configuration, as shown. In addition, each source S has a source resistor **125** (1K ohm) to set the gain nominally at 100, which is similar to most tube stages. The drain and source resistor values are adjusted so that each FET stage **112**, **114**, **116** and **118** with a pinch-off voltage of approximately 6 volts will idle at approximately 180 uA of source current. Each will have approximately +22 volts at the drain D and approximately +6 volts at the source S.

In FIG. **2**, the drain D, or output, of stage **112** is coupled to the gate G, or input, of stage **114** by coupling capacitor **142** and gate resistor **144**. Resistor **146** provides a ground reference to the gate G of stage **114**. Also, resistors **144** and **146** act as a voltage divider. Similarly, as in FIG. **1**, the successive stages **116** and **117** are coupled by a corresponding combination of a coupling capacitor, gate resistor and reference resistor **148**, **150** and **152** and **154**, **156** and **158** respectively. The final stage **118** is coupled to the output by means of coupling capacitor **160**. The input stage **112** has a Miller capacitor **140** between drain D and the gate G, as illustrated.

In FIG. **2**, without the clipping means, the available peak drain signal output capability or swing in the positive direction for any stage is about 18 volts, i.e., the difference between the drain D and the source S. In each amplifier stage **112–118**, the gate G operates as a diode which conducts at a positive peak of about 7 volts. Thus, the ratio of the drain output capability (18 volts) and the gate swing (7 volts) is about 2.57. This ratio is insufficient to cause adequate second harmonic distortion.

In the present invention, clipping means is provided between the stages. In the embodiment disclosed, clipping may be achieved by means of diodes **162**, **164** and **166** provided in parallel with the corresponding reference resistors **146**, **152** and **158** in each of the respective stages **114**, **116** and **118**. Each diode **162**, **164** and **166** has its cathode coupled to ground and its anode coupled to a node between the divider resistors **144/146**, **150/152** and **156/158** of each stage. Each diode **162**, **164** and **166** conducts in the forward direction and thereby establishes a clipping level at approximately +0.5 volts change in gate swing to emulate distortion associated with grid conduction in a tube amplifier. The ratio of the drain signal output capability or swing (+18 V) to the gate clipping level (+0.5 V) is 18/0.5=36, which is not as high as a tube circuit. However, while a higher ratio is desirable, in a solid state circuit, a ratio of about 30 is sufficient to cause adequate second harmonic distortion in each stage. Accordingly, the solid state, multi-stage preamplifier **110** of the invention produces distortion performance which is quite similar to that of the tube circuit in FIG. **1** by employing a low level input clipping means, such as diodes **162**, **164** and **166**.

It should be understood that other solid state devices may be employed other than the J-FET type devices and the diode illustrated. Also, gain, coupling and high-frequency characteristics may be tailored by changing the various element values. However, the clipping means provided between stages is effective to produce the operating second harmonic distortion which causes tubelike sound from a solid state amplifier.

FIG. **3** is an illustration of another embodiment of a multi-stage, solid state **210** amplifier according to the invention. Similar elements have similar reference numbers as shown in FIG. **2** in a **200** series. In FIG. **3**, however, a potentiometer **213** is substituted for the fixed divider resistors **144** and **146** between stages **112** and **114**. The wiper is coupled to the input of stage **214** and the potentiometer **213** allows maximum distortion in the CW direction where the maximum signal is applied to stage **214**.

At the output, a potentiometer **261** is coupled to the coupling capacitor **260**. The wiper acts as the output terminal. The output level is maximum when the wiper is in the full CW position. The ability to independently vary the distortion and to vary the level adds versatility to the circuit and allows the artist to tailor the distorted sound and volume at will.

FIG. **4** is an illustration of another embodiment of a multi-stage, solid state **310** amplifier having stages **312**, **314**, **316** and **318** according to the invention. Similar elements have similar reference numbers as shown in FIG. **2** in a **300** series. In FIG. **4**, however, a pair of Darlington connected transistors **312A–312B**, **314A–314B**, **316A–316B** and **318A–318B** is substituted for each corresponding FET **112**, **114**, **116** and **118**. A multilevel biasing system with an input diode 374 is also provided. An internal base resistor 370 for each stage may be employed for favorably affecting the turn-off characteristics.

In FIG. 4, the circuit function is similar to that of FIG. 2. However, the use of the Darlington connected transistors allows cost and performance advantages. The FET devices described above require a tight pinch-off voltage limit which increases the price considerably. For the transistor devices, the input essentially looks like two diodes which is predictable in terms of performance.

The amplifier 310 of FIG. 4 uses a pair of NPN type Darlington connected transistors 312A–312B, 314A–314B, 316A–316B and 318A–318B for each stage. The NPN type is chosen because the supply is a positive 40 V. A negative supply would allow an opposite device type. Alternatively, a single transistor would operate in the embodiment discussed herein. However, a single transistor would not offer a gain or input impedance to match that of the tube or the FET. Thus, paired transistors are preferred. An integrated Darlington NPN transistor may also be used. However, the discrete transistors offer slightly better performance at high frequencies, because access to the internal base connection is available. In addition, discrete transistors cost less.

The design of FIG. 4 requires a circuit arrangement to bias the transistors. Accordingly, a separate bias supply VCC is provided. The bias voltage for the first stage 312 is adjusted to a first level and a common bias voltage is used for the remaining three stages 314–318. In the exemplary embodiment, all stages 312–318 employ respective 470K bias supply resistors 338, 346, 356 and 358 and 150K collector resistors 320. Each inter-stage has a corresponding input diode 362, 364, 366 to cause the operating point shift therein. Further, each stage has a 470K base resistor 370 from the available internal base to ground in the base emitter circuit between the transistors to improve the turn-off characteristics. The input stage 312 has an input diode 374 for better tube input circuit emulation as discussed hereinafter.

Many of the newer or more contemporary guitars provide a relatively high output voltage, e.g., 3–5 V. It is thus important to have good preamp input overload. Accordingly, the first stage 312 has been biased to duplicate the input overload of a typical 12AX7 tube grid by providing a multilevel bias arrangement.

In the arrangement, bias circuit 380 includes a divider network including resistors 382, 384, 386 and 388 and filter capacitors 391. Resistor 382 is coupled to the anode of input diode 374 and establishes a diode bias level (e.g., 3.5 V DC) for stage 312. Resistor 384 is coupled to the base resistor 338 for establishing the base bias of stage 312 (e.g., 2.5 V DC). Resistor 386 and diode resistor 388 establish a common bias voltage for the downstream stages 314–318, as shown (e.g., 1.5 V DC).

As noted in the exemplary embodiment, VCC is divided to establish a 2.5 volt base bias supply, a 3.5 volt diode supply and a common 1.5 volt supply. In each of the stages 314–318, the corresponding clipping diode 362, 364, 366 is also connected to the common 1.5 volt supply.

In the exemplary embodiment, with the first stage input transistor 312A base is biased at 2.5 volts, the emitter of output transistor 312B finds itself at about 1.5 volts (i.e., two diode drops 0.5 V), which is the same value as the cathode in the previously described tube circuit (FIG. 1). This means that a negative peak swing of 1.5 volts at the input of this stage 312 will cause the operating collector current to go to zero (negative clipping). Such is also the case for that of the tube circuit of FIG. 1. Thus, the circuit of FIG. 4 looks like a tube circuit input.

The first stage input diode 374, conducts at 0.5 V. The 3.5 volt supply, is 1 volt greater than the base supply voltage. The sum of the diode bias voltage of 3.5 V plus the diode drop of 0.5 V equals 4 V which is 1.5 V above the base bias voltage of 2.5 V. This means that a positive peak of 1.5 volts at the input of this stage will cause the input diode 374 to conduct and force an operating point shift (positive clipping). This is also the case for that of the tube circuit. Thus, the circuit of FIG. 4 matches the input dynamic range of the tube circuit.

As noted in FIG. 2, capacitor 340 provides some controlled high frequency roll-off (Miller effect), and the emitter of 312B has series resistors 322 and 325 to ground, with resistor 322 bypassed with capacitor 324. This circuit arrangement provides idle current and gain values to cause overload conditions that match a typical tube first stage. A series circuit, including a resistor 390 and capacitor 392 is coupled across the emitter circuit to achieve a high frequency boost.

The first stage output signal is coupled to the input of the second stage 314 through capacitor 342 and resistor 344 in a way similar to the arrangement of FIG. 2. The remaining three stages 314, 316 and 318 are similar, although component values may change to achieve the desired amount of clipping, operating point shift and frequency response to produce a pleasing overload sound. For example, inter-stage coupling is provided by capacitors 342, 348 and 354 and resistors 344, 350 and 356. In each downstream stage 314–318, the emitters have dual series resistors 322–325 to ground and the ground resistor 322 is bypassed with capacitor 324. Also, each stage employs series resistors 390 and capacitors 392 to ground to provide high frequency boost. The output is delivered via capacitor 360.

The available peak collector swing in the positive direction for any stage is about 20 volts. Each of the three inter-stage clipping diodes 362, 364 and 366 conducts in the forward direction at approximately 0.5 volts. Thus, the ratio of input swing to clipping voltage 20/0.5=40 is adequate to cause the operating point shift of each stage, and the distortion performance is very similar to that of a tube circuit.

It should be understood that the input diode 374 may be employed with an appropriate bias circuit in the arrangement of FIGS. 2 or 3. This circuit allows the circuit to have the same input characteristics as a tube circuit.

In the arrangement of FIG. 4, a volume potentiometer similar to the potentiometer 213 may be inserted in the circuit after capacitor 342, and a master volume potentiometer similar to the potentiometer 261 may be connected after capacitor 360.

It should be noted that this invention is not simply a diode clipping means, and an operating point shifter. Several U.S. patents discuss operating point shift by means of a diode, with resulting second harmonic distortion. The invention shows the important discovery that a diode with a mere 0.5 V forward voltage clipping value, when driven from a typical solid state device with an output capability of about 20 to 30 volts will closely match the particular output/input ratio of the existing tube circuits and therefore closely emulate the tube sound. The present invention also teaches that multiple operating point shifts produced by multiple stages generate a multiplicity of levels and amounts of second harmonic distortion over a wide range input signal levels. The invention further shows a way to duplicate the input overload characteristic of the typical first stage of a tube circuit with a solid state device. The result then, is the so-called tube distortion sound; a sound that can be generated almost exactly with solid state devices.

While there have been described what are at present considered to be the preferred embodiments of the present invention, it will be apparent to those skilled in the art that various changes and modifications may be made therein without departing from the invention, and it is intended in the appended claims to cover such changes and modifications as fall within the spirit and scope of the invention.

What is claimed is:

1. A solid state amplifier for emulating the distortion associated with a flow of current in the grid of an overdriven multi-stage tube amplifier at high input signal levels resulting in a desirable input clipping characteristic comprising:

a plurality of series connected solid state devices for amplifying a signal each including an input circuit and an output circuit having an output signal capability, each device downstream of a first one of said devices having its input circuit coupled to the output circuit of one of such devices immediately upstream thereof; and

clipping means in the input circuit between each of the devices for establishing a clipping level in one direction between such devices, duplicating in such solid state amplifier the desirable input clipping characteristic of a tube amplifier wherein the output signal capability and the clipping level between stages are in a ratio similar to tube circuits sufficient to result in adequate second harmonic distortion to emulate the distortion effect of a tube amplifier.

2. The amplifier of claim 1 wherein each device comprises a field effect transistor.

3. The amplifier of claim 1 wherein each device comprises a transistor.

4. The amplifier of claim 1 wherein each device comprises a Darlington transistor.

5. The amplifier of claim 4 wherein the Darlington transistor is integrated.

6. The amplifier of claim 4 wherein the Darlington transistor comprises discrete elements.

7. The amplifier of claim 6 wherein the Darlington transistor includes a base-emitter circuit and a bias resistor coupled between the base-emitter circuit and ground.

8. The amplifier of claim 1 wherein the clipping means comprises a diode.

9. The amplifier of claim 8 wherein the output signal capability is effective to forward bias the diode in the input circuit of the device immediately downstream thereof.

10. The amplifier of claim 8 wherein the diode produces forward voltage clipping at about 0.5 V.

11. The amplifier of claim 1 wherein the ratio is at least about 30.

12. The amplifier of claim 1 further including variable input means for at least one of said input circuits for varying the distortion.

13. The amplifier of claim 12 wherein variable input means comprises a potentiometer coupled in output circuit of a first one of said devices, said potentiometer having a wiper coupled to the input of a second one of said devices and the clipping means.

14. The amplifier of claim 1 further comprising variable output means coupled in the output circuit of the last stage for varying an overall output level of the amplifier.

15. The amplifier of claim 14 wherein the variable output means comprises a potentiometer.

16. The amplifier of claim 1 further comprising biasing means for the solid state devices.

17. The amplifier of claim 16 further comprising multi-level biasing means including an upstream biasing means for a first upstream one of the devices and a downstream biasing means for downstream ones of said devices.

18. The amplifier of claim 17 wherein the upstream biasing means comprises a base circuit biasing means coupled to a base element of the upstream solid state device.

19. The amplifier of claim 1 further comprising input diode means coupled between an input of the first upstream stage and a reference.

20. The amplifier of claim 19 further comprising upstream biasing means for the diode.

21. The amplifier of claim 1 further comprising a diode circuit biasing means for establishing a diode circuit bias level, a base circuit biasing means for establishing a base circuit bias level and a downstream biasing means for establishing a downstream bias level and wherein the diode biasing level is greater than the base circuit bias level which in turn is greater than the downstream bias level.

22. The amplifier of claim 1 further comprising means for boosting frequency response at high frequency levels coupled to each device.

23. A solid state amplifier comprising:

a plurality of series connected single ended solid state devices for amplifying a signal each device including an input circuit and an output circuit having an output signal capability, each device located downstream of a first one of said devices having its input circuit coupled to the output circuit of one of said devices located immediately upstream thereof;

a clipper connected in the input circuit of each device downstream of the first one of said devices for clipping signals in one direction above a selected level, and wherein the output signal capability and the input signal level between devices are in a ratio similar to tube circuits sufficient to result in second harmonic distortion.

24. A solid state amplifier comprising:

a plurality of series connected solid state devices each including an input circuit and an output circuit having an output capability, each device downstream of a first one of said devices having its input circuit coupled to the output circuit of one of said devices connected immediately upstream thereof;

a clipper connected to the input circuit of each device downstream of the first one of said devices for establishing a clipping level therefor in one direction; and

interstage coupling means for establishing a selected relationship between the output capability of each device and the clipping level similar to tube circuits to emulate the distortion effect of a tube amplifier.

25. A solid state amplifier comprising:

a plurality of series connected Darlington transistors, each Darlington transistor including an input circuit and an output circuit having an output signal capability, each Darlington transistor located downstream of a first one of said Darlington transistor having its input circuit coupled to the output circuit of a Darlington transistor connected immediately upstream thereof; and

a clipper in the input circuit of each Darlington transistor downstream of the first one of said Darlington transistor.

26. The amplifier according to the claim 25 wherein the clipping in the input circuit of each Darlington transistor establishes a clipping level between the Darlington transistors, duplicating in such solid state amplifier the input clipping characteristic of a tube amplifier, and the output signal capability and the input clipping level between the

Darlington transistors being in a ratio sufficient to result in second harmonic distortion.

27. The amplifier according to claim **25** further including interstage coupling means for establishing a selected relationship between the output capability of each device and the clipping level.

* * * * *

US005841875A

United States Patent [19]

Kuroki et al.

[11] **Patent Number: 5,841,875**

[45] **Date of Patent: Nov. 24, 1998**

[54] **DIGITAL AUDIO SIGNAL PROCESSOR WITH HARMONICS MODIFICATION**

[75] Inventors: **Ryuichiro Kuroki**; **Tsugio Ito**, both of Hamamatsu, Japan

[73] Assignee: **Yamaha Corporation**, Hamamatsu, Japan

[21] Appl. No.: **588,288**

[22] Filed: **Jan. 18, 1996**

Related U.S. Application Data

[63] Continuation of Ser. No. 968,539, Oct. 29, 1992, abandoned.

[30] **Foreign Application Priority Data**

Oct. 30, 1991 [JP] Japan 3-311609

[51] **Int. Cl.**[6] **H03G 3/00**

[52] **U.S. Cl.** **381/61**; 381/98

[58] **Field of Search** 381/61, 98, 63

[56] **References Cited**

U.S. PATENT DOCUMENTS

4,987,381	1/1991	Butler	381/61
4,991,218	2/1991	Kramer .	
4,995,084	2/1991	Prichard	381/61

OTHER PUBLICATIONS

Advertisement on Oct., 1980 *Guitar Player,* p. 19, for Ibanez UE–400 and Service Manual for UE–400.

Primary Examiner—Forester W. Isen
Attorney, Agent, or Firm—Loeb & Loeb LLP

[57] **ABSTRACT**

The audio signal processor has a harmonics modifier for processing an input audio signal to produce an output audio signal. The harmonics modifier is comprised of translating means for sequentially translating the input audio signal into an address signal according to a sampled amplitude of the input audio signal. Memory means is provided for storing an input/output conversion table containing amplitude values in addressable manner. Conversion means operates for accessing the memory means in response to the address signal to effect input/output conversion to read out a sequence of the amplitude values which form the output audio signal containing desired harmonic components according to results of the input/output conversion.

24 Claims, 19 Drawing Sheets

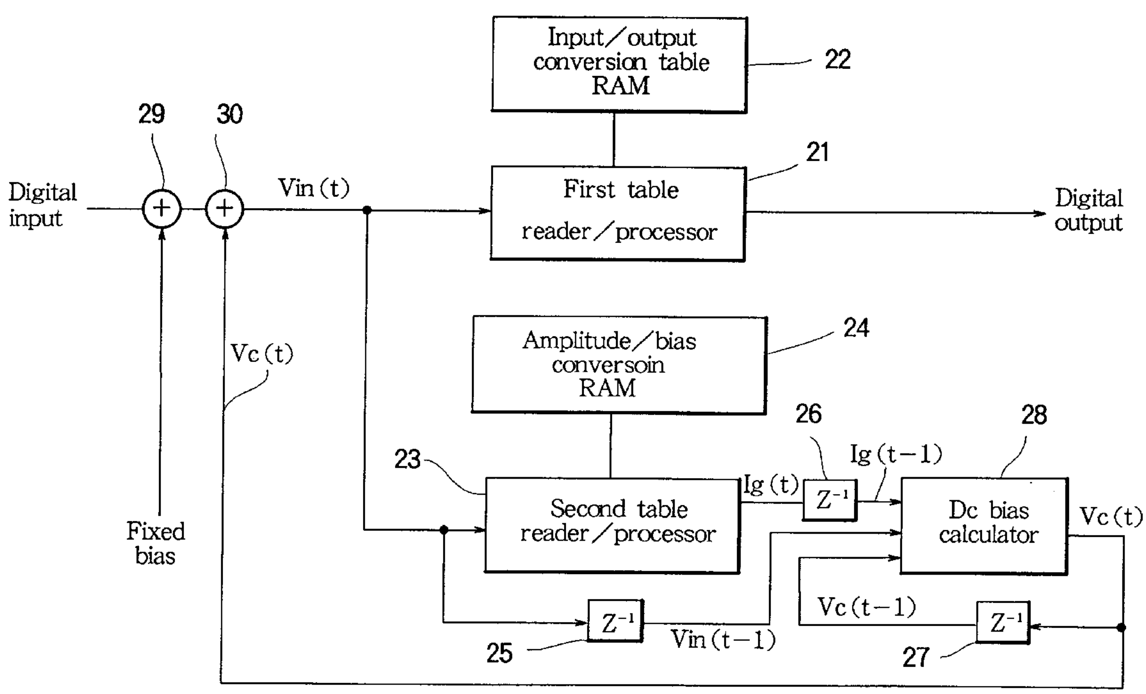
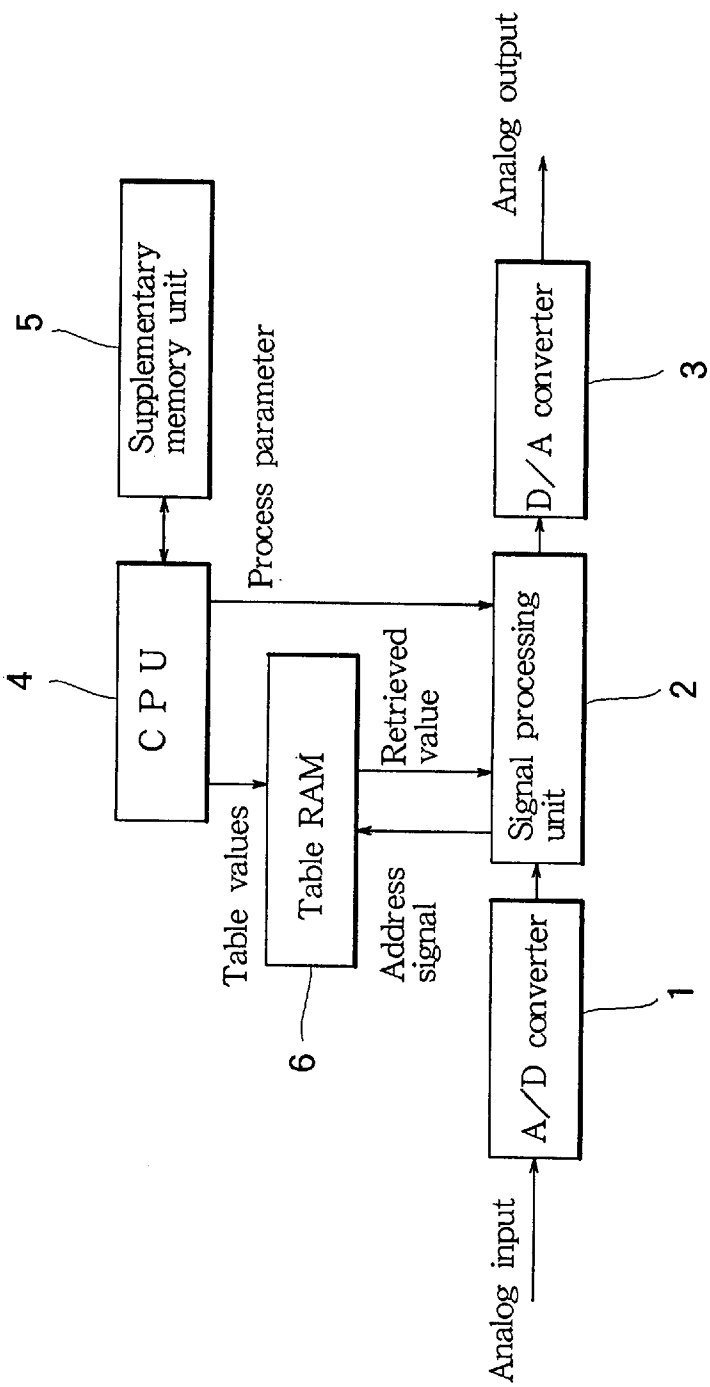


FIG. 1



Supplementary memory unit
5
C P U
4
Process parameter
Table values
Table RAM
6
Retrieved value
Address signal
Signal processing unit
2
D/A converter
3
Analog output
A/D converter
1
Analog input

FIG.2

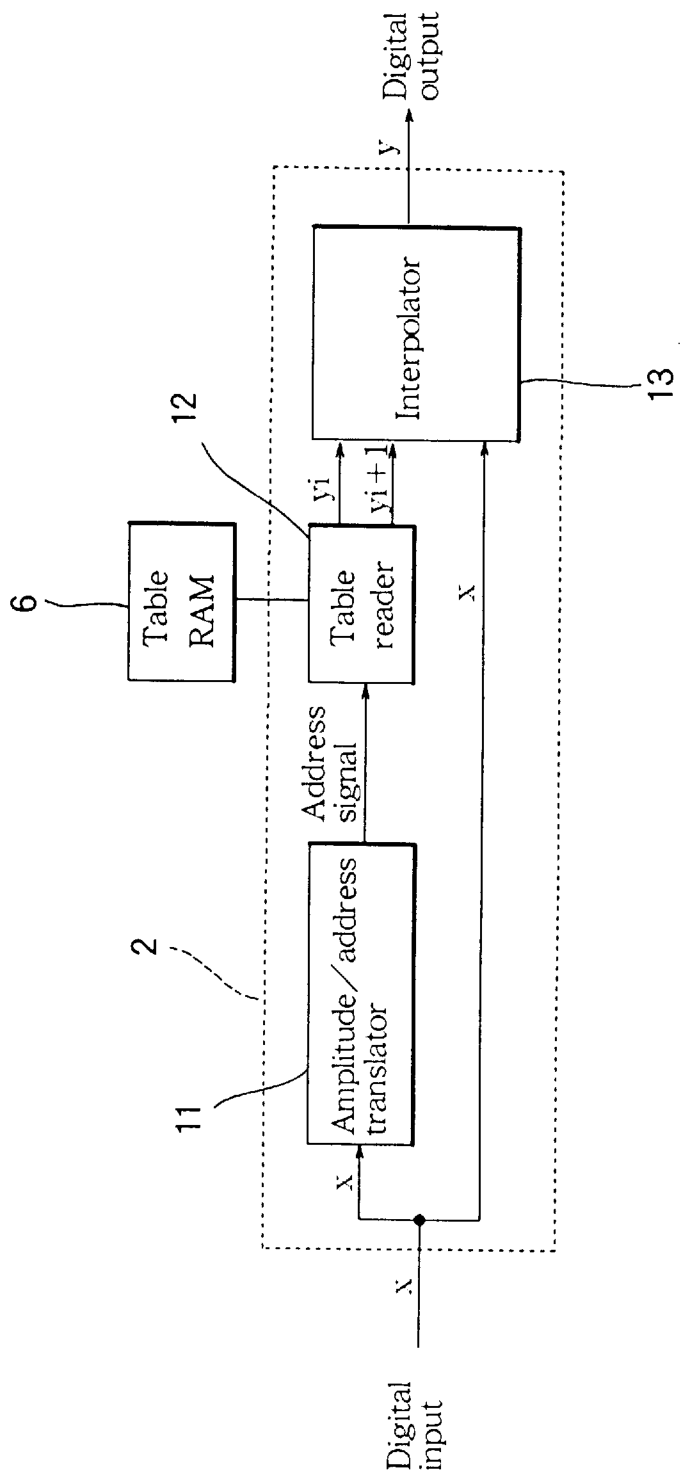

Digital input
x
x
Amplitude/address translator
11
2
Address signal
Table reader
12
Table RAM
6
yi
yi+1
x
Interpolator
13
y
Digital output

FIG. 3(a)

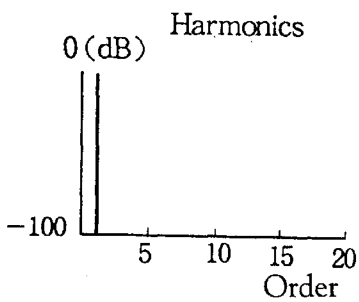


FIG. 3(b)

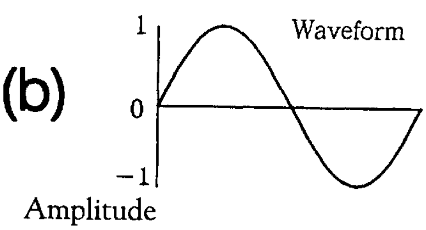


FIG. 3(c)

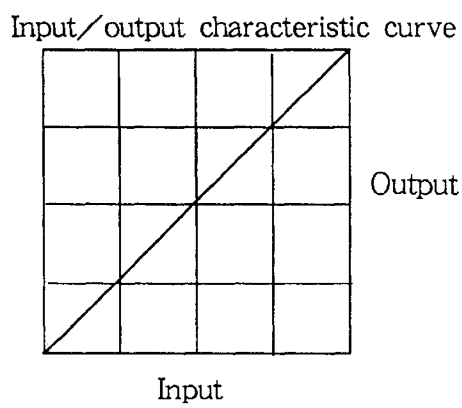


FIG. 4(a)

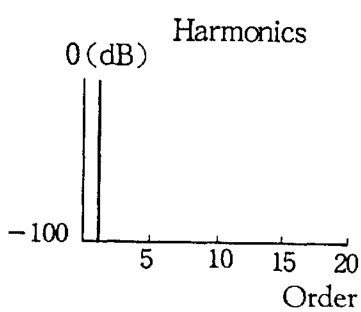


FIG. 4(b)

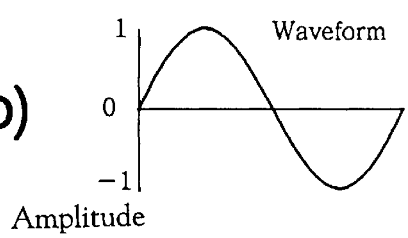


FIG. 4(c)

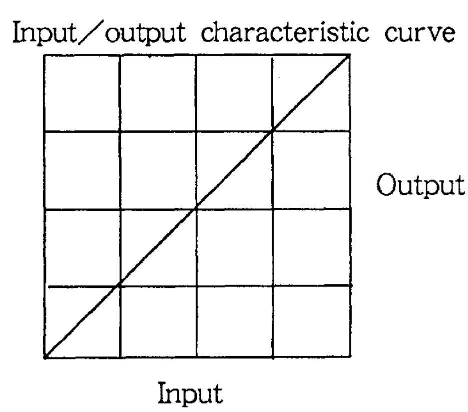


FIG. 5(a)

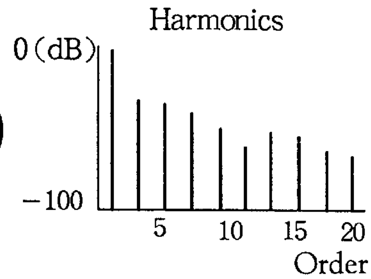


FIG. 5(b)

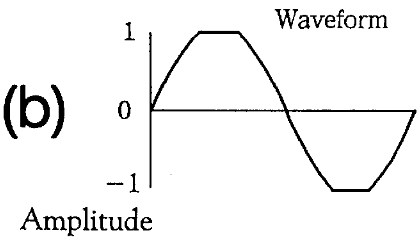


FIG. 5(c)

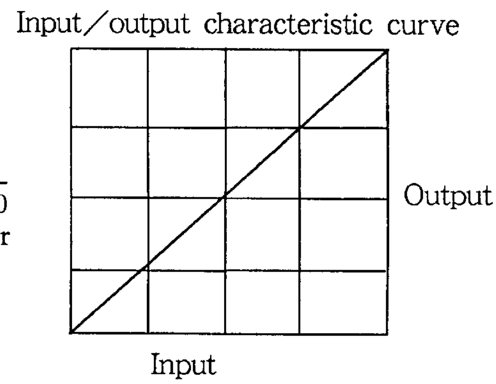


FIG. 6(a)

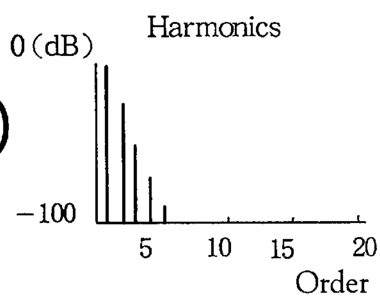


FIG. 6(b)

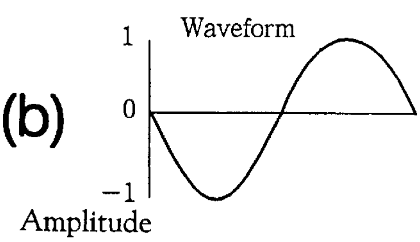


FIG. 6(c)

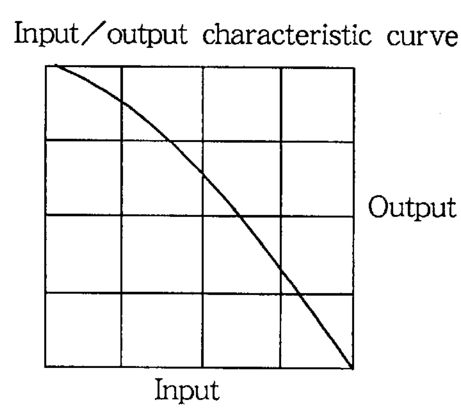


FIG. 7(a)

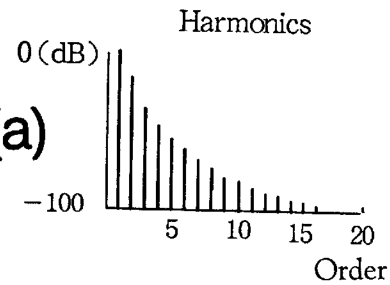


FIG. 7(b)

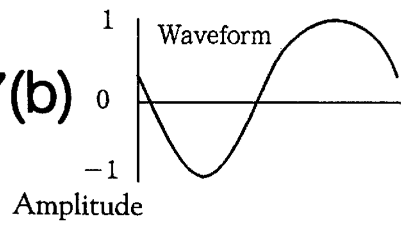


FIG. 7(c)

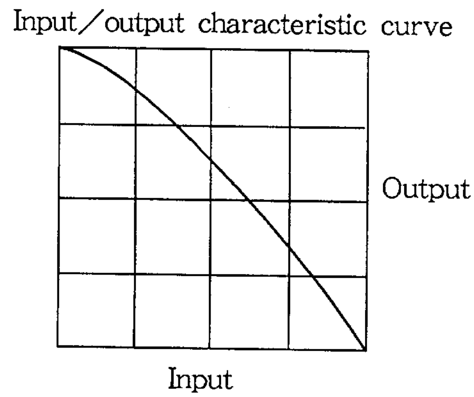


FIG. 8(a)

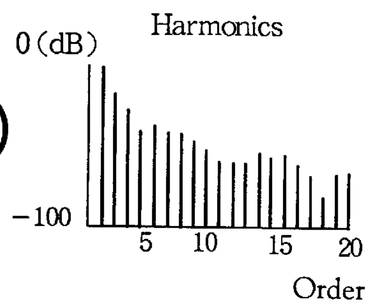


FIG. 8(b)

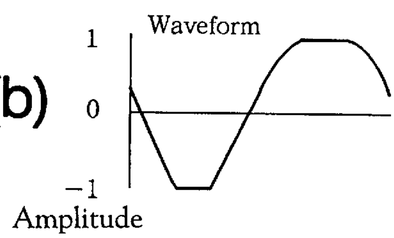


FIG. 8(c)

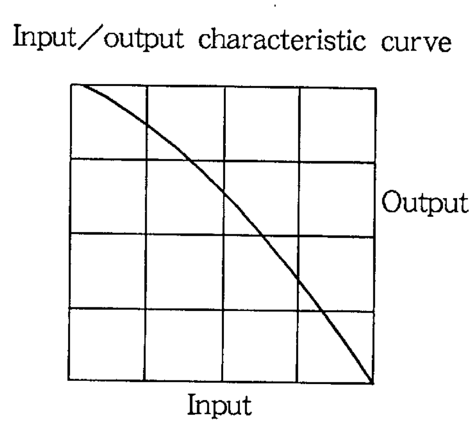
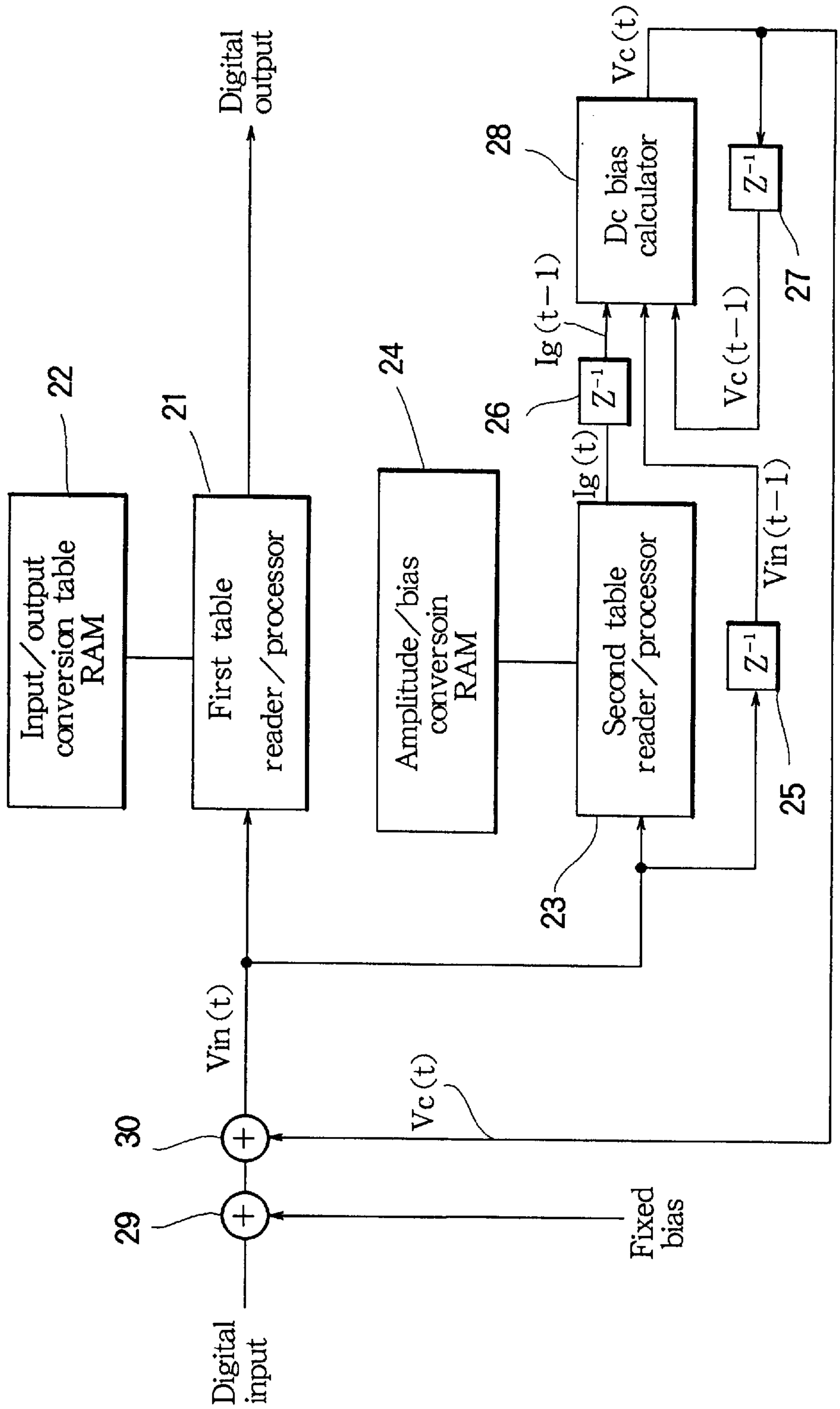


FIG.9



Digital input
29
Fixed bias
30
Vc(t)
Vin(t)
Input/output conversion table RAM
22
First table reader/processor
21
Digital output
Amplitude/bias conversoin RAM
24
Second table reader/processor
23
Ig(t)
26
Z-1
Ig(t−1)
Dc bias calculator
28
Vc(t)
Z-1
27
Vc(t−1)
Vin(t−1)
Z-1
25

FIG. 10

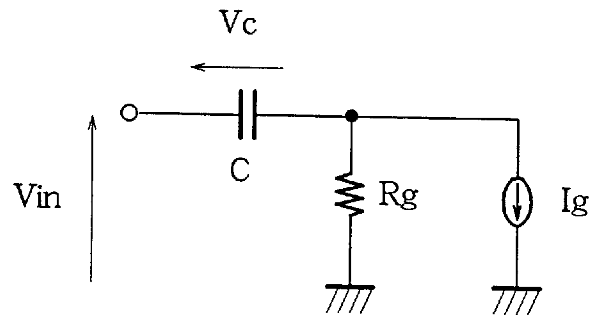


FIG. 11

Input/Output Characteristic Curve

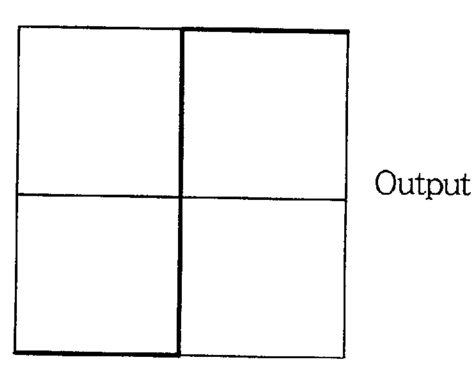


FIG.12

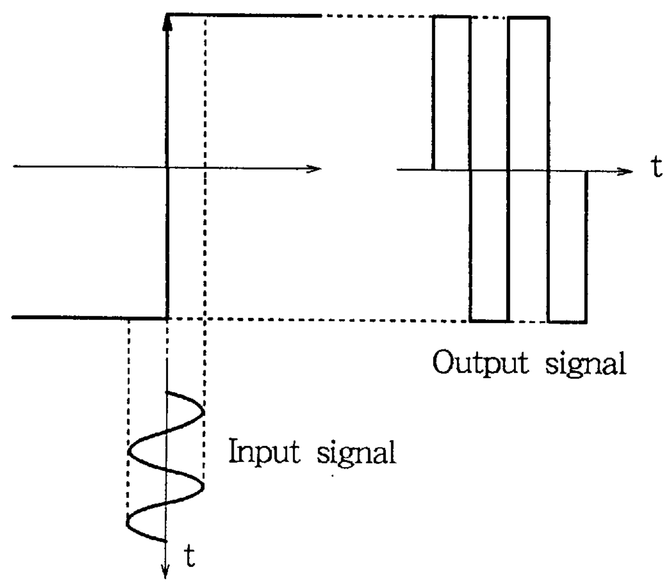

t
Output signal
Input signal
t

FIG.13

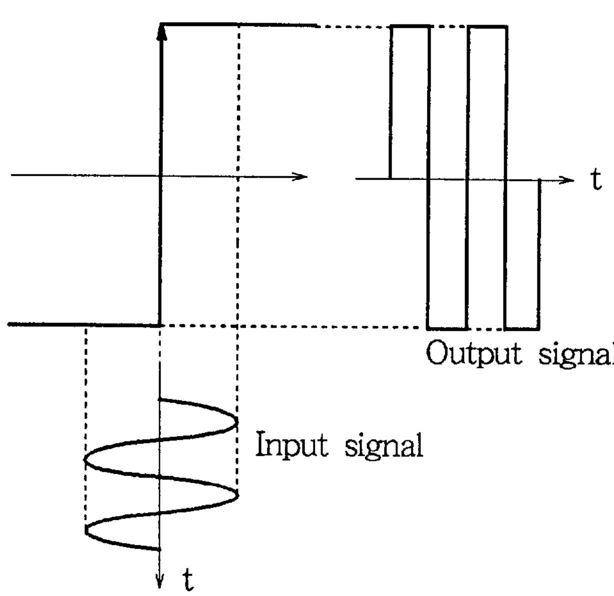

t
Output signal
Input signal
t

FIG. 14

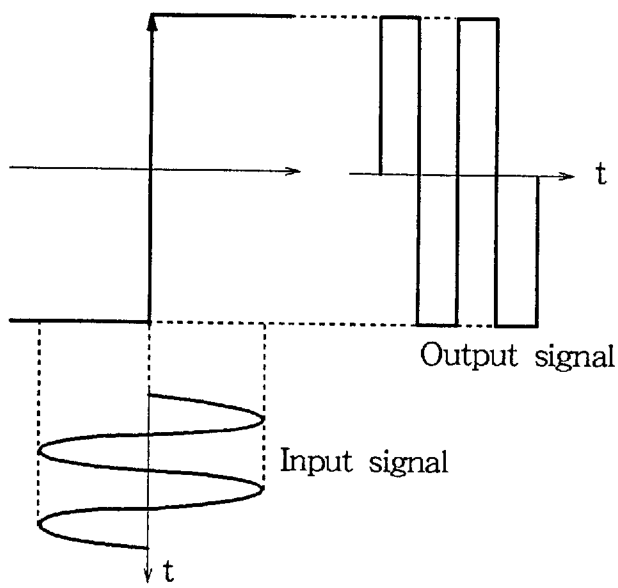

t
Output signal
Input signal
t

FIG. 15

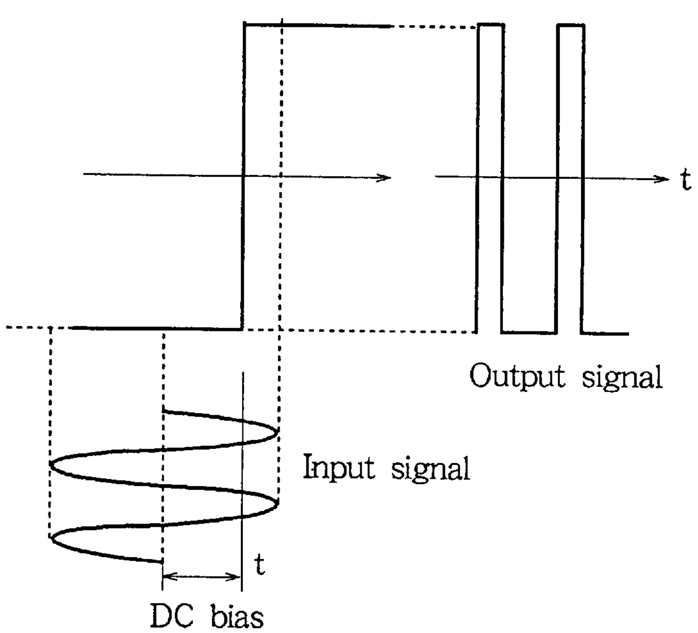

t
Output signal
Input signal
t
DC bias

FIG.16

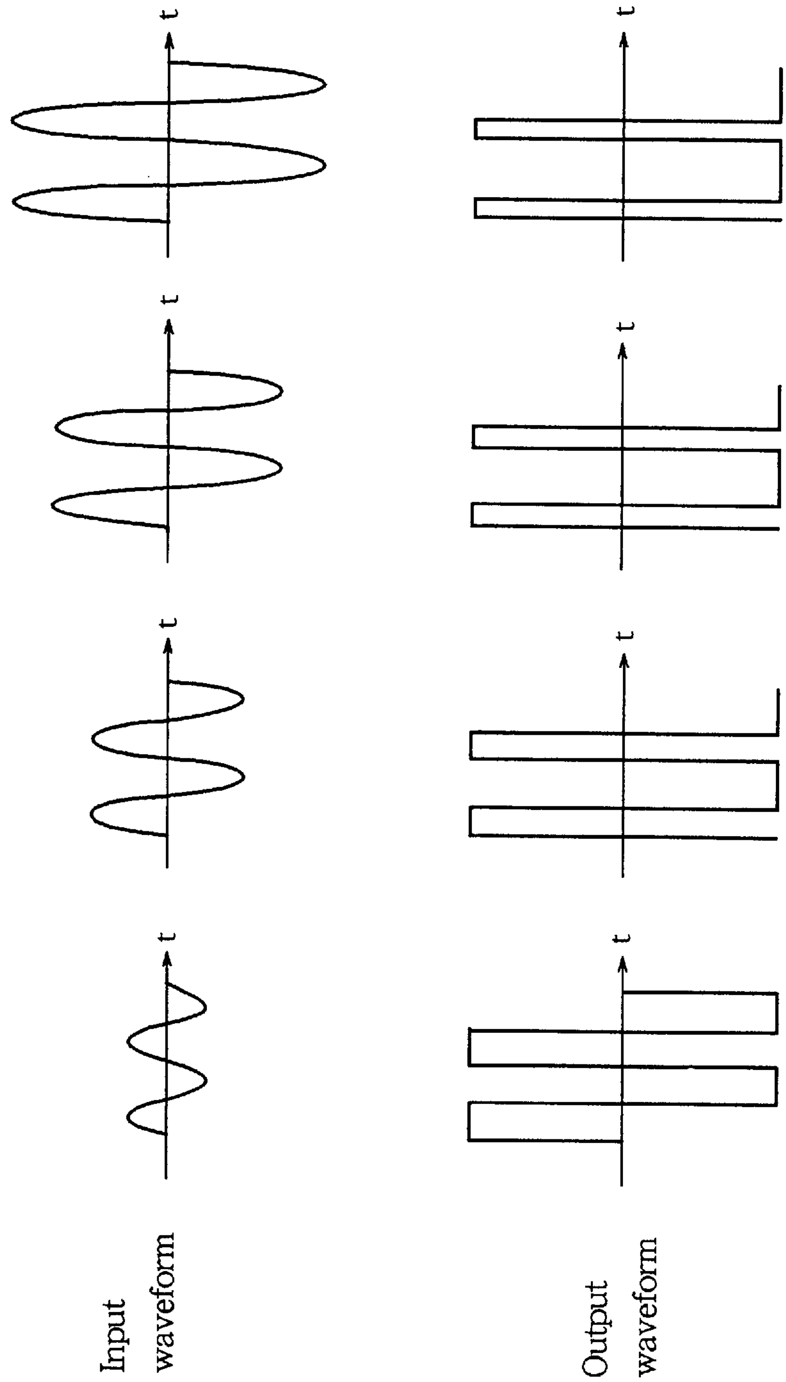


FIG. 17

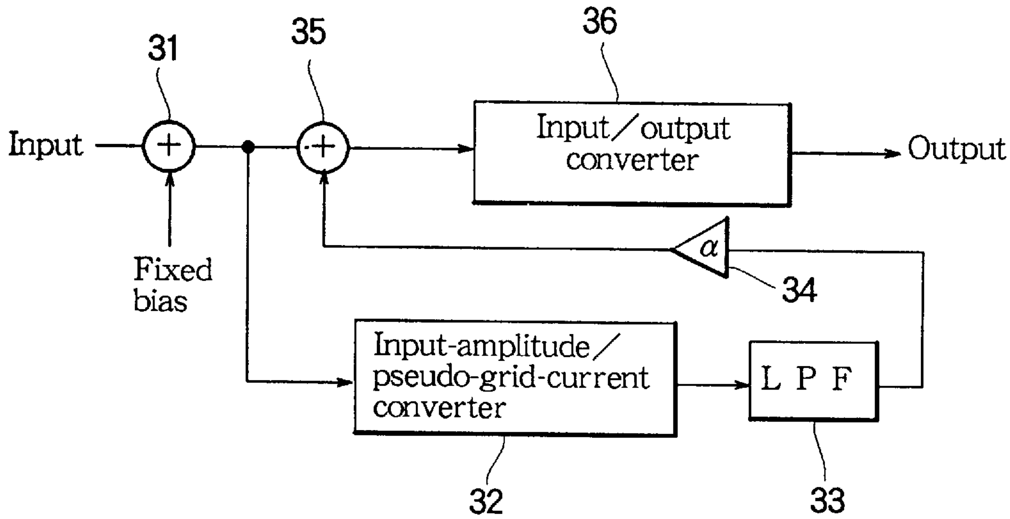

31
35
36
Input
Input／output converter
Output
Fixed bias
α
34
Input-amplitude／pseudo-grid-current converter
L P F
32
33

FIG. 18

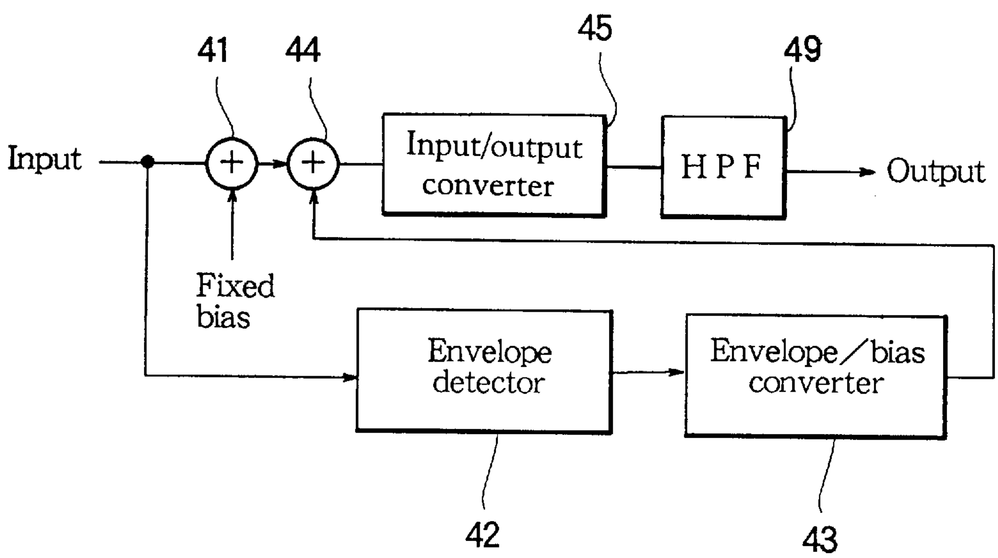

41
44
45
49
Input
Input/output converter
H P F
Output
Fixed bias
Envelope detector
Envelope／bias converter
42
43

FIG.19

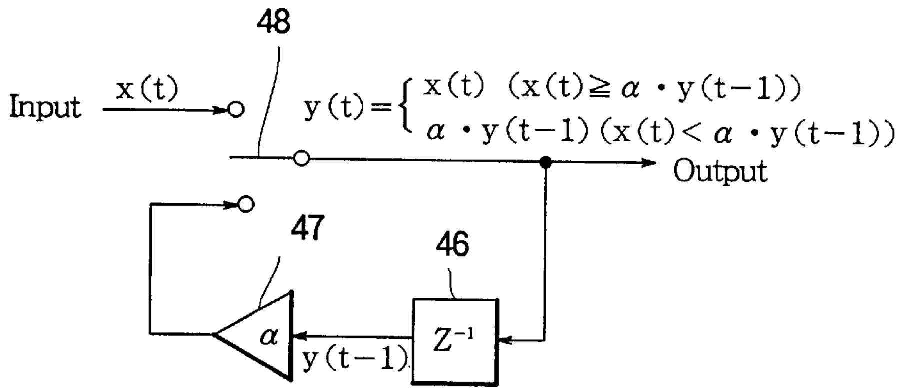


FIG.20

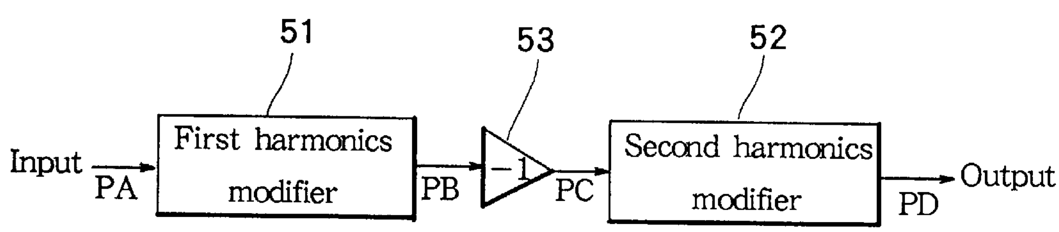


FIG.21

Input/Output Characteristic Curve

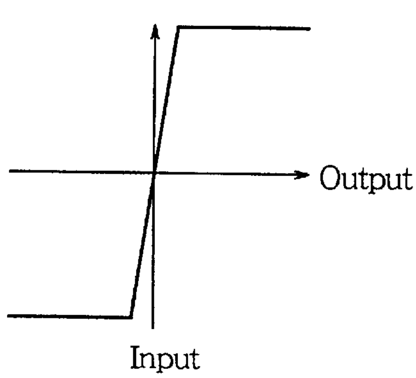


FIG.22

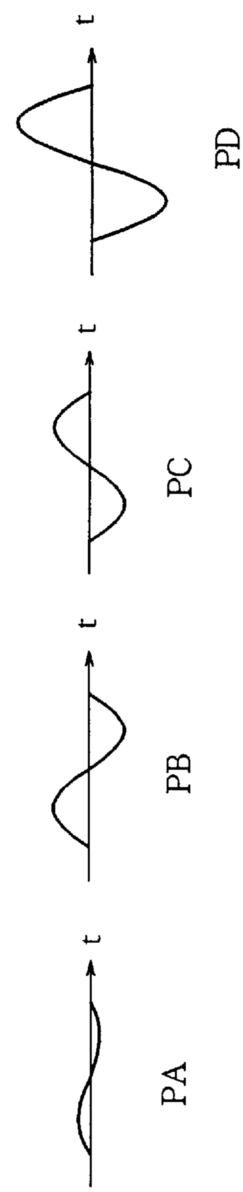


FIG.23

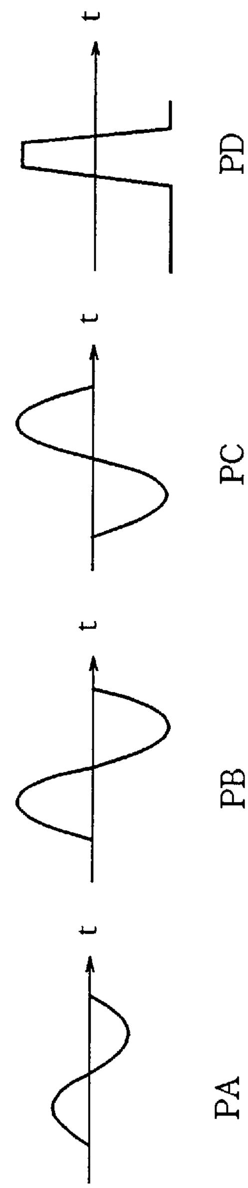


FIG.24

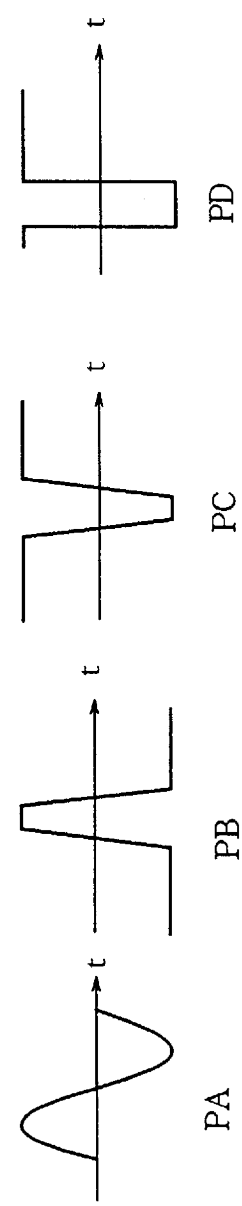


FIG. 25

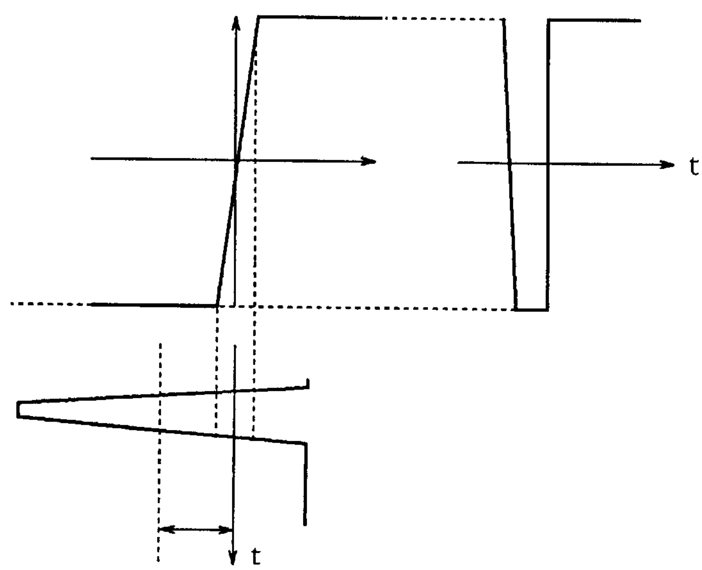

t
t
DC bias

FIG. 26

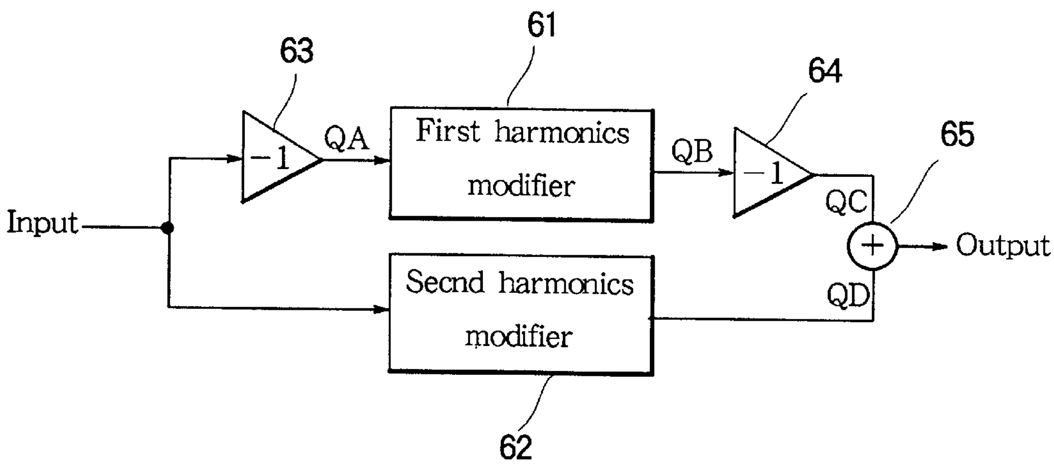

63
61
64
65
−1
QA
First harmonics
modifier
QB
−1
QC
Input
+
Output
Secnd harmonics
modifier
QD
62

FIG. 27

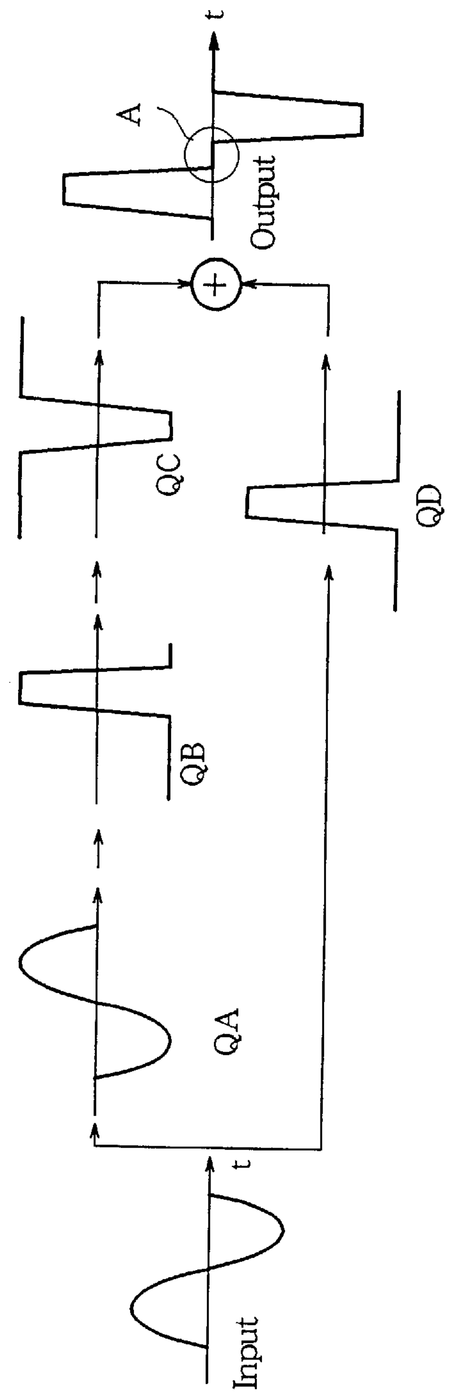


FIG.28

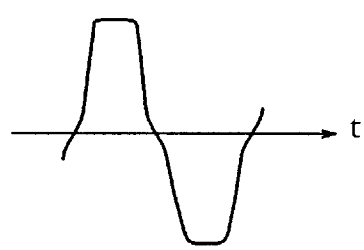
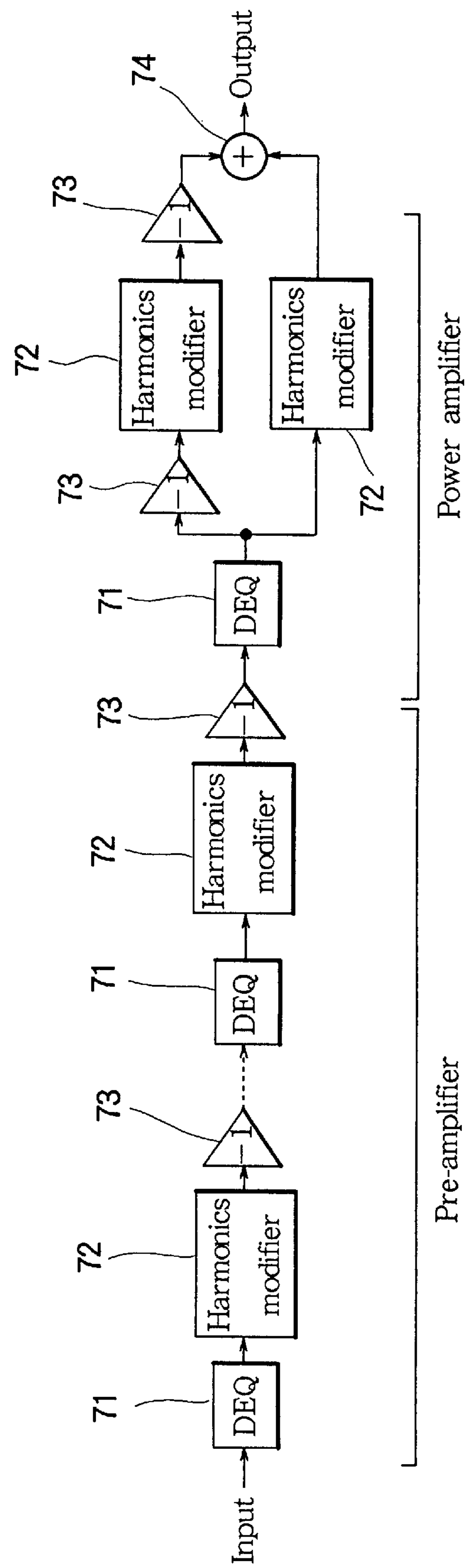


FIG.29


Input
71
DEQ
72
Harmonics modifier
73
−1
71
DEQ
72
Harmonics modifier
73
−1
71
DEQ
73
−1
72
Harmonics modifier
73
−1
72
Harmonics modifier
74
+
Output
Pre-amplifier
Power amplifier

DIGITAL AUDIO SIGNAL PROCESSOR WITH HARMONICS MODIFICATION

This is a continuation of application Ser. No. 07/968,539, filed Oct. 29, 1992, now abandoned.

BACKGROUND OF THE INVENTION

The present invention relates to an audio signal processor having a harmonics modifier used in an audio amplifier of musical instruments such as a guitar for generating harmonics modification effect, and particularly relates to a digital harmonics modifier of the type producing the harmonics modification effect according to digital signal processing.

For example, the musical instrument amplifier such as a guitar amplifier operates not only to amplify an input audio signal, but also to suitably mix harmonic components into the input audio signal to generate a distortion sound having improved timbre expression which is suitable for realistic sound generation of various musical instruments.

Conventionally, there have been adopted two different methods of imparting harmonic components to the audio signal in the musical instrument amplifier. The first method utilizes a vacuum-tube amplifier to generate distortion in an audio signal waveform according to vacuum-tube characteristics. The second method utilizes a distortion circuit composed of a semiconductor device for clipping the signal waveform and an equalizer for adjusting frequency characteristics, thereby generating distortion in the signal waveform.

With regard to the first method utilizing the vacuum-tube amplifier, while the generated distortion has particularly good quality as to timbre, the use of vacuum-tube causes degradation in the durability and reliability at the amplifier. With regard to the second method utilizing the distortion circuit composed of the semiconductor device, while the durability and reliability is superior to the first prior art, satisfactory timbre or quality of the musical sound cannot be obtained. Additionally, these conventional amplifiers are comprised of an analog circuit which can generate only a fixed pattern of tones. A plurality of analog circuits are needed in order to generate different types of tones.

SUMMARY OF THE INVENTION

In view of the above noted drawbacks of the prior art, an object of the present invention is to provide a digital harmonics modifier being capable of generating a variety of tones containing variable harmonic components and featuring high durability and reliability. According to the invention, the digital audio signal processor has a harmonics modifier for processing an input audio signal to produce an output audio signal. The harmonics modifier comprises translating means for sequentially translating the input audio signal into an address signal according to a sampled amplitude of the input audio signal, memory means for storing an input/output conversion table containing amplitude values in addressable manner, and conversion means for accessing the memory means in response to the address signal to effect input/output conversion to read out a sequence of the amplitude values which form the output audio signal containing desired harmonic components according to results of the input/output conversion. In a preferred form, the inventive harmonics modifier has bias means for adding a direct current (DC) bias to the input audio signal according to an amplitude level thereof. The input/output conversion table is addressed for the signal processing based on a level-shifted input audio signal added with the DC bias so as to vary the harmonic component distribution or frequency spectrum. In another preferred form, a plurality of harmonics modifiers are connected in series to one another through a phase inverter so that an output of a preceding modifier is phase-inverted, and is then inputted into a succeeding modifier. In a further preferred form, there are provided a plurality of harmonics modifiers which include mixed ones having phase-inverting means and having no phase-inverting means. These mixed modifiers are connected in parallel to each other.

According to the first aspect of the invention, the harmonics modifier utilizes the input/output conversion table to conduct digital signal processing so as to add desired harmonic components to an arbitrary audio signal, thereby improving durability and reliability of the device. According to the second aspect of the invention, an input audio signal is superimposed with a DC bias prior to the input/output conversion so as to introduce more complex harmonic components to improve significantly variation of the harmonics modification, thereby realizing sophisticated distortion sounds comparable to those produced by the vacuum-tube type audio amplifier. According to the third aspect of the invention, a plurality of the harmonics modifiers are connected in series to one another with intermediate phase-inversion to thereby effect more complicated modification of the audio sound. According to the fourth aspect of the invention, a plurality of the harmonics modifiers are connected in parallel to each other with phase-inversion to thereby enable more complicated modification of the audio sound having realistic timbres comparable to those produced by the vacuum-tube type audio amplifier.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. **1** is a block diagram showing a first embodiment of the inventive digital audio signal processor with a harmonics modifier.

FIG. **2** is a detailed block diagram showing structure of the harmonics modifier in the FIG. **1** embodiment.

FIGS. **3**(i a), **3**(*b*) and **3**(*c*) are diagrams showing harmonics distribution, output audio signal waveform and input/output linear characteristic curve, observed during the operation of the FIG. **1** embodiment in case that the input audio signal has a relatively small amplitude.

FIGS. **4**(*a*), **4**(*b*) and **4**(*c*) are similar diagrams showing the operation of the FIG. **1** embodiment in case that the input audio signal has an intermediate amplitude.

FIGS. **5**(*a*), **5**(*b*) and **5**(*c*) are similar diagrams showing the operation of the FIG. **1** embodiment in case that the input audio signal has a relatively great amplitude.

FIGS. **6**(*a*), **6**(*b*) and **6**(*c*) are diagrams showing harmonies distribution, output audio signal waveform and input/output nonlinear characteristic curve, observed during the operation of the FIG. **1** embodiment in case of a relatively small input amplitude.

FIGS. **7**(*a*), **7**(*b*) and **7**(*c*) are similar diagrams observed in case of an intermediate input amplitude.

FIGS. **8**(*a*), **8**(*b*) and **8**(*c*) are similar diagrams observed in case of a relatively great input amplitude.

FIG. **9** is a block diagram showing a second embodiment of the inventive digital audio signal processor having a harmonics modifier with DC bias application.

FIG. **10** is an equivalent circuit disgram showing the operation of DC bias calculation in the FIG. **9** embodiment.

FIG. **11** is a diagram showing a characteristic curve of the input/output conversion table provided in the FIG. **9** embodiment.

FIG. **12** is a diagram showing the operation of the input/output conversion in case that the input audio signal has a relatively small amplitude without DC bias in the FIG. **9** embodiment.

FIG. **13** is a similar operational diagram in case of an intermediate input amplitude without DC bias.

FIG. **14** is a similar operational diagram in case of a relatively great input amplitude without DC bias.

FIG. **15** is a similar operational diagram in case of a relatively great input amplitude with DC bias.

FIG. **16** is a diagram showing various input and output signal waveforms illustrating the typical operation of the FIG. **9** embodiment.

FIG. **17** is a block diagram showing an example of a DC bias application circuit adoptable in the FIG. **9** embodiment.

FIG. **18** is a block diagram showing another example of the DC bias application circuit adoptable in the FIG. **9** embodiment.

FIG. **19** is a block diagram showing an envelope detector provided in the FIG. **18** circuit.

FIG. **20** is a block diagram showing a third embodiment of the inventive digital audio signal processor.

FIG. **21** is a diagram showing a characteristic curve of the input/output conversion table provided in the FIG. **20** embodiment.

FIG. **22** is a diagram showing various signal waveforms illustrating the operation of the FIG. **20** embodiment in case of a relatively small input amplitude,

FIG. **23** is a similar waveform diagram observed in case of an intermediate input amplitude.

FIG. **24** is a similar waveform diagram observed in case of a relatively great input amplitude.

FIG. **25** is a diagram showing an input/output conversion characteristic of a second harmonics modifier provided in the FIG. **20** embodiment.

FIG. **26** is a block diagram showing a fourth embodiment of the inventive digital audio signal processor.

FIG. **27** is a diagram showing various waveforms illustrative of the operation of the FIG. **26** embodiment.

FIG. **28** is a diagram showing an output waveform of a vacuum-tube audio amplifier.

FIG. **29** is a block diagram showing a fifth embodiment of the inventive digital audio signal processor.

DETAILED DESCRIPTION OF EMBODIMENTS

Hereinafter, preferred embodiments of the invention will be described in conjunction with the drawings. FIG. **1** shows the first embodiment of the inventive digital audio signal processor, which features specific digital signal processing by accessing an input/output conversion table to generate harmonic components.

For the background information, in view of the good sound quality, an audio amplifier of a musical instrument is conventionally comprised of a vacuum-tube circuit, which characteristically features the following two aspects. Firstly, the vacuum-tube audio amplifier has a nonlinear input/output characteristic so that an output audio signal may contain harmonic components which are not contained in an input audio signal, even without clipping the output audio signal. Secondly, a grid current flows in the vacuum-tube as the input audio signal level increases so that a bias voltage superimposed to the input audio signal varies to thereby change harmonics distribution in the output audio signal. The first feature of the vacuum-tube amplifier may significantly improve richness and fullness when amplifying and reproducing a clear sound which initially contains no positive harmonics. The second feature can significantly improve stiffness and strength when amplifying and reproducing a distortion sound which initially contains positive harmonics.

Accordingly, the present invention is directed to digital audio signal processing technology effective to realize such features of the vacuum-tube amplifier so as to generate good quality sounds suitable for musical instrument power amplification. The first embodiment shown in FIG. **1** is constructed for this purpose. In the figure, an analog input audio signal is converted into a corresponding digital signal by an analog/digital converter or A/D converter **1**, and is then fed to a signal processing unit **2** which constitutes a harmonics modifier. The signal processing unit **2** carries out specific digital processing to produce a modified digital signal, which is then treated by a digital/analog converter or D/A converter **3** to produce a corresponding analog output audio signal. The signal processing unit **2** is controlled by a central processing unit or CPU **4**. CPU **4** is provided with a supplementary memory unit **5**. The signal processing unit **2** controlled by the CPU **4** converts the digital input signal sampled from the A/D converter **1** by using an input/output conversion table or look-up table stored in a random access memory or RAM **6** to effect the specific digital processing. The sampled data conversion of the input signal by the look-up table is undertaken such that the table RAM **6** is accessed sequentially by an address signal corrresponding to each sampled amplitude of the input audio signal to retrieve a sequence of registered amplitude values to provide an output digital signal. The contents of the input/output conversion table stored in the RAM **6** can be rewritten by the CPU **4** if desired. In case that there is no need to change the table contents, the look-up table may be stored in a Read-only memory or ROM.

The signal processing unit **2** is comprised of an amplitude/address translator **11**, a table reader **12** and an interpolator **13** as shown in FIG. **2**. The amplitude/address translator **11** translates an input sample data x representative of an amplitude of the input audio signal to a corresponding address signal of the table RAM **6**. The table reader **12** operates using this address signal to retrieve amplitude values yi and $yi+_1$ from the table RAM **6** at one sample sequence. These retrieved values yi and $yi+_1$ are fed to the interpolator **13** together with the input data x to effect interpolation of the first order to produce an output digital data y. In this embodiment, the linear interpolation is carried out by the Interpolator **13** to calculate the output digital data y in order to reduce a table size stored in the RAM **6**. In FIG. **2**, the input sample data is translated to a corresponding address value of the table RAM **6** by the amplitude/address translator **11**. The reader **12** retrieves a table value stored in the designated address from the table RAM **6**. In case that the RAM **6** stores a complete set of table values corresponding to the full range of the input sample data, the table size may be extremely expanded if each input data is composed of a great number of bits. In view of this, in the FIG. **2** structure, each sample data is divided into a set of higher order bits and another set of lower order bits. The higher bit set is used as the address value to access the conversion table to retrieve a pair of coarse table values. These table values are interpolated or truncated by the interpolator **13** based on the linear interpolation according to the divided lower bit set. Alternatively, the table RAM **6** may store fine table values corresponding directly to each of the sample data to enable direct input/output conversion, thereby eliminating the interpolator **13**.

FIGS. **3**–**8** show various operation modes of the FIG. **1** embodiment, where FIGS. **3**–**5** modes are associated to linear input/output conversion with small, intermediate and great input amplitude levels, respectively, and FIGS. **6**–**8** modes are associated to nonlinear input/output conversion with small, intermediate and great input amplitude levels, respectively. As understood from these figures, in case that the digital modifier composed of the signal processing unit **2** and the table RAM **6** etc. has the linear input/output conversion characteristic, the output audio signal does not contain a harmonic component as long as the output audio signal is not clipped as in the FIGS. **3** and **4** cases. On the other hand, when the signal waveform is clipped as in the FIG. **5** case, there are abruptly generated various harmonic components. However, by providing an adequate conversion table having a nonlinear input/output characteristic, harmonic components are generated even if the input amplitude level is relatively small as shown in FIGS. **6**–**8**. As the input amplitude level increases, the harmonic components are accordingly increased. These harmonic components may improve richness of reproduced clear sounds. Further, the use of the nonlinear conversion may avoid abrupt timbre change which would be caused by clipping.

Next, the description is given for the second embodiment of the inventive digital audio signal processor. This second embodiment is provided with an improved harmonies modifier in contrast to the first embodiment. The present embodiment features that a variable DC bias is applied to the input audio signal so as to produce various harmonic components from a prescribed conversion table. Further, the DC bias value is varied according to an amplitude level of the input audio signal so as to change harmonics distribution or frequency spectrum of the output audio signal.

Referring to FIG. **9**, the improved harmonics modifier is comprised of a first table reader/processor **21**, a first table RAM **22**, a second table reader/processor **23**, a second table RAM **24**, data delaying units **25**–**27**, a DC bias calculator **28** and adders **29**, **30**. The combination of the first table RAM **22** and the first table reader/processor **21** has the substantially same structure as the FIG. **1** harmonics modifier composed of the signal processing unit **2**, CPU **4**, supplementary memory unit **5** and table RAM **6**. In the present embodiment, an initial input audio signal is converted into a digital form already by an A/D converter (not shown), and an output signal is given in a digital form which may be treated by a D/A converter it necessary in sound reproduction. The first table RAM **22** stores an input/output conversion table similar to that stored in the table RAM **6** of FIG. **1**. The second table RAM **24** stores an amplitude/bias conversion table which represents or simulates an input-current/grid-current characteristic of a conventional vacuum-tube. The DC bias calculator **28** carries out the following Euler's integration to calculate a DC bias value Vc(t) based on those of a data Vin(t−1) which is obtained by delaying an input data Vin(t) through the delay circuit **25** by one sample timings another data Ig(t−1) which is obtained by delaying an output data Ig(t) of the second table reader/processor **23** through the delay circuit **26** by one sample timing, and a further data Vc(t−1) which is obtained by delaying the output data Vc(t) itself of the DC bias calculator **28** through the delay circuit **27** by one sample timing.

The Euler's integration is represented by:

$$Vc(t)=\{Ig(t-1)+[(Vin(t-1)-Vc(t-1))/Rg]\}/C\cdot Fs$$

where Fs denotes a sampling frequency.

The DC bias value calculated by the Euler's integration corresponds to a terminal voltage Vc of a capacitor C when feeding an input voltage Vin to an equivalent circuit of FIG. **10** which simulates a vacuum-tube and which comprises the capacitor C, a resistor Rg and an electric current source Ig. Referring back to FIG. **9**, the input signal is added with a fixed bias by the adder **29**, and is further added with the DC bias Vc(t) calculated by the DC bias calculator **28** through the other adder **30** to thereby provide the digital input Vin(t) which is given to the first table reader/processor **21**. By such operation, the table reader/processor **21** sequentially retrieves an output data stored in the table RAM **22** according to the input Vin(t) added with the DC bias to produce the digital output signal. In this embodiment, the input signal is provisionally added with the fixed bias which is set to a center level of a desired bias range. Further, the DC bias is added, which varies around the center level according to the amplitude of the input signal.

Next, the description is given for the operation of the FIG. **9** embodiment with reference to FIGS. **11**–**16**. For facilitating better understanding of the operation, the first table RAM **22** stores an input/output conversion table having a particular characteristic curve shown illustratively in FIG. **11**, which simulates a comparator. The input/output conversion characteristic curve of FIG. **11** has an extremely great distortion gain such that the signal waveform is completely clipped. In such a conversion characteristic, as shown in FIGS. **12**–**14**, the output signal waveform is all the same while the amplitude level of the input signal varies provided that the harmonies modifier is not provided with DC bias means.

As opposed to the FIG. **14** condition, when an adequate DC bias is added to the input signal as shown in FIG. **15**, there can be obtained a different output signal waveform from the same input/output conversion table. By utilizing such feature, the DC bias voltage applied to the input signal is varied according to the amplitude level of the input signal so as to further change timbre of the distortion sound having a clipped waveform. For example, FIG. **16** shows a preferred situation where the DC bias is increased as the input signal amplitude is raised. In general, the manner of varying the DC bias value is controlled by the contents of the amplitude/bias conversion table stored in the second table RAM **24**, and values of the capacitor C and the resistor Rg in the DC bias means. Timbre variation due to change of the output waveform can greatly improve strength of the distortion sound in the audio amplifier of a musical instrument.

There may be other methods of varying the DC bias according to the input signal amplitude. For example, FIG. **17** shows one expedient where a low-pass filter (LPF) **33** is utilized to vary the DC bias. In this structure, a fixed bias is added precedingly to the input signal by an adder **31**. Then, the input signal is treated by an input-amplitude/pseudo-grid-current converter **32** to produce a pseudo-grid-current by utilizing a look-up table which represents a certain input-current/grid-current characteristic of the vacuum-tube. The pseudo-grid-current is filtered by the LPF **33** to extract a direct current (DC) component. The DC component is multiplied with a given coefficient α by a multiplier **34** to provide a DC bias, which is added to the input signal by an adder **35**. This DC-based input signal is processed by the harmonics modifier in the form of an input/output converter **36** utilizing an adequate input/output conversion look-up table to thereby produce an output signal.

FIG. **18** shows another example where an envelope of the input signal is detected for varying a DC bias accordingly. In this structure, an evelope detector **42** is provided to detect the envelope of the initial input signal prior to addition of a fixed bias or offset by an adder **41**. The detected envelope data is converted into a DC bias by an envelope/bias converter **43** utilizing an adequate look-up table. The DC bias is added by an adder **44** to the input signal which is precedingly added with the fixed bias by the adder **41**. The DC-biased input signal is treated by an input/output converter **45** utilizing a given look-up table to thereby produce an output signal modified with harmonics.

FIG. **19** shows a detailed structure of the envelope detector **42** and illustrates an envelope detection process. Namely, the input signal x(t) is passed as it is to form the output signal of $y(t)=x(t)$ in case of $x(t)\geqq\alpha\cdot y(t-1)$. On the other hand, in case of $x(t)<\alpha\cdot y(t-1)$, the envelope detector produces the output signal of $y(t)=\alpha\cdot y(t-1)$. The value of a $\alpha\cdot y(t-1)$ is obtained by delaying one sample timing the output signal y(t) through a delay circuit **46** and then by multiplying the same by the given factor α through a multiplier **47**. A selector **48** operates according to the above condition to selectively output either of $\alpha\cdot y(t-1)$ and x(t).

The harmonics modification is carried out by the input/output converter **45** after the input signal is added with the DC bias. In such a case, there might be caused drawbacks that a DC offset is transmitted to a subsequent device which utilizes the output signal. In order to avoid the drawbacks, the improved harmonics modifier having the DC bias means is provided at its output side with means for removing a DC component. For example, a high-pass filter (HPF) **49** is adopted to cut off the DC component.

Next, the description is given for the third embodiment of the inventive digital audio signal processor, which is provided with a plurality of harmonics modifiers connected in series to one another while an intermediate signal is phase-inverted to thereby vary a distribution of harmonic components. Referring to FIG. **20**, the digital audio signal processor is comprised of first and second harmonics modifiers **51**, **52** each having the substantially same construction as that of the FIG. **9** modifier, and a phase inverter **53** interposed between the first and second harmonics modifiers. The first and second harmonics modifiers **51**, **52** are connected in series to each other through the phase inverter **53** so as to more efficiently control variation of timbre.

Next, the description is given for operation of the FIG. **20** embodiment of the double stage construction in conjunction with FIGS. **21–25**. The pair of harmonics modifiers **51**, **52** utilize a given input/output conversion table having a conversion characteristic curve shown in FIG. **21**. Namely, the utilized look-up table has a linear characteristic with a certain gain in a relatively small input range and a flat characteristic in a relatively great input range so as to clip an output waveform.

Firstly, FIG. **22** shows signal waveforms observed in case of a relatively small input level at various nodes of the FIG. **20** circuit, including an input node PA, an output node PB of the first harmonics modifier **51**, an output node PC of the phase inverter **53**, and a final output node PD of the second harmonics modifier **52**.

Next, FIG. **23** shows signal waveforms observed at the various nodes PA, PB, PC and PD in case that the input level moderately increases so that clipping and DC bias variation are occurring in the second harmonics modifier **52**. Further, FIG. **24** shows signal waveforms observed in case that the input level considerably increases so that the clipping and the DC bias variation are occurring also in the first harmonics modifier **51**.

FIG. **25** shows the operation state of the second harmonics modifier **52** in the FIG. **24** condition. As understood from this figure, when the second harmonica modifier receives an input signal having the clipped waveform, the DC bias application effect does not occur substantially. Accordingly in the series connection of the hamonics modifiers, the preceding modifier has a greater DC bias application effect than that of the succeeding modifier. Further, variations of a duty ratio of the modified waveforms are opposite to each other between the preceding and succeeding stages because of the intermediate phase inversion. Consequently, in the series connection structure, the succeeding stage changes the duty ratio of the output waveform according to the DC bias application when the input level is relatively low. Then, the duty ratio can be restored by the DC bias application effect of the preceding stage in response to the input level. Further, the duty ratio can be changed reversely when the input level goes high. The above described harmonics modification phases are substantially separately controlled by corresponding stages of the modifiers to thereby facilitate the harmonics control. In addition, a number of stages may be increased to effect more delicate harmonics control.

Next, the description is given for the fourth embodiment of the inventive digital audio signal processor, which comprises at least a pair of harmonics modifiers having the substantially same structure as those of the first and second embodiments. One of the modifiers has inverted input and output terminals, and another modifier is connected in parallel to that inversion type modifier. Referring to FIG. **26**, a pair of first and second harmonics modifiers **61**, **62** having the same structure as the FIG. **9** embodiment are coupled in parallel to each other. A pair of phase inverters **63**, **64** are connected to input and output terminals of the first harmonics modifier **61**. respectively An adder **65** is connected to the output of the phase inverter **64** and to the output of the second harmonics modifier **62**.

The recognized good quality of the distortion sound by the vacuum-tube amplifier partly originates from a distortion by a power amplifier. The power amplifier is generally comprised of a push/pull circuit. In view of this, the fourth embodiment is constructed as shown in FIG. **26** to effect digital push/pull operation in order to realize such a good quality of the conventional vacuum-tube amplifier. FIG. **27** shows signal waveforms observed during the course of distortion sound reproduction in the FIG. **26** embodiment at various points including start input node, output node QA of the phase inverter **63**, output node QB of the first harmonics modifier **61**, output node QC of the phase inverter **64**, output node QD of the second harmonics modifier **62** and final output node. In this operation, the signal waveform at the final output node has a step portion A, which significantly improves quality of the distortion sound. For the simplicity of the description of the present embodiment, the utilized input/output conversion table has a characteristic curve as shown in FIG. **21**, hence the final output waveform has the rather emphasized step portion A. In the practical vacuum-tube amplifier, the output waveform has a moderate and smooth step portion as shown in FIG. **28**. Such a smooth step portion can be realized by suitably setting the contents of the input/output conversion table. Since the step portion is generated by varying the DC bias added to the input signal, it may be possible to eliminate the step portion when the input signal is relatively small. Such a control may be effected by suitably altering the contents of the amplitude/bias conversion table.

Lastly, the description is given for the fifth embodiment of the invention, which is a combination of the FIG. **20** construction and the FIG. **26** construction with digital equalizers (DEQ) to constitute an amplifier simulation circuit as shown in FIG. **29**. The simulation circuit is comprised of digital equalizers **71**, harmonics modifiers **72**, phase inverters **73** and adder **74**. The FIG. **29** circuit is divided into a pre-amplifier composed of a serial connection similar to the FIG. **20** embodiment and a power amplifier composed of the last stage equalizer **71** and the parallel connection of a push/pull circuit similar to the FIG. **26** embodiment. This structure can simulate various types of conventional amplifiers by controlling parameters, input/output conversion table, amplitude/bias conversion table and equalizer characteristics.

The present invention is not limited to the above described embodiments, but may be applied to various modifications to produce new interesting sounds. For example, the inventive harmonics modifier can be combined to the conventional distortion circuit such as the vacuum-tube amplifier and the semiconductor waveform clipper.

As described above, according to the invention, the digital audio signal processor can form interesting sounds containing various harmonic components. Further, the inventive harmonics modifier has high reliability and durability.

What is claimed is:

1. An audio signal processor having a harmonies modifier for processing an input audio signal to produce an output audio signal, wherein the harmonics modifier comprises:

translating means for sequentially translating the input audio signal into an address signal according to a sampled amplitude of the input audio signal;

memory means for storing an input/output conversion table containing amplitude values in addressable manner;

conversion means for accessing the memory means in response to the address signal to effect input/output conversion to read out a sequence of the amplitude values which form the output audio signal containing desired harmonic components according to results of the input/output conversion; and

DC bias means disposed upstream of the translating means for adding a variable DC bias to the input audio signal, the variable DC bias being varied according to an amplitude level of the input signal, thereby varying a distribution of the harmonic components contained in the output audio signal,

wherein the DC bias means includes an amplitude/bias conversion table used for conversion of the amplitude level of the input audio signal into the variable DC bias, the amplitude/bias conversion table representing an input-voltage/grid-current characteristic of a vacuum-tube amplifier.

2. An audio signal processor having a harmonics modifier for processing an input audio signal to produce an output audio signal, wherein the harmonics modifier comprises:

translating means for sequentially translating the input audio signal into an address signal according to a sampled amplitude of the input audio signal;

memory means for storing an input/output conversion table containing amplitude values in an addressable manner;

conversion means for accessing the memory means in response to the address signal to effect input/output conversion to read out a sequence of the amplitude values which form the output audio signal containing desired harmonic components according to results of the input/output conversion;

DC bias means disposed upstream of the translating means for adding a variable DC bias to the input audio signal, the variable DC bias being varied according to an amplitude level of the input signal, thereby varying a distribution of the harmonic components contained in the output audio signal;

a plurality of harmonics modifiers connected in series to one another; and

an inverter interposed between preceding and succeeding harmonics modifiers for inverting a phase of a transmitted signal from the preceding harmonics modifier to the succeeding harmonics modifier.

3. An audio signal processor having a harmonics modifier for processing an input audio signal to produce an output audio signal, wherein the harmonics modifier comprises:

translating means for sequentially translating the input audio signal into an address signal according to a sampled amplitude of the input audio signal;

memory means for storing an input/output conversion table containing amplitude values in an addressable manner;

conversion means for accessing the memory means in response to the address signal to effect input/output conversion to read out a sequence of the amplitude values which form the output audio signal containing desired harmonic components according to results of the input/output conversion;

DC bias means disposed upstream of the translating means for adding a variable DC bias to the input audio signal, the variable DC bias being varied according to an amplitude level of the input signal, thereby varying a distribution of the harmonic components contained in the output audio signal; and

at least one pair of harmonics modifiers connected in parallel to each other, one of the at least one pair of harmonics modifier having phase-inverters connect at its input and output terminals, and another of the at least one pair of harmonics modifier having no phase-inverter.

4. An audio signal processor according to claim **1**, further including a first plurality of harmonics modifiers connected in series to one another, and a second plurality of harmonics modifiers connected in parallel to one another.

5. An audio signal processor having a harmonics modifier for processing an input audio signal to produce an output audio signal, wherein the harmonics modifier comprises:

translating means for sequentially translating the input audio signal into an address signal according to a sampled amplitude of the input audio signal;

memory means for storing an input/output conversion table containing amplitude values in an addressable manner;

conversion means for accessing the memory means in response to the address signal to effect input/output conversion to read out a sequence of the amplitude values which form the output audio signal containing desired harmonic components according to results of the input/output conversion;

DC bias means disposed upstream of the translating means for adding a variable DC bias to the input audio signal, the variable DC bias being varied according to an amplitude level of the input signal, thereby varying a distribution of the harmonic components contained in the output audio signal; and

a serial connection of harmonics modifiers and equalizers to form a pre-amplifier; and

a parallel connection of harmonics modifiers having selectively inverted terminals and non-inverters terminals to form a power amplifier, wherein the preamplifier and the power amplifier are coupled to each other to constitute an amplification simulator.

6. A digital audio signal processor having a harmonics modifier for processing a digital input audio signal to produce a digital output audio signal, wherein the harmonics modifier comprises:

sampling means operative at a sampling frequency for sequentially sampling an input amplitude level of the digital input audio signal at each sample timing;

DC bias computing means for computing a variable DC bias based on a predetermined formula according to the input amplitude level sampled by the sampling means;

adding means for adding the computed DC bias to the digital input audio signal;

nonlinear conversion means for nonlinearly converting the digital input audio signal added with the computed DC bias into the digital output audio signal;

a plurality of harmonics modifiers connected in series to one another; and

an inverter interposed between preceding and succeeding modifiers for inverting a phase of a transmitted signal from the preceding harmonics modifiers to the succeeding harmonics modifier.

7. A digital audio signal processor according to claim **6**, further including means for adding a fixed DC bias to the digital input audio signal.

8. A digital audio signal processor having a harmonics modifier for processing a digital input audio signal to produce a digital output audio signal, wherein the harmonics modifier comprises:

sampling means operative at a sampling frequency for sequentially sampling an input amplitude level of the digital input audio signal at each sample timing;

DC bias computing means for computing a variable DC bias based on a predetermined formula according to the input amplitude level sampled by the sampling means;

adding means for adding the computed DC bias to the digital input audio signal;

nonlinear conversion means for nonlinearly converting the digital input audio signal added with the computed DC bias into the digital output audio signal; and

a pair of harmonics modifiers connected in parallel to each other, one harmonics modifiers having phase-inverters connected at its input and output terminals, and another harmonics modifier having no phase-inverter.

9. A digital audio signal processor according to claim **8**, further including means for adding a fixed DC bias to the digital input audio signal.

10. A digital audio signal processor having a harmonics modifier for processing a digital input audio signal to produce a digital output audio signal, wherein the harmonics modifier comprises:

sampling means operative at a sampling frequency for sequentially sampling an input amplitude level of the digital input audio signal at each sample timing;

DC bias computing means for computing a variable DC bias based on a predetermined formula according to the input amplitude level sampled by the sampling means;

adding means for adding the computed DC bias to the digital input audio signal;

nonlinear conversion means for nonlinearly converting the digital input audio signal added with the computed DC bias into the digital output audio signal;

a first plurality of harmonics modifiers connected in series to one another; and

a second plurality of harmonics modifiers connected in parallel to one another.

11. A digital audio signal processor according to claim **10**, further including means for adding a fixed DC bias to the digital input audio signal.

12. A method of modifying an audio signal with harmonic components, comprising the steps of:

preparing an input/output conversion table containing amplitude values in addressable manner;

setting a variable DC bias in accordance with an amplitude of an input audio signal;

adding the variable DC bias to the input audio signal;

sequentially translating the biased input audio signal into an address signal according to a sampled amplitude of the biased input audio signal; and

accessing the input/output conversion table in response to the address signal to read out a sequence of the amplitude values which form an output audio signal containing desired harmonic components,

wherein the variable DC bias circuit includes an amplitude/bias conversion table used for conversion of the amplitude level of the input audio signal into the variable DC bias, the amplitude/bias conversion table representing an input-voltage/grid-current characteristic of a vacuum-tube amplifier.

13. An audio signal processor having a harmonics modifier for processing an input audio signal to produce an output audio signal, wherein the harmonics modifier comprises:

a variable DC bias circuit that determines a variable DC bias based upon the amplitude of the input audio signal and a previous bias, the variable DC bias circuit then adding the variable DC bias to the input audio signal to produce an adjusted input signal;

a processor circuit that converts the adjusted input signal into an address signal in accordance with an amplitude of the adjusted input signal;

a memory device containing an input/output conversion table formed by a plurality of amplitude conversion values; and

a conversion circuit that accesses the memory device based upon the address signal to effect an input/output conversion by obtaining one of the plurality of amplitude conversion values from the memory device, the conversion circuit then outputting the obtained one of the plurality of amplitude conversion values as the output audio signal containing the desired harmonic components that are produced in accordance with the input/output conversion,

wherein the variable DC bias circuit includes an amplitude/bias conversion table used for conversion of the amplitude level of the input audio signal into the variable DC bias, the amplitude/bias conversion table representing an input-voltage/grid-current characteristic of a vacuum-tube amplifier.

14. An audio signal processor having a harmonics modifier for processing an input audio signal to produce an output audio signal, wherein the harmonics modifier comprises:

a variable DC bias circuit that determines a variable DC bias based upon the amplitude of the input audio signal and a previous bias, the variable DC bias circuit then adding the variable DC bias to the input audio signal to produce an adjusted input signal;

a processor circuit that converts the adjusted input signal into an address signal in accordance with an amplitude of the adjusted input signal;

a memory device containing an input/output conversion table formed by a plurality of amplitude conversion values;

a conversion circuit that accesses the memory device based upon the address signal to effect an input/output conversion by obtaining one of the plurality of amplitude conversion values from the memory device, the conversion circuit then outputting the obtained one of the plurality of amplitude conversion values as the output audio signal containing the desired harmonic components that are produced in accordance with the input/output conversion;

a plurality of harmonics modifiers connected in series to one another; and

an inverter interposed between preceding and succeeding harmonics modifiers for inverting a phase of a transmitted signal from the preceding harmonics modifier to the succeeding harmonics modifier.

15. An audio signal processor having a harmonics modifier for processing an input audio signal to produce an output audio signal, wherein the harmonics modifier comprises:

a variable DC bias circuit that determines a variable DC bias based upon the amplitude of the input audio signal and a previous bias, the variable DC bias circuit then adding the variable DC bias to the input audio signal to produce an adjusted input signal;

a processor circuit that converts the adjusted input signal into an address signal in accordance with an amplitude of the adjusted input signal;

a memory device containing an input/output conversion table formed by a plurality of amplitude conversion values;

a conversion circuit that accesses the memory device based upon the address signal to effect an input/output conversion by obtaining one of the plurality of amplitude conversion values from the memory device, the conversion circuit then outputting the obtained one of the plurality of amplitude conversion values as the output audio signal containing the desired harmonic components that are produced in accordance with the input/output conversion; and

at least one pair of harmonics modifiers connected in parallel to each other, one of the at least one pair of harmonics modifier having phase-inverters connect at its input and output terminals, and another of the at least one pair of harmonics modifier having no phase-inverter.

16. An audio signal processor according to claim **11**, further including a first plurality of harmonics modifiers connected in series to one another, and a second plurality of harmonic modifiers connected in parallel to one another.

17. An audio signal processor having a harmonics modifier for processing an input audio signal to produce an output audio signal, wherein the harmonics modifier comprises:

a variable DC bias circuit that determines a variable DC bias based upon the amplitude of the input audio signal and a previous bias, the variable DC bias circuit then adding the variable DC bias to the input audio signal to produce an adjusted input signal;

a processor circuit that converts the adjusted input signal into an address signal in accordance with an amplitude of the adjusted input signal;

a memory device containing an input/output conversion table formed by a plurality of amplitude conversion values; and

a conversion circuit that accesses the memory device based upon the address signal to effect an input/output conversion by obtaining one of the plurality of amplitude conversion values from the memory device, the conversion circuit then outputting the obtained one of the plurality of amplitude conversion values as the output audio signal containing the desired harmonic components that are produced in accordance with the input/output conversion;

a serial connection of harmonics modifiers and equalizers to form a pre-amplifier; and

a parallel connection of harmonics modifiers having selectively inverted terminals and non-inverted terminals to form a power amplifier, wherein the pre-amplifier and the power amplifier are coupled to each other to constitute an amplification simulator.

18. An audio signal processor according to claim **1**, wherein the conversion means for the harmonics modifier includes reading means for interpolatively reading out the amplitude.

19. An audio signal processor according to claim **1**, wherein the DC bias means further includes means for adding a fixed DC bias to the input audio signal.

20. An audio signal processor according to claim **1**, wherein the harmonics modifier further includes removing means disposed downstream of the conversion means for removing a DC component from the output audio signal so as to prevent an affect of a DC offset.

21. An audio signal processor according to claim **20**, wherein the removing means comprises a high-pass filter.

22. An audio signal processor according to claim **11**, wherein the conversion circuit for the harmonics modifier includes an interpolation circuit for interpolatively reading out the amplitude value.

23. An audio signal processor according to claim **11**, wherein the variable DC bias circuit further includes an input to add a fixed DC bias to the input audio signal.

24. An audio signal processor according to claim **11**, wherein the harmonics modifier further includes a removing circuit disposed downstream of the conversion circuit for removing a DC component from the output audio signal so as to prevent an affect of a DC offset.

* * * * *

United States Patent [19]

Brown, Sr. et al.

[11] **Patent Number: 4,811,401**

[45] **Date of Patent: Mar. 7, 1989**

[54] **SUPERDISTORTED AMPLIFIER CIRCUITRY WITH NORMAL GAIN**

[75] Inventors: **James W. Brown, Sr.; Jack C. Sondermeyer,** both of Meridian, Miss.

[73] Assignee: **Peavey Electronics Corporation,** Meridian, Miss.

[21] Appl. No.: **207,926**

[22] Filed: **Jun. 14, 1988**

Related U.S. Application Data

[63] Continuation of Ser. No. 63,924, Jun. 19, 1987, abandoned.

[51] **Int. Cl.**[4] **H03G 3/00**
[52] **U.S. Cl.** **381/61;** 84/1.19
[58] **Field of Search** 381/61, 98; 84/1.16, 84/1.19

[56] **References Cited**

U.S. PATENT DOCUMENTS

3,663,735	5/1972	Evans .	
3,860,876	1/1975	Woods .	
4,150,253	4/1979	Knoppel	381/101
4,290,335	9/1981	Sondermeyer .	
4,405,832	9/1983	Sondermeyer	381/61
4,479,238	10/1984	Spector .	
4,584,700	4/1986	Scholz	381/61

OTHER PUBLICATIONS

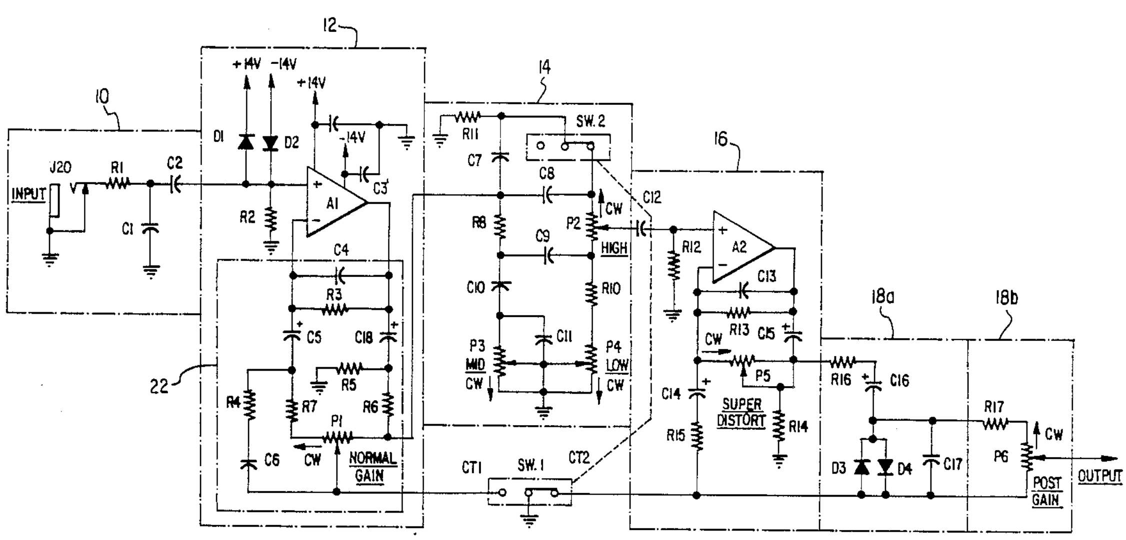
Vox Distortion Boaster V8162 (Thomas Organ Co.) 1969.

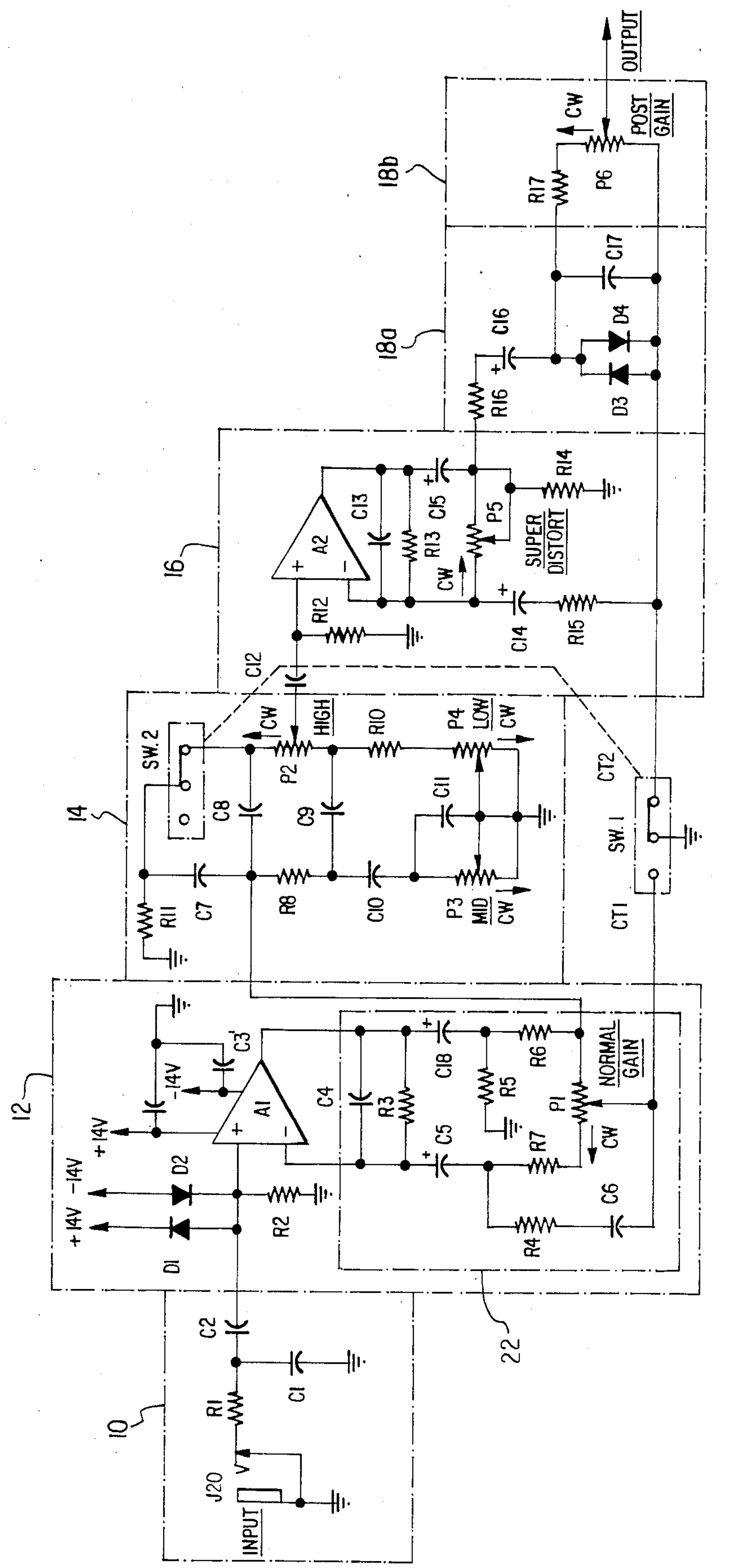
Primary Examiner—Forester W. Isen
Attorney, Agent, or Firm—Huff & Associates

[57] **ABSTRACT**

An amplifier circuit for use with a guitar. It has two modes, a "clean" output mode, and a "super-distortion" mode, which mode has independent gain control prior to a distortion stage. There is also post distortion gain control.

39 Claims, 5 Drawing Sheets





OUTPUT
POST GAIN
CW
P6
R17
18b
C17
18a
C16
D4
D3
R16
R14
C15
P5
SUPER DISTORT
C13
R13
A2
16
R12
C14
R15
C12
SW.2
HIGH
P2
R10
P4
LOW
14
C8
C9
C11
CT2
SW.1
CT1
R11
C7
R8
C10
P3
MID
12
C3
-14V
+14V
A1
C4
R3
C18
R5
R6
NORMAL GAIN
P1
C5
R7
D2
R2
C6
R4
D1
22
C2
C1
R1
10
J20
INPUT

FIG. 1

FIG. 2a

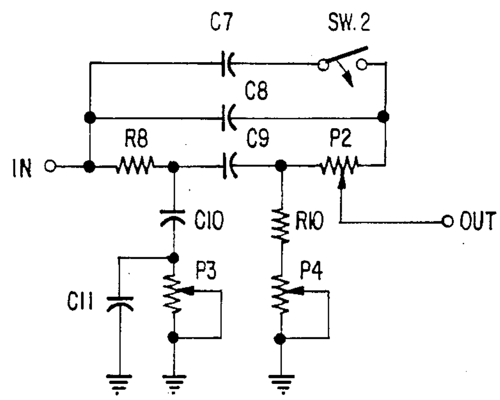


FIG. 4

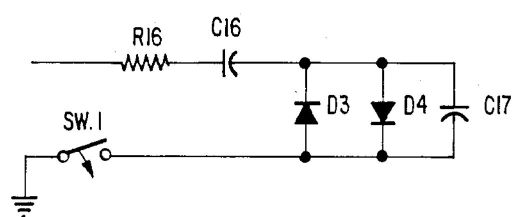
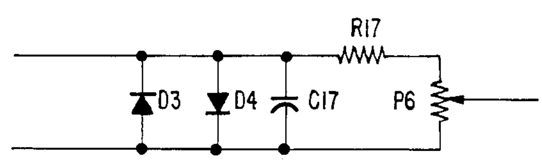


FIG. 5



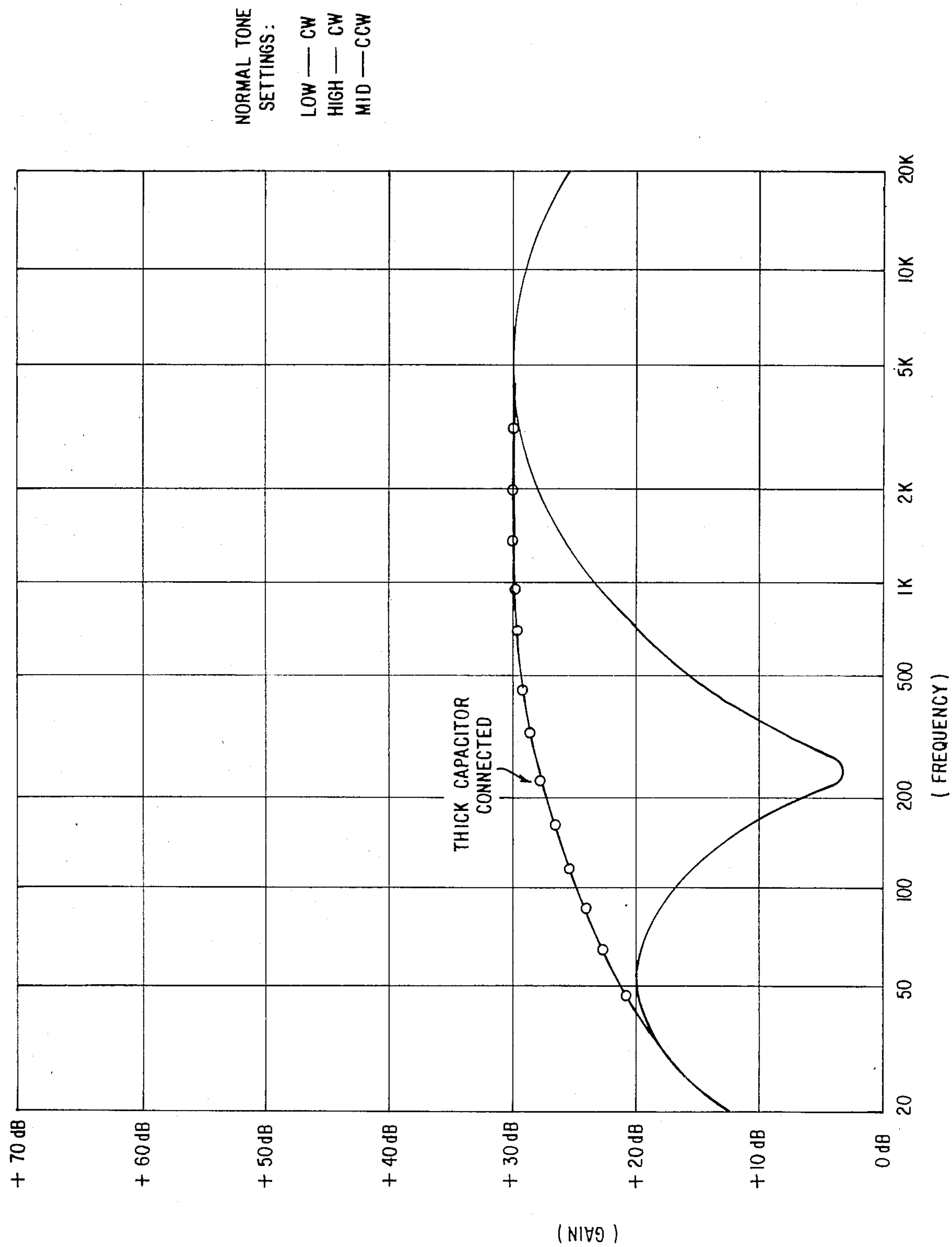

NORMAL TONE SETTINGS:
LOW — CW
HIGH — CW
MID — CCW
+ 70 dB
+ 60 dB
+ 50 dB
+ 40 dB
+ 30 dB
+ 20 dB
+ 10 dB
0 dB
(GAIN)
20
50
100
200
500
1K
2K
5K
10K
20K
(FREQUENCY)
THICK CAPACITOR CONNECTED

FIG. 2b

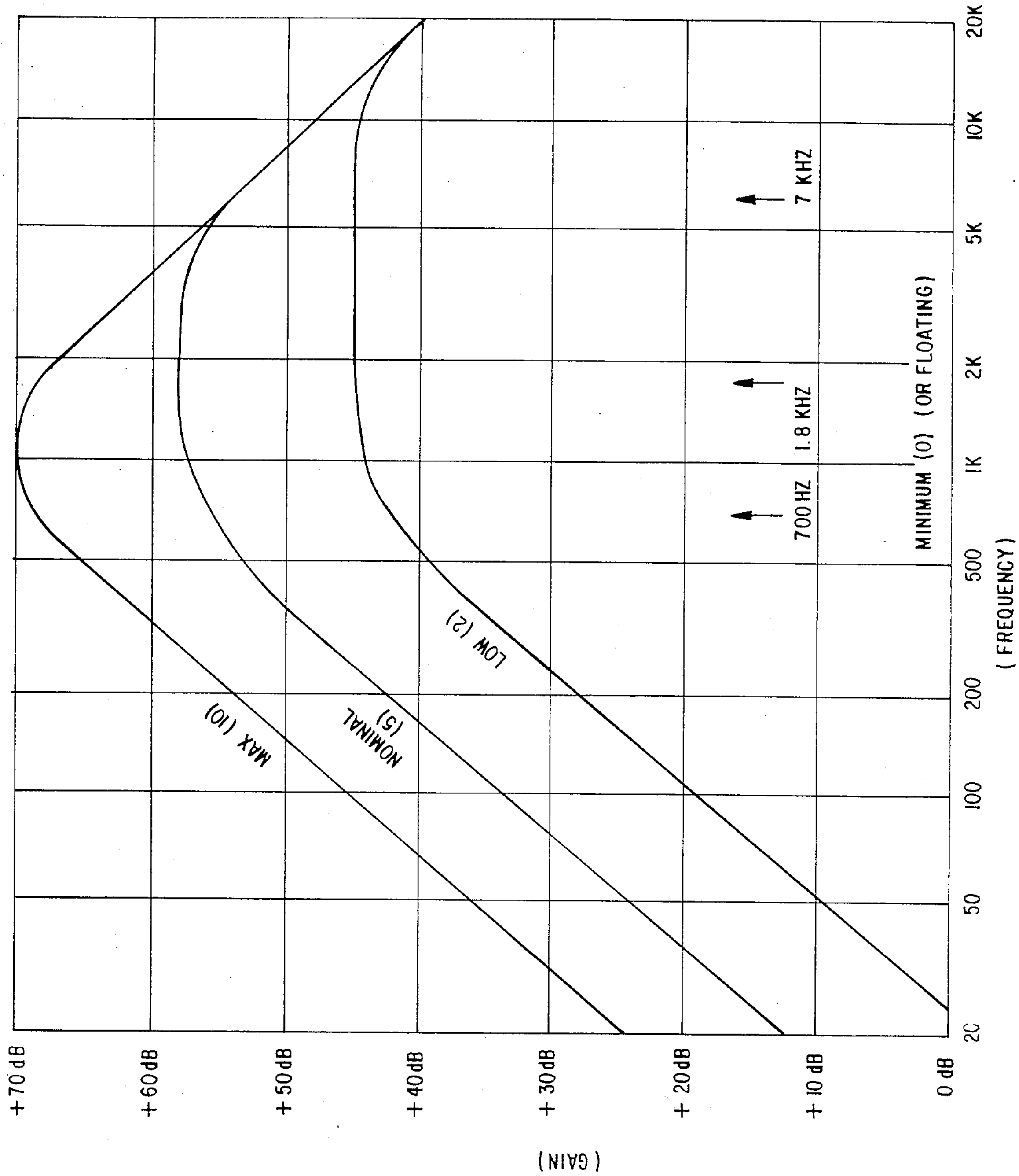
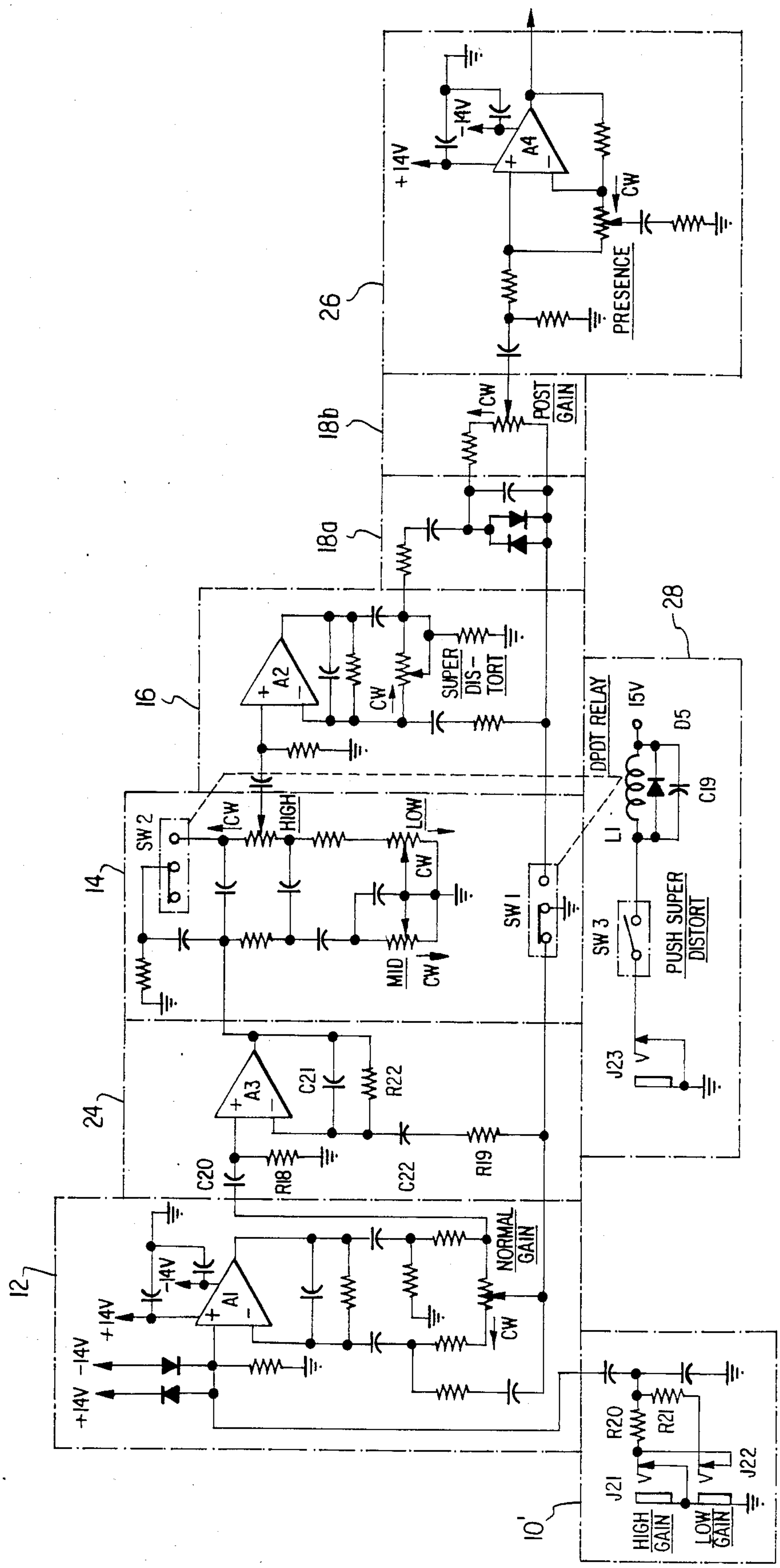

+70 dB
+60 dB
+50 dB
+40 dB
+30 dB
+20 dB
+10 dB
0 dB
(GAIN)
20
50
100
200
500
1K
2K
5K
10K
20K
(FREQUENCY)
MAX (10)
NOMINAL (5)
LOW (2)
MINIMUM (0) (OR FLOATING)
700 HZ
1.8 KHZ
7 KHZ

FIG. 3



26
18b
18a
16
14
24
12
10'
28
+14V
-14V
A4
A3
A2
A1
CW
PRESENCE
POST GAIN
SUPER DIS- TORT
HIGH
LOW
MID
SW 2
SW 1
SW 3
DPDT RELAY
15V
D5
C19
L1
PUSH SUPER DISTORT
J23
C21
R22
C22
R19
C20
R18
NORMAL GAIN
R20
R21
J21
J22
HIGH GAIN
LOW GAIN

FIG. 6

SUPERDISTORTED AMPLIFIER CIRCUITRY WITH NORMAL GAIN

This application is a continuation of application Ser. No. 63,924, filed 06/19/87 and abandoned concurrently with the filing of this application.

BACKGROUND OF THE INVENTION

1. Field of the Invention

This invention is directed to audio amplifier circuitry, and in particular to such circuitry for simulating two-channel audio amplifiers, i.e. audio amplifiers having two alternate channels for signal amplification, but using only a single amplifying channel for providing either selective distortion or non-distorted ("clean" audio output) modes of amplification of audio signals.

2. Background of the Invention

This invention is related to the subject matter disclosed in U.S. Pat. No. 4,405,832, "CIRCUIT FOR DISTORTING AN AUDIO SIGNAL", issued in the name of Jack C. Sondermeyer (a co-inventor of the subject application) on Sept. 20, 1983. The disclosure of that U.S. patent is incorporated herein by reference.

U.S. Pat. No. 4,405,832 discloses distortion generating circuitry including high gain amplification with a variable, controlled feedback network for varying the gain and frequency response thereof. The specific distortion generation is disclosed as a pair of oppositely poled, anti-parallelly-connected diodes, functioning to clip the output signal of an amplifier stage in the single amplifying channel. The feedback network includes a first control means for controlling gain and frequency response of the amplifier associated with the feedback network; and a second control means for mixing the output of the amplifier with the output of the clipping or distortion circuit in variable proportions. The first and second controls are preferably ganged, thereby enabling "tracking" between the amount of gain and the degree of clipping simultaneously with frequency response modification of the amplified output signal to enhance the harmonics of the input signal to the amplifying channel circuitry.

The amplifying circuitry of the aforementioned U.S. patent describes independently operable control switches for switching between normal amplification of audio input signals, whereby a "clean" sound is output, or distortion amplification of the audio input signals, whereby the amplified signals are distorted to enhance the sound of the audio output.

The ganged normal/distortion amplification control of the aforementioned U.S. patent has the disadvantage that it provides relatively broad band gain, as the gain of a pre-amplifier stage is effectively electrically connected constantly regardless of the mode of operation, i.e., normal or distorted output.

SUMMARY OF THE INVENTION

As is evident from the above description, the present invention, while employing the techniques of the distortion control circuitry in U.S. Pat. No. 4,405,832, modifies such techniques at least by the inclusion of additional amplification (gain) prior to the distortion stage where the pre-amplified audio input signal is clipped to generate harmonics. This causes the audio signal in the distortion stage, typically a pair of anti-poled, parallelly-connected diodes, to become operationally "superdistorted", thereby generating additional harmonics

which further enhance the sound of the amplified audio signals, and particularly those audio signals within the frequency range produced by guitars.

Additionally, the present inventors have discovered that the provision of additional amplification prior to clipping or distorting the already amplified audio input signal, produces significant enhancement of the "quality" of the output sound, and particularly with application to the amplification of audio frequencies in the range normally associated with guitars. Reference is made to the discussion of the effects of distortion on such audio frequencies under BACKGROUND OF THE INVENTION in U.S. Pat. No. 4,405,832.

Furthermore, the "superdistortion" circuit of the present invention simulates a two-channel audio amplifier circuit by enabling alternate mode selection between normal gain control to produce a "clean" audio output, and "superdistortion" control with commensurate independent amplification control of the variable gain of a superdistortion amplifier stage preceding the distortion stage, and post distortion gain control, to produce a distorted sound output. Thus, when the distortion mode is selected, the operator has the ability to independently vary the amplification of the audio signal to be distorted and the amplification of the audio signals after they have passed the distortion stage.

Furthermore, switching to the normal gain control mode automatically disables the superdistortion amplifier stage, the distortion stage and the post gain control stage, such that amplification of the audio signal in this mode is controlled solely by the gain of the preamplifier stage.

Another feature of the present invention is the employment of tone control for low, middle and high frequencies in both the non-distorted and distorted modes of operation and which includes a switch ganged with the mode control switch such that the mid-frequency range of the preamplified audio input signals are reintroduced with selection of the "superdistortion" mode of operation.

The present invention also is adaptable for operation with other well-known "presence" amplifier stages, brightening circuits and DPDT relay switches for effecting mode control, for example.

In a modified embodiment of the invention, an intermediate amplifier stage is included after a preamplifier stage to enable decreased amplification (gain) of the preamplifier stage, thereby prohibiting clipping of large amplitude audio input signals in the preamplifier stage.

A primary object of the present invention is to provide single channel audio amplification circuitry capable of operating in at least two modes, one mode providing normal gain (clean sound output) and the other mode providing greatly enhanced distorted sound output with commensurate flexibility in normal gain control, post gain control and "superdistortion" gain control.

A further object of the present invention is to provide significant enhancement of the distorted sound produced by clipping amplified audio sound frequencies, and particular those frequencies associated with guitars.

Yet a further object of the present invention is to provide a widely variable gain adjustment of a distortion stage in audio amplification circuitry in conjunction with an independently variable post distortion gain control.

And yet another object of the present invention is to provide tone control of the low, mid-range and high frequencies of the preamplified audio input signals with automatic re-introduction of the mid-range frequencies when in the "superdistortion" control mode of operation.

Yet another object of the invention is to provide all of the objects, features and advantages of the amplification circuitry as disclosed in U.S. Pat. No. 4,405,832, but with much greater flexibility with respect to adjustment of the normal gain, distortion gain and post gain distortion control and with commensurately greater enhancement of the distorted sound output.

BRIEF DESCRIPTION OF THE DRAWINGS

The above objects, features and advantages of the invention are readily apparent from the following description of preferred embodiments of the best mode of carrying out the invention when taken in conjunction with the following drawings, wherein:

FIG. **1** is an exemplary preferred embodiment of a two-amplifier version of the superdistorted amplifier circuitry of the present invention with tone control and using a double-pole-double-throw (DPDT) switch to effect mode conversion control;,

FIG. *2a* represents an exemplary bridged-T tone control circuit forming a portion of the tone filter control circuit having application with the circuit of FIG. **1**;

FIG. *2b* is a gain vs. frequency plot representing the operation of the bridged-T tone control circuit of FIG. *2a* in both the normal gain control mode and the superdistortion control mode;

FIG. **3** is a gain vs. frequency plot at specified maximum, nominal and minimum settings of the superdistortion stage of FIG. **1**;

FIG. **4** illustrates the distortion stage of the superdistorted amplifier circuitry of FIG. **1**;

FIG. **5** illustrates the post gain stage of the superdistortion amplifier circuitry of FIG. **1**;

FIG. **6** is a modified embodiment of the superdistorted amplifier circuitry of FIG. **1** with the inclusion of an intermediate amplifier stage, an exemplary peripheral circuit, namely a "presence" circuit; and an external footswitch circuit to effect mode conversion control with the embodiments of FIGS. **1** and **2**.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

The exemplary preferred embodiment of the superdistortion control circuitry of the invention uses a preamplifier stage, the output of which is input to a tone control circuit providing low, mid-range and high frequency control with subsequent amplification of the audio signals such that the amplified audio signals will superdistort the distortion stage, thereby affording greatly enhanced distorted audio sound output from the superdistortion control circuitry.

The mode control selection between normal gain and superdistortion is effected by a DPDT switch, although other type switches are employable with the superdistortion control circuitry of the invention as will be more readily apparent from the following description with respect to FIG. **6**. For simplification of the following description, the superdistortion amplifier circuitry of FIG. **1** is divided into the following functional components: (1) input stage **10**, (2) preamplifier stage **12**, (3) tone control filter stage **14**, (4) superdistortion gain stage **16**, (5) distortion stage **18*a*** and (6) output stage **18*b***.

Input stage **10** includes: input jack **J20** for electrical connection with a source of audio signals, such as those produced by a guitar; resistor **R1** and capacitor **C1**, forming a very high frequency low pass filter network; and DC isolation capacitor **C2**.

In preamplifier stage **12**, one terminal of diodes **D1** and **D2** are connected in back-biased relationship to the non-inverting input of amplifier Al to provide protection for the preamplifier stage from high level audio input signals. Resistor **R2** is connected between the non-inverting input of amplifier **A1** and ground. Capacitor **C3** is connected between the cathode of diode **D2** (and the +14 V power supply voltage) and ground to function as a positive by-pass for the +14 V power supply. Capacitor **C3'** provides the same by-pass function for the −14 V power supply voltage.

Amplifier **A1** is a standard broad band operational amplifier of the non-inverting type and powered by plus and minus supply voltages as indicated in FIG. **1**. Throughout the following description it is understood that the power supply voltages (and type, i.e. bi-polar or non-bi-polar) shown are only exemplary, as those skilled in the audio amplification art will readily recognize that other power voltage levels as well as non bi-polar power supply types can be used with the superdistortion circuitry of the invention.

Feedback network **22** provides high frequency and low frequency roll-off to essentially shape the frequency response of non-inverting amplifier **A1** to a desired characteristic for preamplification of the input signals at input jack **J20**. Feedback network **22** includes feedback capacitor **C4** connected from the output of amplifier **A1** to the inverting input thereof, which provides high frequency roll-off for operational amplifier **A1**. Feedback resistor **R3** is parallelly connected across feedback capacitor **C4**. The terminal of resistor **R3** (connected to the inverting input of operational amplifier **A1**) is connected through DC isolation capacitor **C5** to series-connected resistor **R4** and capacitor **C6**, and the other terminal of resistor **R3** is connected through DC isolation capacitor **C18** to the junction node of resistors **R5** and **R6**. Resistor **R7** is connected between the junction node of capacitor **C5** and resistor **R4** and the CW terminal of normal gain potentiometer **P1**. The movable tap of potentiometer **P1** is connected to **C6** and stationary contact **CT1** of mode control switch **SW1**.

The following gain and frequency response is obtained with mode control switch **SW1** switched to have movable contact **CT2** thereof in contact with the movable tap of potentiometer **P1** such that it is short-circuited to ground. The aforementioned position of ganged mode control switch **SW1** affords normal or a clean audio output of the superdistortion amplifier circuit shown in FIG. **1**. The function of ganged mode control switch **SW2** associated with tone control filter circuit **14** is discussed more fully, infra. With the mode control established in the normal mode, as described supra, maximum gain of operational amplifier **A1** is obtained with the movable tap of potentiometer **P1** at the CW position shown in FIG. **1**, and movement of the potentiometer tap CCW causes reduced gain of operational amplifier **A1**.

In the normal gain control mode, it is convenient to consider the gain of operational amplifier **A1** at three settings, namely: (1) minimum, (2) maximum and (3) nominal setting of potentiometer **P1**, where minimum and maximum setting respectively correspond to the

CCW and CW settings of potentiometer **P1** as shown in FIG. **1**; and the nominal setting conforms to the mid-range setting of potentiometer **P1**. With the circuit component values as set forth in Table I, the following gains of preamplifier stage **12** are obtained using the well-known equation for the gain of an operational amplifier:

$$A_v=(R3/R+1)\times(P1/(P1+R6)),$$

where R is the effective resistance from the inverting input of operational amplifier **A1** to ground.

(1) 0 or -infinity dB;
(2) 55.12 or +35 dB; and
(3) 4.57 or +13 dB

The associated frequency response of operational amplifier **A1** is as follows using as a reference the aforementioned nominal setting of normal gain potentiometer **P1**. The high frequency response is determined as:

$$f=1/(6.28\times R\times C)=1/(6.28\times 33K\times 100\times 10^{-12})=48 \text{ kHz } (-3 \text{ dB}),$$

where R is the resistance and C the capacitance having the nominal values shown in Table I.

The low frequency response is determined as:

$f=1/(6.28\times R\times C)=1/(6.28\times 5400\times 2.2\times 10^{-6})$ 15 Hz (-3 db), where R is the effective impedance and C is the capacitance of capacitor **C5** (all component values as indicated in Table I).

Preamplifier stage **12** is also designed to provide high frequency boost (determined as follows with the aforementioned nominal setting of normal gain potentiometer **P1**).

$$A_{V(VHF)}=33K/(5K//\ 3.3K)+1=17.5 \text{ or } +25 \text{ dB}$$

The effective very high frequency boost is then determined as 25 dB−3 dB (the gain computed, supra, with normal gain potentiometer **P1** at its nominal gain setting), or +12 dB.

All of the above gains and frequency roll-off values have been determined from the component values of FIG. **1** as listed in Table I at the end of the specification.

It is a significant feature of the invention that, in the superdistortion mode of operation with ganged switch **SW1** connected as shown in FIG. **1**, the gain of preamplifier stage **12** is essentially "unity" (0 dB), as operational amplifier **A1**, feedback network **22**, and normal gain control potentiometer **P1** are "floating". Therefore, the gain of preamplifier stage **12** in the superdistortion mode of operation is independent of the setting of normal gain potentiometer **P1**, unlike the gain of the preamplifier stage (operational amplifier **10**) as shown in FIG. **1** of U.S. Pat. No. 4,405,832.

The output from preamplifier stage **12** is taken from the node junction between resistor **R6** and normal gain potentiometer **P1** and input to tone control filter stage **14** at the node junction of capacitor **C7**, **C8** and resistor **R8**. Tone control filter stage **14** is illustrated in FIG. **1** as comprising passive low, mid-range, and high frequency filters, respectively. However, those skilled in the audio amplifier art will recognize that active filters could also be used in place of passive filters. Capacitor **C8** and potentiometer **P2** form a high-pass frequency filter; capacitor **C9**, resistor **R10** and potentiometer **P4** form a low pass frequency filter; and capacitors **C10**, **C11** and potentiometer **P3** form a mid-range frequency filter. The low, mid-range and high frequency filters of tone control filter stage **14** have respective center frequencies of 100 Hz, 250 Hz and 5 kHz (with the nominal component values shown in Table I). The design of tone control filter stage **14** is well-known to those skilled in the audio amplifier art. However, the use of such a three-frequency range filter is of particular significance with application of the superdistortion amplifier circuitry of the present invention to the amplification of audio signals produced by guitars.

With continuing reference to FIG. **1**, and with ganged mode control switch **SW2** in the position for "normal" gain mode control, resistor **R11** and capacitor **C7** are not electrically connected in the tone control filter circuit **14**. However, with ganged switches **SW1** and **SW2** switched to the position shown in FIG. **1**, such that the superdistortion amplifier circuit is set in the superdistortion mode, the common node between capacitor **C7** and resistor **R11** is electrically connected to the CW terminal of potentiometer **P2**. This places capacitor **C7** in parallel with capacitor **C8** of hi-pass filter in tone control filter stage **14** to effectively destroy the mid-frequency notch of the tone filters therein, thereby emphasizing the mid-range frequencies with the superdistortion amplifier circuit of FIG. **1** connected in the superdistortion mode of operation (switches **SW1** and **SW2** connected as illustrated therein).

FIG. **2***a* shows the pertinent components of the tone control circuit **14** of FIG. **1**, and FIG. **2***b* illustrates the gain vs. frequency characteristics of the tone control filter **14** for low, mid-range and high frequencies. With the superdistortion amplifier control circuit of FIG. **1** set in the normal gain mode of operation (ganged switches **SW1** and **SW2** set oppositely as shown in FIG. **1**), capacitor **C7** (known as a "thick" capacitor to those skilled in the audio amplifier art) is not connected in parallel with capacitor **C8** and therefore the gain vs. frequency response as shown in FIG. **2***b* is the normal tone response as shown in solid lines. However, with ganged switch **SW2** set in the superdistortion mode, "thick" capacitor **C7** is now electrically connected in tone control filter circuit **14** and the response characteristics thereof are altered as shown by the dotted line in FIG. **2***b*, with a commensurate mid-boost in gain.

From the above description it is evident that the addition of the "thick" capacitor (1) destroys the 250 Hz notch; (2) renders the low and mid-range filters ineffective; and (3) limits or changes the high frequency range control. These characteristics of the frequency response are of particular interest in enhancing the sound output of the superdistortion amplifier circuit of FIG. **1** in the audio frequency range of guitars.

The output of tone control filter stage **14** is electrically connected to the non-inverting input of high gain operational amplifier **A2** in superdistortion amplifier stage **16** through interstage coupling capacitor **C12**. Resistor **R12** provides a normal biasing function for operational amplifier **A2**. Feedback resistor **R13** is connected in parallel across feedback capacitor **C13**. This parallel combination of resistor **R13** and capacitor **C13** is connected between the output and the inverted input of operational amplifier **A2**. One terminal of superdistortion gain control potentiometer **P5** is connected to the common terminal of resistor **R13** and DC isolation capacitor **C14**. The other terminal of superdistortion gain control potentiometer **P5** is connected to the movable tap thereof, which in turn is connected to ground through bleeder resistor **R14**. Capacitor **C15**, connected between the respective terminals of resistor R13 and the CW terminal of superdistortion control potentiometer P5, provides DC isolation.

Resistor R15 is serially connected to capacitor C14 and to ground in the superdistortion control mode with SW1 connected as shown in FIG. 1. In the normal gain control mode with switches SW1 and SW2 switched to their respective opposite terminals as shown in FIG. 1, resistor R15 is "floating", the effect of which will be more fully described, infra.

The gain of operational amplifier A2 in superdistortion amplifier stage 16 is determined by the equation:

$$Av=R_{(eff)}/R15+1,$$

where $R_{(eff)}$ is defined as the parallel combination of R13 and P5.

As with the preamplifier stage 12 previously described, it is again convenient for purposes of defining the operational gain characteristics of superdistortion stage 16 to select (1) minimum, (2) maximum and (3) nominal settings of potentiometer P5 (with potentiometer P5 set full CCW defining minimum setting, and set full CW defining maximum setting as shown in FIG. 1, and set at the mid-point for defining the nominal setting).

With the above settings of superdistortion potentiometer P5 the following gains are obtained with the component values as set forth in Table I:

Minimum setting: Gain=1 (0 dB)
Maximum setting: Gain=3200 (approx. +70 dB)
Nominal setting: Gain=830 (+58 dB)

The inventors consider that approximately 60 dB gain is necessary to achieve the phenomena of "sustain" whereby acoustic feedback is achieved between the guitar and the guitar amplifier/speaker in the audio frequency range normally associated with guitar music.

Thus, movement of the movable tap of superdistortion gain control potentiometer P5 throughout its range of effectiveness (from full CCW to CW rotation), results in a gain change of 1 to 3200, with the superdistortion control amplifier of FIG. 1 set to the superdistortion mode of operation.

The commensurate frequency response of superdistortion stage 16 of FIG. 1 is as follows (using the nominal and maximum potentiometer settings as described, supra):

$$f_{(high)nom}=1/(6.28\times R\times C)=1/(6.28\times 83K\times 270\times 10^{-12})=7\text{ kHz }(-3\text{ dB}).$$

$$f_{(high)max}=1/(6.28R\times C)=1/(6.28\times 320K\times 270\times 10^{-12})=1.8\text{ kHz }(-3\text{ dB}).$$

$$f_{(low)}=1/(6.28\times R\times C)=1/(6.28\times 100\times 2.2\times 10)=700\text{ Hz }(-3\text{ db}).$$

The above computations are made using the component values set forth in Table I.

FIG. 3 illustrates the gain vs. frequency characteristics of superdistortion stage 16 with the representative component values shown in Table I, for maximum, nominal, low and minimum settings of superdistortion potentiometer P5.

However, in contradistinction to the aforementioned gain and frequency characteristics obtained in the superdistortion mode, the gain is "unity" (0 dB) with the superdistortion control circuit of FIG. 1 set to operate in the normal gain control mode as established by the setting of switches SW1 and SW2 previously described. That is, the gain remains "unity" (0 dB) regardless of the setting of superdistortion potentiometer P5 and there is no alteration of the frequency response of superdistortion amplifier stage 16 in the normal gain mode of operation. This operation is obtained because superdistortion amplifier stage 16 is "floating" and the gain thereof is "unity" (0 dB) regardless of the setting of superdistortion potentiometer P5.

The gain of superdistortion amplifier stage 16 is designed to be high so that the output of operational amplifier A2 in combination with the distortion stage 18a is clipped, thereby generating a number of harmonics of the frequencies of the input signals thereto from the output of tone control filter stage 14. The presence of such harmonics is believed by the inventors to enhance the sound output from the amplifier circuitry in the superdistortion control mode, and especially those harmonics associated with the mid-range frequencies. In particular, guitars are known to produce predominantly low frequency notes, and the harmonics generated by the amplification of the superdistortion amplifier stage 16 and distortion stage 18 (described, infra.) are believed to enhance the sound of the output signal generated by the superdistortion amplifier circuitry as shown and described in FIG. 1.

With continuing reference to FIG. 1 and additional reference to FIG. 4, the output signal from superdistortion amplifier stage 16 is input through series-connected resistor R16 and capacitor C16 in distortion stage 18a, and as shown in the preferred embodiment of FIG. 1, to anti-pole, parallelly-connected clipping diodes D3 and D4. Capacitor C17 is connected in parallel across clipping diodes D3 and D4, to provide high frequency roll-off as discussed more fully, infra.

In the normal gain control mode, superdistortion stage 18a is "floating" (as is superdistortion amplifier stage 16), and therefore clipping diodes D3 and D4 are ineffective such that no clipping of the output of superdistortion stage 18 occurs by diodes D3 and D4, and capacitor C17 is not electrically connected so that the frequency response of the output signal from superdistortion stage 16 is not modified by superdistortion clipping stage 18a.

In the superdistortion mode of operation, with switch SW1 at contact CT2, superdistortion clipping stage 18a is operative such that the amplified audio signal output from superdistortion amplifier stage 16 is clipped at approximately 1.2 volts (peak-to-peak) by diodes D3 and D4. With the component values shown in Table I, including the gain of operational amplifier A2 in superdistortion stage 16, the amplitude of the amplified output signal from superdistortion amplifier stage 16 is approximately 28 volts (peak-to-peak).

Further enhancement of the audio signals, especially in the case where the audio input to the superdistortion amplifier circuit of FIG. 1 is produced by guitars, is provided by the high-frequency roll-off produced by the series-connected resistor R16 and capacitor C17 at a frequency of 5 kHz as determined by the following well-known equation:

$$f=1/(6.28\times R16\times C17)=1/(6.28\times K\times 0.033\times 10^{-6})=5\text{ kHz}$$

As mentioned above, this enhancement of the audio signal takes place only with the superdistortion circuit of FIG. 1 connected in the superdistortion mode of operation. The nominal component values of R16 and C17 are as shown in Table I.

With continuing reference to FIG. 1, coupling capacitor C16 in distortion stage 18*a* provides DC isolation.

The output of the superdistortion amplifier control circuitry shown in FIG. 1 is obtained from the audio signal output of superdistortion clipping stage 18*a* (the audio signal across capacitor C17) and input to post gain stage 18*b* (also shown in detail in FIG. 5). Post gain stage 18*b* consists of resistor R17 connected to capacitor C17 and master volume control potentiometer P6. The maximum gain setting of master volume control potentiometer P6 is at the CW position indicated in FIG. 1, the minimum gain setting of potentiometer P6 is at the opposite, or CCW setting, and the nominal gain setting is with potentiometer P6 set to a mid-position (as are the nominal gain settings of potentiometers P1 and P5, as discussed, supra.). These settings of post gain control potentiometer P6 produce the following gains (with the superdistortion amplifier circuitry of FIG. 1 set to operate in the superdistortion mode):

$A_{max}=\frac{1}{2}$ (−6 dB);

$A_{min}=0$ (−infinity dB); and

$A_{nom}=\frac{1}{4}$ (−12 dB)

In the normal gain control mode of operation of the superdistortion amplifier control circuitry of FIG. 1, the post gain control stage 18*b* is "floating" and therefore the gain of that stage is "unity" (0 db) regardless of the setting of master volume control potentiometer P6 (in conjunction with the unity gain operation of superdistortion clipping circuit 18*a* and superdistortion amplifier stage 16, as described, supra.)

It is evident to those skilled in the audio amplifier art that the aforementioned operation of post gain control stage 18*b* affords additional flexibility in controlling the amplification of audio signals in that the gain of post gain control stage 18*b* is effective only in the superdistortion control mode of operation and ineffective in the normal gain mode.

The output from post gain control potentiometer P6 is input to a power amplifier (not shown) to provide the audio output.

In the modified embodiment of the superdistortion control amplifier circuit shown in FIG. 6, the preamplifier stage 12, tone control filter stage 14, superdistortion amplifier stage 16, distortion stage 18*a* and post gain control stage 18*b* are identical to that of FIG. 1. However, the superdistortion control amplifier of FIG. 6 has been modified to include the following: (1) a dual high/low gain input circuit; (2) an intermediate amplifier stage 24; (3) a "presence" circuit 26 connected to the output from post distortion gain potentiometer P5; and (4) an external footswitch circuit 28, which is substituted for the push switches SW1 and SW2.

Dual high/low gain input circuit 10′ (consisting of input jacks J21 and J22 and resistors R20 and R21) reduces the gain of an input signal by ½ (with resistors R20 and R21 of equal value) when the audio input signal is connected with the low gain portion of the input jack. This feature is desirable for high level audio input signals to avoid saturation of operational amplifier A1 in preamplifier stage 12.

Continuing with the circuitry shown in FIG. 6, intermediate amplifier stage 24 connected between the output of preamplifier stage 12 and the input of the tone control filter stage 14 provides amplification of the output of preamplifier stage 12. The output of preamplifier stage 12, obtained from the normal gain potentiometer P1, is input through capacitor C20 to the non-inverting input of operational amplifier A3. Resistor R18 provides normal biasing of operational amplifier A3. A feedback network consisting of parallelly-connected resistor R22 and C21 is connected between the output and inverting input of the operational amplifier as shown in FIG. 6. Capacitor C22 is connected to the common node of resistor R22 and capacitor C21, and capacitor C22 is serially connected to the common node of the movable tap of normal gain potentiometer P1 and switch SW1 of the relay through resistor R19. The gain Av of intermediate amplifier stage 24 is determined as:

$$Av=1+R22/R19=4.3$$

with the component values as indicated in Table I.

This gain is obtained only in the normal gain control mode of operation (ganged mode control switches SW1 and SW2 set as shown in FIG. 6). In the superdistortion control mode of operation, with the ganged switches SW1 and SW2 set opposite to that illustrated in FIG. 6, intermediate amplifier stage 24 is "floating" and has a gain of "unity" (0 dB). This independent operation of intermediate amplifier stage 24 provides flexibility in controlling the gain of the audio frequency signals input to the superdistortion amplifier control circuitry of FIG. 6, consistent with the operation of the superdistortion control circuitry previously described with respect to FIG. 1.

The modified superdistorted audio amplifier circuit of FIG. 6 shows "presence" amplifier circuit 26 connected to the output of the superdistorted audio amplifier circuit of FIG. 1, namely the output of post distortion gain control potentiometer P6. Presence circuit 26 is known to the art and is illustrated in FIG. 6 to demonstrate the use of "peripheral" circuitry with the superdistorted audio amplifier control circuitry of the present invention, and therefore no further description of its structure and operation is necessary for the purposes of this invention.

FIG. 6 illustrates an external footswitch circuit 28 adapted to be used with the superdistortion audio amplifier control circuitry of the present invention in a manner similar to that of DPDT switches SW1 and SW2 of FIG. 1. Additionally, footswitch circuit 28 also includes diode D5 and parallelly-connected capacitor C19, which in turn are parallelly connected across the coil L1 of the DPDT relay associated with footswitch circuit 28. Diode D5 and capacitor C19 prevent "flyback" and "ringing" upon activation and deactivation of superdistortion mode control switch 28 by depression of a remote shorting switch connected to footswitch jack J23, as is known to those skilled in the audio amplifier art.

Table I lists the various components and their respective associated exemplary component values of the embodiments of the superdistorted audio ampifier control circuit described herein.

The preceding specification describes exemplary preferred embodiments of the best mode of carrying out the invention, and is therefore not intended to represent limitations of the scope of the invention. It is understood that equivalent variations and modifications of the invention will be apparent to those skilled in the audio amplification art. For example, the distortion stage 18*a* could consist of multiple series/parallel-connected diodes (germanium or silicon), zener diodes, or other clipping means. Also, the switches SW1 and SW2 in FIG. 1 could consist of two individual switches instead of ganged switches. Such variations, modifications and equivalents are within the scope of the invention as set forth in the claims appended hereto, with such claims interpreted to obtain benefit of all of the equivalents to which the invention is entitled.

TABLE I

COMPONENT	VALUE	COMPONENT	VALUE
Resistors		Capacitors	
R1	1 kohm	C1	100 pF
R2	220 kohms	C2	.1 uF
R3	33 kohms	C3	.1 uF
R4	3.3 kohms	C3	.1 uF
R5	22 kohms	C4	100 pF
R6	2.7 kohms	C5	2.2 uF
R7	470 ohms	C6	.033 uF
R8	47 kohms	C7	.015 uF
R10	10 kohms	C8	270 pF
R11	220 kohms	C9	.047 uF
R12	470 kohms	C10	.1 uF
R13	470 kohms	C11	.015 uF
R14	22 kohms	C12	.1 uF
R15	100 ohms	C13	270 pF
R16	1 kohm	C14	2.2 uF
R17	10 kohms	C15	2.2 uF
R18	47 kohms	C16	2.2 uF
R19	10 kohms	C17	.033 uF
R20	22 kohms	C18	2.2 uF
R21	22 kohms	C19	2.2 uF
R22	33 kohms	C20	.1 uF
		C21	100 pF
		C22	2.2 uF
Potentiometers			
P1	10 K (linear)		
P2	250 K (linear)		
P3	50 K (linear)		
P4	250 K (audio)		
P5	1 M (audio)		
P6	10 K (linear)		

What is claimed is:

1. Circuitry for amplifying audio signals, comprising:

preamplifier means for amplifying the audio signals and including a first operational amplifier for receiving the audio signals at a non-inverting input thereof and providing a first amplified signal output, first feedback means interconnected between the output of said operational amplifier and an inverting input thereof for variably controlling the gain and frequency response of said operational amplifier within a predetermined range of audio frequencies;

amplifier means responsive to said first amplified signal output and including a second operational amplifier for providing a second amplified signal output and including a second feedback means for selectively controlling the gain and frequency response of said second operational amplifier within a second predetermined range of audio frequencies;

distortion means for selectively distorting said second amplified signal output to provide at least one of a distorted audio output signal and a non-distorted audio output signal;

post gain control means for independently amplifying said distorted audio output signal to provide a third amplified signal output; and

switching means for selectively controlling said preamplifier means, amplifier means and said distortion means to provide said distorted audio output signal with said switching means at one position, and said non-distorted audio output signal with said switching means at another position.

2. Circuitry according to claim 1 wherein said first feedback means includes first circuit means for varying the gain of said first operational amplifier.

3. Circuitry according to claim 2 wherein said first circuit means includes a variable potentiometer electrically connected to said switching means.

4. Circuitry according to claim 1 further comprising tone control filter means for filtering said first amplified signal output in at least one of several frequency ranges to provide a filtered output signal; and wherein said switching means includes first and second switching sections each having respective one and another operating positions with said second switching section being electrically connected in said tone control filter means with said switching means set to said one position, said second switching section modifies said filtered output signal with said switching means set to said another position, and said second switching section does not modify said filtered output signal with said switching means set to said one position.

5. Circuitry according to claim 4 wherein said first and second switching sections are ganged.

6. Circuitry according to claim 4 wherein said tone control filter means further includes at least first, second and third filter sections respectively operating in different frequency ranges of said audio signals, and a second potentiometer electrically connected in one of said at least first, second and third filter sections for selectively varying the gain of said filtered output signal.

7. Circuitry according to claim 4 wherein said first and second switching sections are ganged.

8. Circuitry according to claim 4 wherein said switching means is an external footswitch for operating said first and second switching sections.

9. Circuitry according to claim 4 further comprising intermediate amplifier means responsive to first amplified signal output and providing an intermediate amplified signal output to said tone control filter means.

10. Circuitry according to claim 2 wherein, with said switching means at said one position, said first circuit means is electrically disconnected from said first operational amplifier, and said amplifier means, distortion means and post gain control means are enabled to provide said distorted audio output signal.

11. Circuitry according to claim 10 wherein said second feedback means includes second circuit means for varying the gain and frequency response of said second operational amplifier at respective predetermined low and high frequencies to vary said second amplified signal output.

12. Circuitry according to claim 11 wherein said second circuit means is a selectively variable potentiometer for providing a broad range of selective gain control of said second amplified signal output.

13. Circuitry according to claim 11 further comprising tone control filter means for providing a filtered audio output signal and including second switching means operable with said switching means such that with said switching means at said one position, said filtered audio output signal is not affected.

14. Circuitry according to claim 13 said tone control filter means further includes at least first, second and third filter sections respectively operating in different frequency ranges of said audio signals, and at least one

of said first, second and third filter sections includes a "thick" capacitor connected to said second switching means such that with said first switching means at said one position said "thick" capacitor is connected in one of said first, second and third filter sections, and with said first switching means at said another position said "thick" capacitor is not connected in one of said first, second and third filter sections.

15. Circuitry according to claim **14** wherein said tone control filter means includes passive filter circuitry.

16. Circuitry according to claim **1** wherein said distortion means includes clipping means responsive to said second signal output for providing said distorted audio output signal.

17. Circuitry according to claim **16** wherein said clipping means is operable when said switching means is at said one position.

18. Circuitry according to claim **1** wherein said switching means is a DPDT push button switch.

19. Circuitry according to claim **1** wherein said switching means is an external footswitch.

20. Circuitry according to claim **1** further comprising input jack means connected to the non-inverting input of said first operational amplifier for receiving an audio input signal and including means for providing at least one reduced amplified signal level output.

21. Circuitry according to claim **1** wherein said post gain control means is a variable potentiometer.

22. Circuitry according to claim **21** wherein said post gain control means is operable with said switching means at said one position.

23. Circuitry for selectively amplifying audio signals in distorted and non-distorted modes, comprising:

a series connection of variable gain first amplifier means, variable gain second amplifier means and clipping means; and

switching means for selectively controlling the first amplifier means, the second amplifier means and the clipping means so that the second amplifier means provides substantially unity gain irrespective of any gain to which it is adjusted while the first amplifier means provides that gain to which it is adjusted and the clipping means is disabled in one position of the switching means to produce a non-distorted audio signal output, and in another position of the switching means first amplifier means has unity gain, the second amplifier means has variable gain, and the clipping means is enabled to produce a distorted audio signal output.

24. Circuitry as defined in claim **23** tone control filter means connected between the first amplifier means and the second amplifier means for filtering low- mid- and high-range frequencies at the one position of the switching means and for emphasizing frequencies between the low- and high-range frequencies at the another position of the switching means.

25. Circuitry as defined in claim **24** including post gain control means for selectively controlling gain of the circuitry at said another position of the switching means and which is disabled at the one position of the switching means.

26. Circuitry as defined in claim **25** wherein the gain of the post gain control means is variable between about 0 dB and −12 dB.

27. Circuitry for selectively amplifying audio signals in distorted and non-distorted modes, comprising:

low gain amplifier means having an input for receiving the audio signals and an output for providing a first amplified signal output;

first feedback means interconnected between the output and input of the low gain amplifier means for selectively controlling the gain and frequency response of the low gain amplifier means within a predetermined range of audio frequencies;

high gain amplifier means having an input in series connection with the first amplified output signal and an output for providing a second amplified output signal;

second feedback means interconnected between the output and the input of the high gain amplifier means for selectively controlling the gain and frequency response of the high gain amplifier means within a second predetermined range of audio frequencies;

distortion means having an input connected to the output of the high gain amplifier means for providing a distorted audio output signal when the high gain amplifier means is adjusted through its feedback means to provide a second amplified output signal of high gain; and

switching means for selectively controlling the low gain amplifier means, the high gain amplifier means and the distortion means so that the high gain amplifier means provides substantially unity gain irrespective of any setting of its feedback means while the low gain amplifier means provides that gain to which its feedback means is set and the distortion means is disabled in one position of the switching means, and in another position of the switching means the low gain amplifier means has unity gain, the high gain amplifier means has variable gain, and the distortion means is enabled.

28. Circuitry as defined in claim **27** including tone control filter means between the low gain amplifier means and the high gain amplifier means for filtering low- mid- and high-range frequencies at the one position of the switching means and for emphasizing frequencies between the low- and high-range frequencies at the another position of the switching means.

29. Circuitry for selectively amplifying audio signals in distorted and non-distorted modes, comprising:

a series connection of variable gain first amplifier means, tone control means, variable gain second amplifier means and clipping means; and

switching means for selectively controlling the first amplifier means, the second amplifier means and the clipping means so that the second amplifier means provides substantially unity gain irrespective of any setting of its variable gain while the first amplifier means provides that gain to which it is adjusted, the tone control means gives a non-distortion frequency response and the clipping means is disabled in one position of the switching means to produce a non-distorted audio signal output, and in another position of the switching means, the first amplifier means has unity gain, the second amplifier means has variable gain, the tone control means has a distortion frequency response, and the clipping means is enabled to produce a distorted audio signal output.

30. Circuitry as defined in claim **29** the tone control means for filtering low- mid- and high-range frequencies at the one position of the switching means and for emphasizing frequencies between the low- and high-range frequencies at the another position of the switching means.

31. Circuitry as defined in claim 30 wherein the tone control means includes a "thick" capacitor switched into circuit at the another position of the switching means and switched out of circuit at the one position of the switching means.

32. Circuitry for amplifying audio signals, comprising:

first amplifier means having an input for receiving the audio signals and an output for providing a first amplified signal output, first feedback means interconnected between the output and input of the first amplifier means for selectively controlling the gain and frequency response of said first amplifier means within a predetermined range of audio frequencies;

second amplifier means serially connected with the first amplifier means having an input responsive to the first amplified output signal and an output for providing a second amplified output signal and including a second feedback means for selectively controlling the gain and frequency response of said second amplifier means within a second predetermined range of audio frequencies;

distortion means for selectively distorting said second amplified signal output to provide at least one of a distorted audio output signal and a non-distorted audio output signal dependent upon unity gain and non-unity gain of the preamplifier means;

post gain control means for independently amplifying the output of the distortion means; and

switching means for selectively controlling said preamplifier means so that it is "floating" with unity gain in one position of the switching means and has a non-unity, low gain as set by the first feedback means at another position of the switching means, for selectively controlling the amplifier means so that it is "floating" with unity gain in the another position of the switching means and has a non-unity, high gain as set by the second feedback means when the switching means is in the one position thereof, so that said distortion means provides said distorted audio output signal with said switching means at the one position, and said non-distorted audio output signal with said switching means at the another position.

33. Circuitry as defined in claim 32 including tone control filter means for filtering said first amplified output signal in high, mid and low frequency ranges and including a "thick" capacitor altering high, mid and low range filtering, the switching means connecting and disconnecting the "thick" capacitor in circuit in response to the one and another positions of the switching means respectively.

34. Circuitry for selectively amplifying audio signals in superdistorted and clean, non-distorted audio output modes, comprising:

a series connection of variable gain first amplifier means for producing the clean, non-distorted audio output mode from the circuitry, tone control filter means for selectively providing frequency filtering of a first kind and of a second kind, variable gain superdistortion amplifier means for providing a high gain output, clipping means for clipping the high gain output to provide the superdistorted audio output mode from the circuitry, and post gain control means for selectively varying the gain of the superdistorted audio output mode; and

switching means for selectively controlling the first amplifier means, the superdistortion amplifier means, the clipping means and the post gain control means so that the superdistortion amplifier means provides substantially unity gain irrespective of any setting of its variable gain while the first amplifier means provides that gain to which it is adjusted, the tone control filter means is adjusted to frequency filtering of the first kind and the clipping means and post gain control means are disabled at one position of the switching means to produce the clean, non-distorted audio output mode from the circuitry, and the first amplifier means provides substantially unity gain irrespective of any setting of its variable gain while the superdistortion amplifier means provides that gain to which it is adjusted, the tone control filter means is adjusted to frequency filtering of the second kind, and the clipping means and the post gain control means are enabled at another position of the switching means to produce the superdistorted audio output mode from the circuitry.

35. Circuitry as defined in claim 34 wherein the frequency filtering of the first kind comprises low- mid- and high-range frequency filtering with a notch at a predetermined low frequency and the frequency filtering of the second kind comprises low- to high-range frequency filtering which eliminates the notch.

36. Circuitry as defined in claim 35 wherein the gain of the post gain control means is variable between about 0 dB and −12 dB.

37. Circuitry as defined in claim 34 wherein the first amplifier means is variable in gain up to about 35 dB and the superdistortion amplifier means is variable in gain up to about 70 dB.

38. Circuitry as defined in claim 35 wherein the first amplifier means is variable in gain up to about 35 dB and the superdistortion amplifier means is variable in gain up to about 70 dB.

39. Circuitry as defined in claim 36 wherein the first amplifier means is variable in gain up to about 35 dB and the superdistortion amplifier means is variable in gain up to about 70 dB.

* * * * *

United States Patent [19]

Scholz

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[54] **ELECTRONIC AUDIO SIGNAL PROCESSOR**

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[21] Appl. No.: **420,280**

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[51] **Int. Cl.**[4] **H03G 3/00**

[52] **U.S. Cl.** **381/61;** 381/98

[58] **Field of Search** .... 381/61, 98, 101, 102, 381/103, 104, 106, 118; 84/DIG. 9; 333/14, 17 L

[56] **References Cited**

U.S. PATENT DOCUMENTS

2,037,285	4/1936	Tasker	381/103
3,304,369	2/1967	Dreyfus	381/106 X
4,146,745	3/1979	Froeschle	381/103 X
4,180,707	12/1979	Moog	381/61
4,211,893	7/1980	Smith	381/118 X
4,363,006	12/1982	Ishigaki et al.	381/106 X
4,363,007	12/1982	Haramoto et al.	381/106 X
4,368,435	1/1983	Bloy	381/98 X
4,400,583	8/1983	Bloy	381/98
4,412,100	10/1983	Orban	381/106 X

OTHER PUBLICATIONS

H. Tremaine, *Audio Cyclopedia,* H. Sams Co., 1969; pp. 937–939.

Electronics Today International, vol. 8, No. 10, Oct. 1979, pp. 47–49; "Audio Compressor".

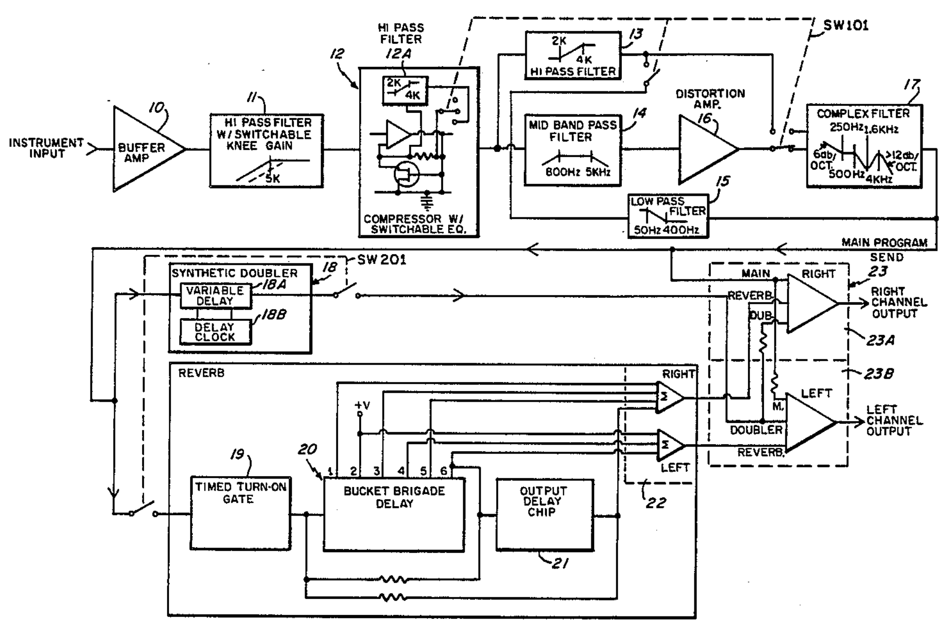
Primary Examiner—Keith E. George

Attorney, Agent, or Firm—Wolf, Greenfield & Sacks

[57] **ABSTRACT**

An electronic audio signal processor especially suitable for electrical instruments such as electric guitars is provided including a controlled distortion and tone alteration portion and a reverb portion. The controlled distortion and tone alternation portion in one form comprises in cascade a compression stage which compresses the amplitude level of an inputted audio signal, a mid band pass filter, a distortion amplifier for adding controlled distortion to said signal and a complex filter having a roll-off of increased attenuation with increased frequency range in the lower and upper audio frequency ranges, and a generally flat response in the middle audio frequency range except with a dip followed by a peak in the upper portion of the mid audio frequency range. The reverb circuit includes a synthetic doubler which provides an output cyclicly varying in pitch from its input and a stereo analog shift register reverb device having two summers which combine staggered adjacent output lines from an analog shift register in different combinations. Two output mixers in conjunction with a switch provide reverb alone, doubling alone or reverb with doubling.

21 Claims, 5 Drawing Figures



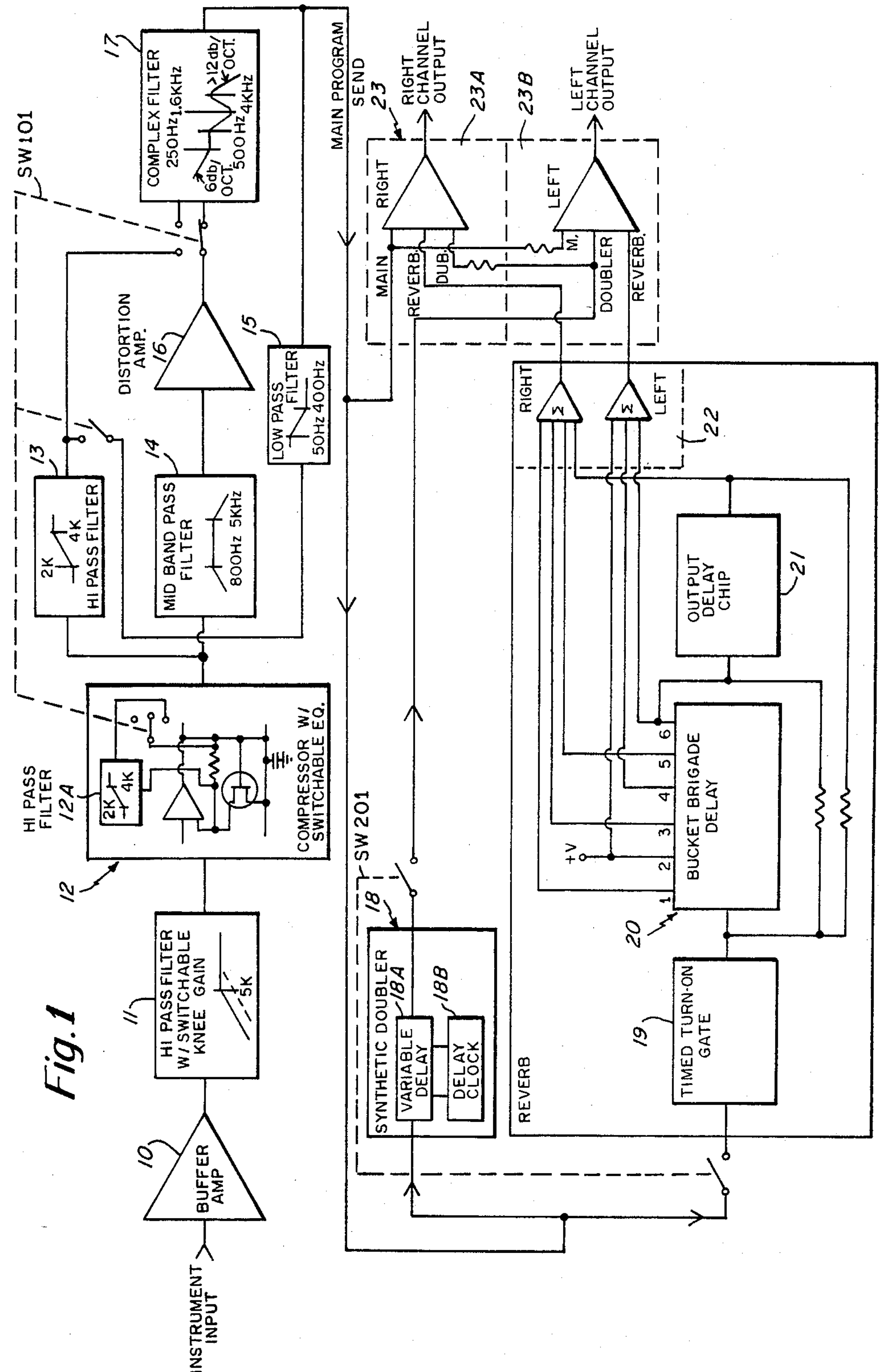
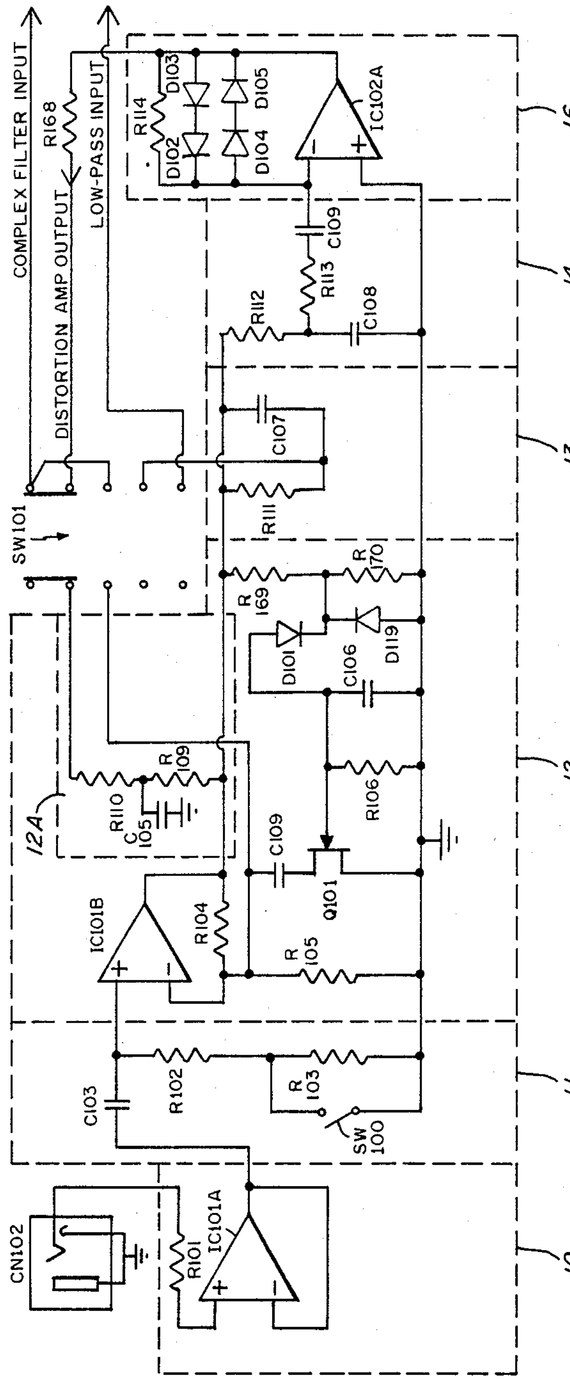

INSTRUMENT INPUT
BUFFER AMP
10
HI PASS FILTER W/SWITCHABLE KNEE GAIN
5K
11
HI PASS FILTER
12A
2K
4K
COMPRESSOR W/ SWITCHABLE EQ.
12
SW101
HI PASS FILTER
13
MID BAND PASS FILTER
800Hz 5KHz
14
LOW PASS FILTER
50Hz 400Hz
15
DISTORTION AMP.
16
COMPLEX FILTER
250Hz 1.6KHz
6db/OCT.
500Hz 4KHz
12db/OCT.
17
MAIN PROGRAM SEND
23
RIGHT CHANNEL OUTPUT
23A
23B
LEFT CHANNEL OUTPUT
RIGHT
LEFT
MAIN
REVERB.
DUB.
M.
DOUBLER
REVERB.
SW 201
18
SYNTHETIC DOUBLER
18A
18B
VARIABLE DELAY
DELAY CLOCK
REVERB
TIMED TURN-ON GATE
19
20
BUCKET BRIGADE DELAY
+V
OUTPUT DELAY CHIP
21
22

Fig. 1

Fig. 2


COMPLEX FILTER INPUT
DISTORTION AMP OUTPUT
LOW-PASS INPUT
R168
R114
D102
D103
D104
D105
IC102A
C109
R113
R112
C108
C107
R111
SW101
R 169
R 170
D101
D119
C106
R 109
R110
C 105
12A
R106
C109
Q101
IC101B
R104
R 105
C103
R102
R 103
SW 100
CN102
R101
IC101A
16
14
13
12
11
10

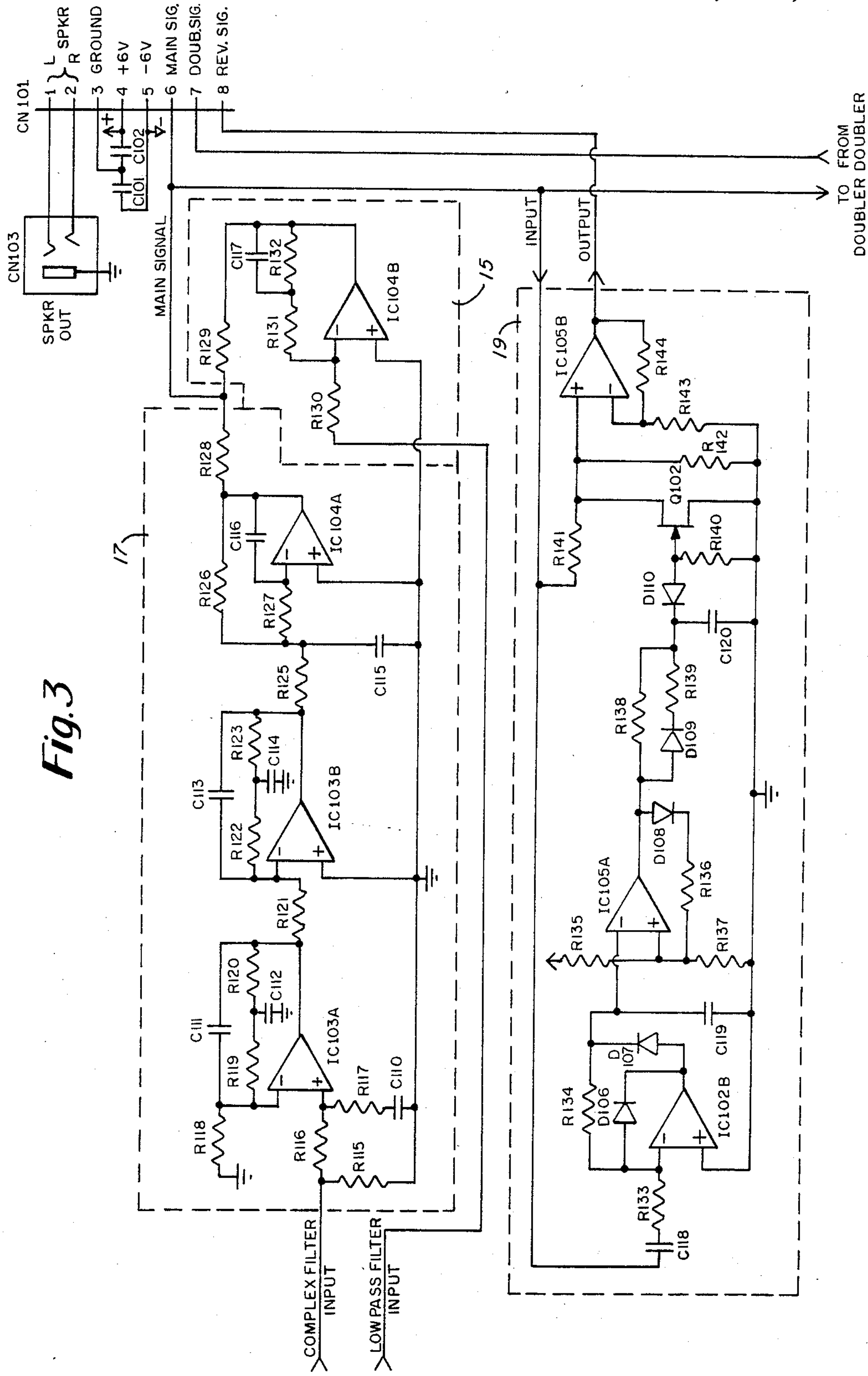

CN101
1 L SPKR
2 R
3 GROUND
4 +6V
5 -6V
6 MAIN SIG.
7 DOUB.SIG.
8 REV. SIG.
CN103
SPKR OUT
MAIN SIGNAL
TO DOUBLER
FROM DOUBLER
INPUT
OUTPUT
COMPLEX FILTER INPUT
LOWPASS FILTER INPUT
IC103A
IC103B
IC104A
IC104B
IC105A
IC105B
IC102B
Q102
15
17
19

Fig.3

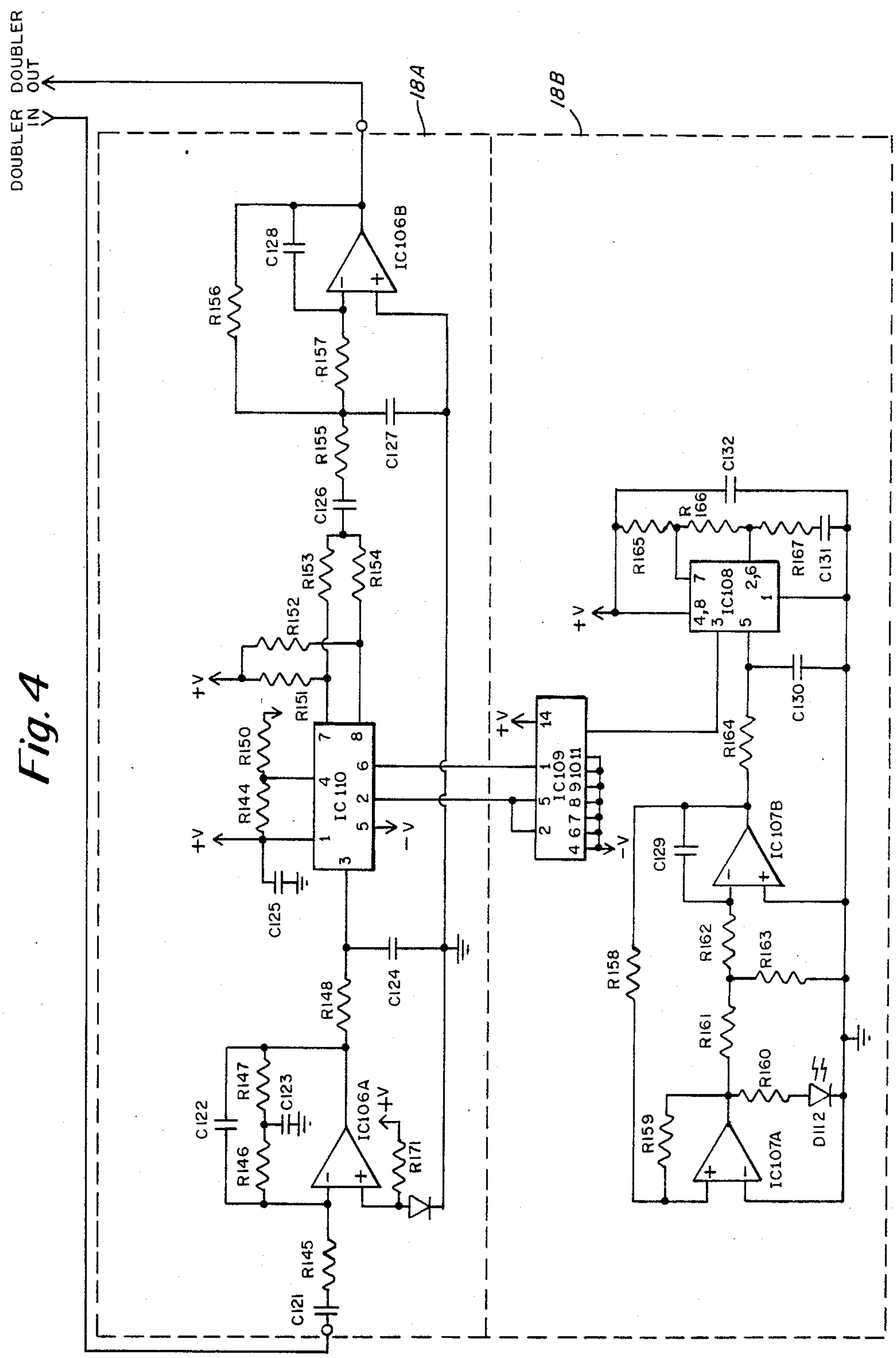

DOUBLER IN
DOUBLER OUT
18A
18B
C128
IC106B
R156
R157
R155
C127
C126
R153
R154
R152
R151
+V
R150
R144
IC110
C125
R148
C124
C122
R147
C123
R146
IC106A
R171
R145
C121
IC109
-V
C132
R 166
R165
IC108
R167
C131
C130
R164
C129
IC107B
R158
R162
R163
R161
R160
R159
IC107A
D112

Fig. 4

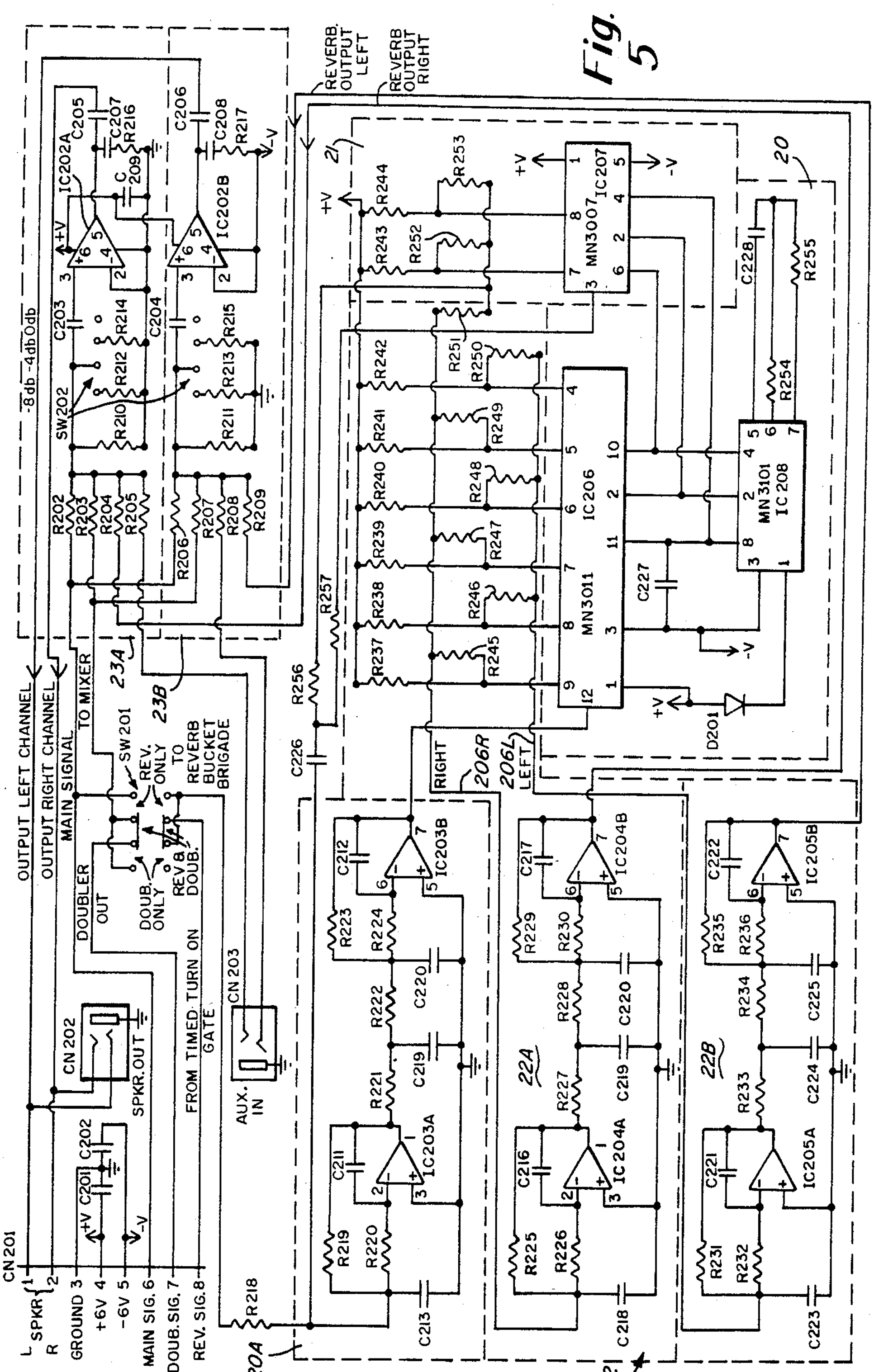

OUTPUT LEFT CHANNEL
OUTPUT RIGHT CHANNEL
MAIN SIGNAL
TO MIXER
DOUBLER OUT
SW201
REV. ONLY
DOUB. ONLY
REV & DOUB.
TO REVERB BUCKET BRIGADE
FROM TIMED TURN ON GATE
CN201
SPKR L R
GROUND 3
+6V 4
-6V 5
MAIN SIG. 6
DOUB. SIG. 7
REV. SIG. 8
CN202
SPKR. OUT
CN203
AUX. IN
SW202
-8db -4db 0db
IC202A
IC202B
REVERB OUTPUT LEFT
REVERB OUTPUT RIGHT
MN3007 IC207
MN3011 IC206
MN3101 IC208
D201
IC203A
IC203B
IC204A
IC204B
IC205A
IC205B
RIGHT
LEFT
20
20A
21
22
22A
22B
23A
23B
206R
206L

Fig. 5

ELECTRONIC AUDIO SIGNAL PROCESSOR

TECHNICAL FIELD

This invention is directed to devices which alter the electrical audio signals, and more particularly to devices for producing controlled distortion in audio output signals and for enhancing the tonal quality thereof.

BACKGROUND OF THE INVENTION

There are many prior art devices available which alter the tonal quality of electrical audio signals. For example, one prior art device has a distortion generator or a distortion compressor stage followed by a filter with a roll-off or attenuation with increased frequency, along with means to adjust either the amount (steepness) of the roll-off, or the point (knee) of the roll-off. However, the filter in such a device is very crude. Further the adjustment means requires the operator to experiment with different settings or combinations of settings in order to define a desirable sound, and even then the device is limited in the quality of sound which it is capable of producing. Moreover, the arrangement just described does little if anything to tailor or enhance the character or quality of the tone of the signal produced by the distortion generator or compressor stage.

Many prior art devices are available for electrically introducing reverberation effects into an audio electrical signal. Many of these devices are susceptible to mechanical jarring, and produce "Boing" type sounds when subject to such jarring or mechanical vibration and from short transient sounds. At least one prior art reverb unit incorporates a multiple output bucket brigade device, i.e. analog shift register. However, for certain applications this device does not provide sufficient delay of the inputted signal, produces undesireable echo with pulse inputs, and is limited in the type and quality of the reverb that it provides.

SUMMARY OF THE INVENTION

An object of the invention is to add controlled distortion to an audio signal to change the dynamics or sustain characteristics of an audio signal, and to alter the tonal quality of the audio signal.

A further object of the invention is to add reverberation to an audio signal such that the resultant signal has superior reverberation characteristics.

In accordance with the present invention, different combinations of filters and other devices are connected serially in different chains. In one form of the invention, a mid band pass filter receives an electrical audio input signal and provides the output to a distortion amplifier which receives the output of the mid band pass filter and adds higher harmonic audio signals to the received signal, compresses it further, and alters the waveform. A complex filter receives the output of the distortion amplifier and provides an output signal having enhanced tonal qualities. The complex filter has a roll-off of increased attentuation with increased frequency range, a boost in the low frequency range, a dip in the upper portion of the low frequency range, a dip in the mid audio frequency range, a dip followed by a peak in the upper frequency portion of the mid audio frequency range, followed by a roll-off of increased attenuation with increased frequency in the upper audio frequency range.

In another form of the invention, a high pass audio filtering circuit receives an electrical audio input signal and provides an output signal to a compressor circuit which produces an output signal having increased sustain. A complex filter with characteristics as described above may be provided after the compressor circuit.

In another form of the invention, a compressor circuit receives an audio signal and produces an output signal having increased sustain, a mid pass filter receives this signal, and the filtered signal is provided to a distortion amplifier which adds more compression and higher audio harmonic signals. A complex filter, having characteristics as described above, may be provided after the distortion amplifier.

In one form of the invention for providing reverberation, a timed turn on gate receives a main audio signal and gates this signal to an analog shift register device only after this signal exceeds a certain signal level for a certain time period. The analog shift register provides delayed output signals at a plurality of staggered delay taps. At least one summing device receives the output signals at several delay taps and outputs a signal having reverb characteristics or delay ("echo") components. By providing a timed turn on gate in front of the analog shift register, much unwanted noise of short duration and transient peaks at the start of notes are removed and therefore an output signal having higher quality reverberation is obtained.

In another form of the invention for providing reverberation to an audio signal, an analog shift register receives a main audio signal and provides delayed outputs at a plurality of staggered delay taps. An output delay circuit receives an output signal from one of the staggered delayed taps, preferably the last in the series, and delays the received signal a time period substantially different from the delay time period between any two of the adjacent staggered delay taps. Two summing devices receive output signals from the delay taps, and one of the summing devices receives the output from the output delay circuit. By summing the signals inputted thereto, the summing devices provide two different channels of audio output signals having different delay components. The output delay circuit following the analog shift register provides additional reverberation components to the resultant output signal, which is different from the sound obtained by using a single analog shift register.

Numerous other advantages and features of the present invention will become readily apparent from the following detailed description of the invention and one embodiment thereof, from the claims and from the accompanying drawings.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is an overall block diagram of the electronic audio signal processor according to the invention;

FIG. 2 is an electrical schematic of a portion of the block diagram of FIG. 1, showing the input buffer amplifier stage, the high pass filter stage, the compressor with switchable equalization, another high pass filter stage, a mid band pass filter stage and controlled distortion amplifier stage;

FIG. 3 is an electrical schematic diagram of some of the blocks of FIG. 1, including the low pass filter stage, the complex filter stage and the timed turn on gate for the reverberation device;

FIG. 4 is an electrical schematic diagram of the synthetic doubling circuit stage of FIG. 1; and

FIG. 5 is an electric schematic diagram of certain of the blocks of FIG. 1, including the bucket brigade stage, the delay output circuit, and the output amplifiers and mixers.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

While this invention is susceptible of embodiment in many different forms, there is shown in the drawings and will herein be described in detail one specific embodiment with the understanding that the present disclosure is to be considered as an exemplification of the principles of the invention and is not intended to limit the invention to the embodiment illustrated. While the description of the preferred embodiment may at times refer to audio signals from musical instruments such as electric guitars, it is to be understood that application of the invention is not limited to musical instruments or electric guitars.

As used herein, the term "low" when used in conjunction with low pass filters and the like is intended to refer to a range starting at about 50 Hz and ending at about 250 Hz to 800 Hz. In the same context, the word "middle" or "mid" is intended to refer to the range starting at about 250 Hz to 800 Hz and ending at about 2 KHz to 5 KHz. Lastly, the word "high" is intended to refer to the range starting about 2 KHz to 5 KHz and ending somewhere in the upper audio frequency spectrum.

The compressor as described herein is intended to refer to a device which compresses the intensity range of the output signal as compared to the range of the input signal, and more particularly to a device which amplifies weak signals and attenuates strong signals to produce a smaller output range for a given input range. The distortion amplifier is intended to refer to a device which functions as a linear amplifier up to a certain point of input signal level and then clips above that certain level in order to produce controlled distortion. In the preferred embodiment, the distortion amp functions to cause intermodulation of the input signals and to produce high harmonics of the low range and mid range audio content of the input signal, generally independently of the high range content of the input signal. The doubler (synthetic doubler) produces an output signal which varies in pitch from its input signal, so that its output signal simulates an instrument different from the instrument providing the input signal. When the output of the doubler is combined with the input by a summer or mixer the result is like two separate instruments.

For purposes of description, the preferred embodiment according to the invention has two main portions: a controlled distortion and tone alteration and sustain alteration portion, and a reverberation portion.

The portion of the preferred embodiment which is directed to controlled distortion tone alteration and sustain operates in one of four modes, as controlled by a selector switch. In each mode a different combination of filters and devices are connected serially in a chain after a buffer amp **10** and high pass filter **11** as shown in FIG. **1**. The filter **11** increases the mid and some of the high range part of the input signal which decay faster, causing the compressor to react more to the mid range part of the signal than to the low range part of the signal. This allows the compressor to maintain the mid range at a more constant level as a note decays, which is more pleasing when heard directly, and is important when its output is connected to the distortion amp **16** and a complex filter **17**. In the second mode, the chain consists of the compressor **12** with the high end EQ boost **12A**, a high pass filter **13** and the complex filter **17**. In the third mode, the chain consists of the compressor **12** without the high end EQ boost **12A**, the high pass filter **13** and the complex filter **17**. In the fourth mode, the chain consists of the compressor **12** without the high end EQ boost **12A**, and a low boost EQ **15**.

In the first operational mode, the distortion amp **16** is used for adding substantial controlled distortion. The mid band pass filter **14** reduces the high and low signal content before the signal goes through the distortion amp **16**. Rolling off the highs results in less noise at the output of the distortion amp and reduces the amount of highs from the input signal heard after the distortion amp **16**. This is important because in this substantial distortion mode it is important that the high end content of the output signal be made up primarily of high harmonics produced by distorting the mid range portion of the signal which are of long duration, rather than by the natural high harmonics contained in the input signal which are of short duration. Also, the high pass filter **11** is modified in this mode by opening the switch **100** which causes the filter to level off at a lowered frequency thus providing less high end content. The rolling off of the lows is important as this reduces modulation of the output signal by the low end content of the input signal. Actually, the low signal content is reduced twice; once at the high pass filter **11** after the buffer amp **10**, and again at the mid band pass filter **14**.

The compressor **12** receives a wide amplitude range of signals and outputs an output signal having a relatively narrow amplitude range. The compressor **12** is designed so that its output is fixed at a good level for generating harmonics within the distortion amplifier **16**. Therefore, one advantage of having the compressor **12** in front of the distortion amp **16** is so that the harmonics generated by the distortion amp **16** can be controlled by the operation of the compressor **12**.

The importance of the compressor **12** will be understood more readily if one considers what the resultant signal would be like without a compressor. If a distortion amplifier were to receive signals directly from a stringed musical instrument a very loud signal is produced when the string is first plucked, and a certain associated distortion characteristic will be produced. When the signal dies out or decays, the character of the signal changes dramatically. Therefore the difference in distortion outputs, with the signal increased, is very pronounced and significant.

One aspect of the invention is directed to minimizing the difference between the initial output of the distortion amplifier **16** and the subsequent sustained output of the distortion amplifier. In order to get sustain out of a musical note, a compressor **12** is used to prevent the signal from dying out or decaying as quickly and keeps the signal near a maximum output level for a certain time period. This signal is fed into the distortion amplifier **16** or distortion generator which generates harmonics.

The mid band pass filter **14** in front of the distortion amplifier **16** is fairly important in obtaining a distorted musical sound having a good waveform quality, as is the compressor **12** bipass EQ **11**. The complex filter **17** which receives the output of the distortion device, processes this output into an ouput signal having excellent tonal qualities. Without this filter, the output would be both "harsh" and "muddy" in tonal quality.

In a second operational mode, the gain of an operational amplifier in the compressor stage **12** will be reduced, thereby cancelling some of the effect of the compressor unit **12** and reducing the level of the signal going into the distortion amp **16**. The distortion amp **16** will not stay in the distortion state quite as long. Each time a note is played on the guitar, distortion will occur, but only for a brief time period.

The distortion amplifier **16** produces more high harmonics as the amp **16** is driven harder. Therefore, when the distortion amp **16** is not driven hard, fewer high harmonics are produced. In order to compensate for this, a high end EQ boost **12A** (high pass filter) can be switched into in the compressor state **12**, resulting in additional high end signal content, when this reduced gain mode is selected.

As the signal decays, the generated highs will diminish as the distortion amp **16** returns to the linear range of operation and no longer outputs a distorted signal. Since the distortion amp is no longer producing as much high end, a high end EQ boost **12A** in the compressor is switched in this second mode. The high end produced will compensate for the fact that the distortion amp **16** is not producing as much high end, resulting in approximately the same amount of high signal content, but without as much distortion. This mode of operation may be desirable for guitar players who desire only a slight amount of distortion for pop music, instead of heavy rock and roll type sustained distortion.

The importance of having the high end EQ boost **12A** before the distortion amp **16** can be illustrated by considering what sound would result by having a high end EQ boost after instead of before a distortion amp. Then the high harmonics synthetically generated by the distortion amp would also be amplified or boosted, and the distorted tones would be boosted, and the true guitar sounds would be masked too much by the distorted guitar tones. However, by putting a high end EQ boosst before the distortion amp **16**, the boost has substantially no effect on the high harmonics that the distortion amp produces because the output of the distortion amp is more dependent on the mid range content of the signal than the high range. Therefore, it is important that the high end EQ boost **12A** associated with the compressor **12** be placed in front of the distortion amp **16** when the distortion amp is driven at lowered signal levels. This output is then processed by the complex filter **17** to improve its tonal qualities.

In the third operational mode, the chain consists of the compressor **12** without the high end EQ boost **12A**, a high pass filter **13** and the complex filter **17**. This operational mode might be used by musicians who desire a clean sound without controlled distortion. The distortion amplifier **16** used in the first operational mode outputs a relatively large amount of high end signal content by adding high harmonics. Since the distortion amplifier is not used in this operational mode, the high pass filter **13** increases the higher harmonic content of the signal and thus compensates for the absence of the distortion amplifier **16**. The complex filter **17** was designed primarily to process the output of the distortion amplifier **16** but is used in this mode to make the tone more similar to that of the first and second operational mode. The complex filter **17** functions so that its output has a relatively large amount of low end and mid range signal content and rolls off dramatically at its upper end due to the large high end signal content produced when the distortion amp is being used. However, since the distortion amplifier is not used in the third operational mode, instead of eliminating the complex filter and replacing it with a separate second complex filter for use in this second operation mode, a simpler high pass filter **13** is provided in cascade with the complex filter **17**. The high pass filter **13** will compensate somewhat for the bass heavy response of the complex filter **17**.

Since the complex filter **17** has a peak in the mid range at about 500 Hz with a dip at 250 Hz and 1.6 KHz, the device will process the signal from a rather toneless guitar into a signal with enhanced tonal qualities in the same way the good stringed instruments with good tonal qualities have heavy response areas in the mid range. For guitars which already have good tonal response in the mid range, some additional mid range tone will be obtained.

In the fourth operational mode, the chain consists of the compressor **12** without the high end EQ boost **12A**, and a low end EQ boost **15**. This operational mode omits the distortion amplifier **16** and complex filter **17** present in other operational modes, and is primarily for keyboard instruments or for jazz guitarists who want a truer sound without substantial emphasis or de-emphasis of the tonal qualities of the musical instrument. The lower end of the audio frequency spectrum is boosted by the low end lost through the high pass filter **11**. However, total compensation is not achieved, since if the high pass filter **13** and low pass filter **15** are superimposed, the resultant filter would be flat from 50 to 400 Hz and then climb to about 5 KHz where it would flatten out.

Referring now to FIG. 2, certain parts of the controlled distortion and tone alteration portion of the preferred embodiment will now be described in greater detail. Buffer amplifier **10** comprising integrated circuit IC **101A** receives an electrical input signal from a musical instrument or any other device producing audio signals through monaural connector CN **102** and resistor R **101**. The output of the buffer amplifier **10** is provided to a high pass filter circuit **11** comprising resistors R **102** and R **103**, capacitor C **103** and switch SW **100**.

Switch SW **100** provides a means to adjust the point of the roll-off or knee between one frequency position of about 5 KHz (for "clean" sounds) and a higher frequency position (for "distorted" sounds). The high pass filter **11** has a roll-off of increased attenuation with a decrease in frequency of about 6 db per octave. When the switch position dictates a lower knee, the gain of the mid-range is higher by about 6 db. Accordingly, with the increase in gain the large signal inputted to the op amp IC **101B** will probably push it into distortion at all times. Actually SW **100** is mechanically tied to SW **101**, so that SW **100** is open only when SW **101** is in its uppermost position. In this position the device operates in the first mode, i.e. with the mid band pass filter, without the high end EQ **12A** in the compressor stage **12**.

The output of the high pass filter **11** is provided to a compressor circuit **12**. As explained above, the compressor circuit **12** amplifies weak signals and attenuates strong signals to produce a smaller amplitude range compared to the amplitude range of its input. The compressor circuit comprises essentially an amplifier IC **101B** and an FET transistor Q **101** which serves to compress or reduce the amplitude range of the signal appearing at the input of amplifier IC **101B**.

The output of the op amp IC **101** B goes through two resistors R **169** and R **170** to ground. The signal between those resistors goes through a diode D **101** to the gate of FET Q **101**. When the output of the op amp IC **101**B exceeds a certain level the resistance for the FET goes up and cuts down the feedback of the op amp. Between the junction of resistors R **169** and R **170** and ground is a diode D **119** which serves to limit the amount of compressing that the FET can perform. When the output signal from the op amp increases, diode D **119** effectively reduces the resistance across resistor R **170**. As soon as the signal gets above the threshold level of this diode D **119**, the signal is passed to ground. Therefore, as the signal gets larger, the FET gate increases resistance until it gets to a certain point. At that point the signal level across the gate of the FET will not increase. If the op amp signal increases, the FET stops compressing at a certain point and intentionally lets the signal build up going through the op amp.

One reason why an upper limit is placed on the FET is related to the operating characteristics of the FET. As the signal increases at the gate of the FET, the resistance across it increases. At first the resistance goes up smoothly and relatively linearly. However, above a certain point the resistance goes up very quickly. This would reduce the gain of amp IC **101** B drastically until capacitor C **106**, which charges up in response to signals, could discharge. A large signal across this capacitor would keep it charged and it would take a long time for the signal to bleed off. Therefore, if diode D **119** was not connected, a large signal could charge the capacitor keeping the FET at a high impedence, and one would not be able to hear weaker sounds played immediately after it. The discharge time of capacitor C **106** is set long enough to produce smooth decay of sounds in the guitar frequency range.

On a guitar the first sound or pulse that comes out can be a huge peak which is almost always much stronger than the signal which follows within a few milliseconds. A guitar amplifier tends to smooth out these sounds because it cannot respond to them fast enough, because it clips (distorts) large signals, and because the speakers have slow response. If the amplifier is turned up high it will simply distort the output amp or the speaker or both for those few milliseconds, and one will hear extra harmonics on the front of the note, without any large pulse coming through.

In accordance with the invention for louder notes, the signal is normally compressed, and the peaks are held to just below where the op amp is starting to clip. The signal immediately following is amplified up to this same point as capacitor C **106** discharges within about 50 milliseconds or less. Any extra signal will not be compressed since the diode D **119** prevents the signal at the FET from surpassing a certain limit.

Thus for overly large signals, the peak of the signal will cause distortion of the op amp TC **101** B, which is acceptable because distortion is a widely understood indicator that the input signal is too large, and the musician will likely reduce the volume of the instrument. Also, the clipping (distortion) of peaks is often accepted as normal for guitar amplifiers.

The above described arrangement not only results in obtaining sustain out of the guitar, it also eliminates large pulses at the front and keeps them down to a moderate level.

Compressor circuit **12** also includes a switchable high end EQ boost portion **12A** comprising resistors R **109**, R **110** and capacitor C **105**. When switch SW **101** (the operation of which will be described in greater detail below) is in its second upper position, the high end EQ boost portion **12A** is switched into the IC **101** B feedback loop, so that the high pass filter with a knee at about 2 KHz is added to the compressor circuit **12**.

The high pass filter **13** comprises a resistor R **111** and capacitor C **107** and is connected in the circuit when the switch SW **101** is in the third and fourth positions. The filter is ineffective in the fourth position, however, due to the high input impedence of filter **15**.

The mid band pass filter **14** comprises resistors R **112** and R **113** and capacitors C **108** and C **109**. The mid band pass filter **14** receives its input from the output of the compressor circuit **12** and outputs a filtered signal which is fed to the input of distortion amp **16**.

Distortion amp **16** comprises an integrated circuit IC **102A**, and a feedback loop comprising diodes D **102** through D **105** and resistor R **114**. The diodes serve to clip both the negative and positive going amplitudes of the output voltage to produce distortion when the input signal level is above a certain point. However below that certain point, the distortion amplifier **16** functions essentially as a linear amplifier. The output of distortion amplifier **16** is provided to a terminal of switch SW **101**.

Switch SW **101** is a **10** terminal, four position slide switch having right and left slide members which are insulated from each other but which move together by a manual switching actuator. Each of the right and left slide members connect two adjacent terminals at a time. Thus, when the switch is in the extreme upper position, the upper two terminals on each side will be connected to each other. In the upper position, the controlled distortion portion of the preferred embodiment operates in the first mode (i.e. the middle chain with the mid band pass filter). In this position the output of the distortion amp **16** is connected to the input of the complex filter **17**, and the EQ portion **12A** of circuit **12** is not connected. When switch SW **101** is connected in the second uppermost position, the condition of the device is essentially the same as just described, except that the equalization portion **12A** is connected in circuit with compressor section **12**, so that the controlled distortion portion of the preferred embodiment operates in the second mode.

When switch SW **101** is in its third uppermost position, the output of high pass filter **13** is connected to the input of complex filter **17** so that the control distortion portion of the preferred embodiment operates in the third mode of operation. Also, the EQ portion **12A** of compressor circuit **12** is not connected. When the switch SW **101** is in its lowermost position, the output of high pass filter **13** is connected to the input of low pass filter **15** and the control the fourth operational mode, and equalization portion **12A** of compressor circuit **12** is not connected. Note that, as explained earlier, the high pass filter **13** does not substantially boost the high end in this mode.

Referring now to FIG. **3**, the complex filter **17** comprises three substantially similar cascaded amplifier and filter stages having different value resistors and capacitors which define different frequency response characteristics for each of the stages and a passive filter stage providing a lower pass filter at the beginning. When cascaded together, the resultant frequency response is that shown in FIG. **1**, i.e. a roll off of increased attenuation with increased frequency from 80 Hz to 250 Hz of about 4 db per octave, a decrease in attenuation with increased frequency to a peak at 500 Hz, followed by a dip at about 1.6 KHz and a peak at about 4 KHz, and a roll-off of increased attenuation with increased frequency of over 12 db per octave in the upper audio frequency range at frequencies above 4 KHz.

The low pass filter **15** as shown in FIG. **3** comprises an amplifier IC **104B**, input resistor R **130** and a feedback loop comprising resistors R **131**, R **132** and capacitor C **117**. The frequency response of the low pass filter **15** is shown in FIG. **1** and has a generally flat response below 50 Hz, with increased attenuation with increased frequency between 50 Hz and 400 Hz, with a generally flat response above 400 Hz. As described above, low pass filter **15** is switched into the circuit when SW **101** is in the lowermost position, i.e. the fourth operational mode.

The portion of the preferred embodiment which is directed to reverberation comprises a doubling circuit **18**, a timed turn on gate **19**, an analog shift register bucket brigade device **20** with delay taps including its associated input buffer amp and filter circuit **20A**, an output delay circuit **21**, an output summing and amplifier circuit **22**, and an output amplifier and mixing circuit **23**. This portion of the preferred embodiment operates in one of three modes to provide doubling alone, reverb alone, or both doubling and reverb.

Turning now to FIG. **3**, the operation of the timed turn on gate **19** will now be described. The timed turn on gate **19** receives a main audio signal which is fed into amplifier IC **102B**. Amplifier IC **102B**, in conjunction with amplifier IC **105A** and associated resistors R **133** through R **140**, capacitors C **118** through C **120** and diodes D **106** through D **110**, will effect switching of FET transistor Q **102** (to gate the main audio signal to IC **105B**) about 20 milliseconds after a main audio signal of sufficient magnitude is present on the main signal line. The main audio signal that is gated comes through resistor R **141**.

When the input signal is low the resistance across the FET will be low and the signal will be attenuated to a very low amount, essentially off. When the signal to the FET is high, the FET will turn on and open its gate to let the main audio signal pass virtually unattenuated as long as a certain amount of voltage is maintained at the gate of the FET. The value of capacitor C **120**, in conjunction with resistor R **138**, determines the turn on time which is about 40 milliseconds. As soon as a signal of sufficient magnitude appears at the input of IC **102B**, the signal at the output of IC **102B** begins charging capacitor C **120**. When C **120** is charged to a sufficient amount, the signal is passed to IC **105A**. Therefore, adequate turn on voltage does not get to the FET gate for 40 milliseconds after the signal is present at the input of op amp IC **102B**.

Capacitor C **120**, in conjunction with R **139**, sets the release time of the timed turn on gate which is a few milliseconds. Thus, if the signal voltage suddenly drops, the voltage across the capacitor C **120** will not disappear immediately, but will bleed off gradually through resistor R **139**. Therefore, the FET will not clamp down shut suddenly but instead will slowly turn off so that the sound into the reverb does not end abruptly.

By providing a timed turn on gate some unwanted noise spikes of short duration (e.g. a few milliseconds), and most high amplitude peaks at the start of "stuccato" guitar notes, are eliminated. Without a timed turn on gate according to the invention, the spikes would pass to the main reverb unit and would result in numerous discrete echoes. One way to reduce the effect of spikes is to provide a large number of echo repeats, i.e. about 300 repeats per second. However, this would be quite costly. Therefore, by providing a timed turn on gate according to the invention, spikes will be eliminated even in reverb units having a small number of stages. If a note is played and then another note is played immediately thereafter, the reverb is already turned on so a spike would get through, but the spike would not be noticed because program material would mask it.

The doubling circuit **18** essentially functions to simulate a second instrument which is slightly off key and slightly out of time with an initial instrument. This is done by cyclicly varying the pitch of the initial instrument signal back and forth about its nominal pitch. For example, if the nominal pitch of the initial instrument signal is an F note then the doubler will output a sharp F note for a while and then a flat F note for a while followed by a sharp F note again and so on.

Cyclic pitch variation can be achieved by inputting the initial instrument signal into an analog delay device and then varying the clock frequency of the clock which drives the delay device. If the delay device is a bucket brigade, the bucket brigade receives an initial instrument signal and shifts the signal within the brigade from bucket to bucket at speed determined by the frequency of the clock which drives the bucket brigade. By varying the frequency of the clock signal the pitch of the signals passed by the buckets can be varied. By reducing the clock frequency the pitch will reduce. To hold the pitch at the reduced pitch level, one must keep reducing the clock speed at the same rate of change. However if this is continued the resultant delay of the bucket brigade will be delayed further and further until eventually the output would be minutes behind its input. In order to provide a pitch differential while still keeping the overall delay to about 15 to 20 milliseconds, the pitch is increased and then reduced and so on in a cyclical manner. Of course the delay will vary within the range of about 15 to 20 milliseconds.

The doubling circuit **18** comprises essentially two circuit portions: an analog delay portion **18A** and a delay clock portion **18B**.

The analog delay portion **18A** comprises a bucket brigade device IC **110** which has an input buffer amp IC **106A**, and an output buffer amp IC **106B**, each having associated resistors and capacitors as shown. The bucket brigade IC **110** at its pins **2** and **6** receives a series of clock pulses of opposite phase from IC **109**. IC **108** and **109** create a high frequency clock whose frequency varies about a nominal rate.

In order to create a slow variation in this clock rate, a low frequency oscillator comprising IC **107** A and B, along with associated resistors and capacitors, provides a triangle waveform signal of frequency about $0.5H_2$ to pin **3** of IC **109**. In response to this triangle wave form, IC **108** and **109** will produce clock pulses of slowly varying frequency. The bucket brigade will respond to these clock pulses to cyclicly vary the pitch of its output signal to either side of the pitch of its input signal. The output of the doubling circuit will thus simulate a second instrument slightly off key and out of time with an instrument whose signal is inputted to the doubling circuit.

As shown in FIG. **5**, the output from the timed turn on gate **19** and the doubling circuit **18** is provided to terminals of switch SW **201**. Switch SW **201** is an eight terminal three position slide switch having an upper sliding member which engages two adjacent terminals at a time, and a lower sliding member which also engages two terminals at a time and moves in conjunction with the upper sliding member. The sliding members are moved by manual switch actuating element. When the switch actuator is on the extreme left, the reverberation portion of the preferred embodiment provides a doubling output but no reverb output to the output mixers. When the switch actuator is in the middle position, the reverberation portion of the preferred embodiment will provide both a doubling component and a reverberation component to the output mixers. When the switch actuator is on the extreme right, the reverberation portion of the circuit will provide a reverberation signal but no doubling component to the output mixers.

When switch SW **201** is in either the middle or extreme right position, the bucket brigade circuit **20** will receive a signal at the input of its buffer amplifier and filter circuit portion **20**A. The buffer amplifier and filter circuit portion comprises two integrated circuits IC **203**A and IC **203**B, and associated resistors and capacitors, and provides an amplified and filtered signal to pin **12** of the bucket brigade device IC **206**. The integrated circuit IC **206** is an analog shift register having 6 output delay taps at pins **4–9** thereof.

Integrated circuit IC **208** is an analog shift register clock generator/driver which drives both integrated circuits IC **206** and IC **207**. The period of the switching of the timer is dependent upon the circuit values of resistors R **254**, R **255** and capacitor C **228**. The bucket brigade IC **206** receives an input signal at pin **12** and provides this signal at different delay periods to the output delay taps (pins **4–9**). The delay between adjacent delay taps is about 15 to 40 milliseconds, so that the input signal is outputted at the first delay tap (pin **9**) about 25 milliseconds after it is received at pin **12**. The signal is outputted at the last delay tap (pin **4**) about 150 milliseconds after it is received at input pin **12** of IC **206**. The output of the last delay tap (pin **4**) is provided to pin **3** of an additional output delay integrated circuit chip IC **207**, which is also an analog shift register like IC **206**, but with fewer stages. The IC **207**, at pins **7** and **8**, provides a delayed output about 50 milliseconds after it receives an input at pin **3**.

The output of output delay taps **4–9** of bucket brigade IC **206** and delay taps **7** and **8** of IC **207** are fed into a resistor summing network comprising resistors R **245** through R **251**. As seen from the Figure, the outputs of alternate pins **4**, **6** and **8** are summed on the lower output line (left channel), whereas the outputs of alternate pins **5**, **7** and **9** are summed on the upper output line (right channel). Further, the output of the additional output delay chip IC **207** is fed to the upper output line only. The output of the upper output line (right channel) is fed to the input of a right output amplifier and filter comprising integrated circuits IC **204**A and IC **204**B, associated resistors R **225** through R **230**, and capacitors C **216** through C **220**. The output of this right output amplifier and filter appearing at pin **7** of IC **204**B is connected to a resistor R **204** at the input of output amplifier and mixing circuit **23**.

Similarly, the output of the lower line of summing resistors (left channel) is fed to the left output amplifier and filter circuit comprising IC **205**A and IC **205**B, associated resistors R **231** through R **236**, and capacitors C **221** through C **225**. The output of the left output amplifier and filter circuit appears at pin **7** of IC **205**B and is connected to resistor R **209** at the input of output amplifier and mixing circuit **23**.

The output amplifier and mixing circuit **23** comprises essentially two different, but substantially identical, output amplifier and mixing circuits **23**A and **23**B. The upper output amplifier and mixing circuit **23**A comprises four input summing resistors R **202** through R **205** and an amplifier mixer IC **202**A. In like manner, the lower output amplifier and mixing circuit **23**B comprises four input summing resistors R **206** through R **209** and an amplifier mixer IC **202**B.

The main signal from the controlled distortion and tone alteration portion of the circuit always appears at the left side of input summing resistors R **202** and R **206**. When switch SW **201** is in the middle or right position, reverberation signals will appear at the left side of input summing resistors R **204** and R **209**. A doubling signal will appear at the left side of input summing resistors R **203** and R **207** when switch SW **201** is in either the left or middle position, but not when SW **201** is in the right position. However, when SW **201** is in the right position, the main audio signal will appear at the left side of resistors R **203** and R **207** in place of the doubling signal to compensate for the absence of the doubling signal. In this way, the combined signal level of the main audio and doubling signals to each mixer is maintained relatively constant. An auxiliary input signal can be inputted to connector CN **203** if desired and will then appear at the right side of input summing resistors R **205** and R **208**. The summing resistors R **202**, R **206**, R **203**, and R **207** are chosen so that the main signal will appear to be substantially, but not entirely at one side of the stereo mix and the doubling signal will appear to be substantially, but not entirely, at the other side when switch SW **201** is in the left or middle position. This is important in order to achieve some phase cancellation between the signals and at the same time provide stereo separation between the main signal and the artificial doubled signal.

Switch SW **202** in the output amplifier and mixing circuit **23** provides a means to selectively attenuate the mixed signals in both channels before they pass through amplifiers IC **202**A and IC **202**B. Switch SW **202** is a three position, eight terminal slide switch substantially identical in structure and operation to switch SW **201**. When the switch contacts are in the extreme right position, 0 db attenuation is achieved. When the switch is in the middle position, 5 db attenuation is obtained, and when the switch is in the left position 10 db of attenuation is achieved.

The output of output amplifier and mixing circuit **23** provides two separate channels of output signals having different signal characteristics. The signals are provided to connector CN **202** which is a stereo output connector, and to terminals **1** and **2** of connector CN **201**, also a stereo output connector. The signals from these two separate channels can be provided to a sound transducer, a stereo amplifier and speaker system, a mixing console or sound recording device.

Table 1 attached hereto lists the values of the circuit components described herein. However, it is to be understood that the invention is not limited to the precise circuit values or even the specific embodiment described above, and no limitation with respect to the specific apparatus illustrated herein is intended or should be inferred. It can be appreciated that numerous variations and modifications may be effected without departing from the true spirit and scope of the novel concept of the invention. It is of course intended to cover by the appended claims all such modifications as fall within the scope of the claims.

TABLE I

R 101	10K	R 131	100K
R 102	5.6K	R 132	560K
R 103	18K	R 133	1 M
R 104	180K	R 134	1 M
R 105	12K	R 135	1 M
R 106	22 M	R 136	1 M
R 109	1K	R 137	4.7K
R 110	10K	R 138	1 M
R 111	18K	R 139	150K
R 112	3.3K	R 140	10 M
R 113	33K	R 141	120K
R 114	1 M	R 142	10K
R 115	2.7K	R 143	10K
R 116	82K	R 144	68K
R 117	8.2K	R 145	150K
R 118	100K	R 146	82K
R 119	100K	R 147	82K
R 120	100K	R 148	6.8K
R 121	47K	R 149	22K
R 122	100K	R 150	2.2K
R 123	100K	R 151	100K
R 125	13K	R 152	100K
R 126	13K	R 153	4.7K
R 127	3.9K	R 154	4.7K
R 128	2.2K	R 155	47K
R 129	2.2K	R 156	47K
R 130	120K	R 157	22K
R 158	27K	R 217	10Ω
R 159	39K	R 218	100K
R 160	220Ω	R 219	100K
R 161	120K	R 220	33K
R 162	220K	R 221	47K
R 163	6.8K	R 222	56K
R 164	330K	R 223	100K
R 165	2.7K	R 224	33K
R 166	560K	R 225	100K
R 168	10K	R 226	33K
R 169	390Ω	R 227	47K
		R 228	56K
R 171	10K	R 229	100K
R 202	120K	R 230	33K
R 203	39K	R 231	100K
R 204	220K	R 232	33K
R 205	33K	R 233	47K
R 206	39K	R 234	56K
R 207	120K	R 235	100K
R 208	33K	R 236	33K
R 209	180K	R 237	56K
R 210	2.2K	R 238	56K
R 211	2.2K	R 239	56K
R 212	1K	R 240	56K
R 213	1K	R 241	56K
R 214	2.7K	R 242	56K
R 215	2.7K	R 243	100K
R 216	10Ω	R 244	100K
R 245	100K	C 116	.001 uf
R 246	100K	C 117	.005 uf
R 247	120K	C 118	.01 uf
R 248	120K	C 119	.05 uf
R 249	150K	C 120	.05 uf
R 250	150K	C 121	3.3 uf
R 251	150K	C 122	62 pf
R 252	5.6K	C 123	1500 pf
R 253	5.6K	C 124	2700 pf
R 254	120K	C 125	22 uf
R 255	22K	C 126	3.3 uf
R 256	470K	C 127	.0033 uf
R 257	390K	C 128	.001 uf
		C 129	.15 uf
C 102	22 uf	C 130	.01 uf
C 103	.001 uf	C 131	15 pf
C 104	3.3 uf	C 132	3.3 uf
C 105	.1 uf	C 201	220 uf
C 106	.082 uf	C 202	220 uf
C 107	.01 uf	C 203	.1 uf
C 108	.033 uf	C 204	.1 uf
C 109	.01 uf	C 205	220 uf
C 110	.033 uf	C 206	220 uf

TABLE I-continued

C 111	.001 uf	C 207	.05 uf
C 112	.0082 uf	C 208	.05 uf
C 113	82 pf	C 209	.1 uf
C 114	.0015 uf	C 211	220 pf
C 115	.047 uf	C 212	220 pf
C 214	2700 pf	D 101-D 111	IN 914
C 215	2700 pf	D 112	LED
C 216	220 pf	D 113	LED ($V_B = 2.2$)
C 217	220 pf	D 201	IN 914
C 218	2700 pf	Q 101	2N 4340 FET
C 219	2700 pf	Q 102	5457 FET
C 220	2700 pf	IC 105	TL 072
C 221	220 pf	IC 106	TL 072
C 222	220 pf	IC 107	TL 072
C 223	2700 pf	IC 108	IC 7555
C 224	2700 pf	IC 109	CD 4013B
C 225	2700 pf	IC 110	MN 3007
C 226	3.3 uf	IC 201	LM 386
C 227	3.3 uf	IC 202	LM 386
		IC 203	TL 072
C 228	220 pf	IC 204	TL 072
IC 101	TL 072	IC 205	TL 072
IC 102	TL 072	IC 206	MN 3011
IC 103	TL 072	IC 207	MN 3007
IC 104	TL 072	IC 208	MN 3101
		VR 101	EVM-31G

What is claimed is:

1. An electronic audio signal processor for processing signals in the audio frequency range, comprising:
   - a mid band pass filter having an input and output and having a bandpass in the middle audio frequency range for receiving an audio input signal;
   - a distortion amplifier having an input and output and connected to receive the output of said mid bandpass filter for adding harmonic audio signals to said received signal;
   - a complex filter connected to receive the output of said distortion amplifier, said complex filter having a low audio frequency range, a mid audio frequency range including lower and upper portions thereof, and an upper audio frequency range, said complex filter having a roll-off of increased attenuation with increased frequency in the low audio frequency range, a generally flat response in the mid audio frequency range, but having a dip followed by a peak in the upper frequency portion of said mid audio frequency range, and a roll-off of increased attenuation with increased frequency in the upper audio frequency range.

2. The electronic audio signal processor according to claim 1 further including:
   - an audio signal compressor circuit before said mid bandpass filter for receiving the audio input signal and for providing the mid bandpass filter with a signal having reduced amplitude variation relative to variations in the input signal amplitude.

3. The electronic audio signal processor according to claim 2 further including:
   - a high pass audio boost stage connected in circuit with said compressor circuit.

4. The electronic audio signal processor according to claim 2 further including:
   - a high pass audio filtering circuit before said compressor circuit for receiving the audio input signal and for providing the compressor circuit with a filtered signal having a decreased low and mid range audio signal content.

5. The electronic audio signal processor according to claim 1 wherein said dip is at a frequency on the order of 1.6 KHz and said peak is at a frequency on the order of 4 KHz.

6. An electronic audio signal processor for processing signals in the audio frequency range, comprising:
a high pass audio filtering circuit for receiving an electrical audio input signal and having an input and output;
an audio signal compressor circuit for receiving the output of said high pass audio filter and for producing an output signal having reduced amplitude variation relative to the variation in the amplitude of the input signal,
a complex filter connected to receive the output of said compressor circuit, said complex filter having a low audio frequency range, a mid audio frequency range, and an upper audio frequency range, said complex filter having a roll-off of increased attenuation with increased frequency in the low audio frequency range, a generally flat response in the mid audio frequency range, and a roll-off of increased attenuation with increased frequency in the upper audio frequency range,
a distortion amplifier connected between said compressor circuit and complex filter for providing to said complex filter, a signal having a harmonic audio signal content increased relative to the harmonic signal content from said compressor circuit,
and a switch means associated with said highpass audio filtering circuit, said switch means having at least two positions including a first position in which the roll-off of the filter is at a first frequency and a second position in which the roll-off is at a second frequency higher than said first frequency.

7. The electronic audio signal processor according to claim 6 wherein the complex filter has in its mid audio frequency range, a dip followed by a peak in the upper frequency portion of the mid audio frequency range.

8. The electronic audio signal processor according to claim 6 further including:
a mid band pass audio filter connected between said compressor circuit and said distortion amplifier.

9. The electronic audio signal processor according to claim 6 further including:
an input buffer amplifier connected in front of said high pass audio filtering circuit for receiving the audio input signal and for providing the high pass filtering circuit with an amplified audio signal.

10. An electronic audio signal processor for processing signals in the audio frequency range, comprising:
a high pass audio filtering circuit for receiving an electric audio input signal and having an input and output;
an audio signal compressor circuit for receiving the output of said high pass audio filter and for producing an output signal having reduced amplitude variation relative to the variation in the amplitude of the input signal,
a complex filter connected to receive the output of said compressor circuit, said complex filter having a roll-off of increased attenuation with increased frequency in the low audio frequency range, a generally flat response in the mid audio frequency range, but having a dip followed by a peak in the upper frequency portion of said mid audio frequency range, and a roll-off of increased attenuation with increased frequency in the upper audio frequency range,
a second high pass audio filtering circuit connected between said compressor circuit and complex filter for providing the complex filter with a signal having increased high audio signal content relative to the low and mid audio signal content of the signal received from the compressor circuit.

11. An electronic audio signal processor for processing signals in the audio frequency range, comprising:
a high pass audio filtering circuit for receiving an electrical audio input signal and having an input and output;
an audio signal compressor circuit for receiving the output of said high pass audio filter and for producing an output signal having reduced amplitude variation relative to the variation in the amplitude of the input signal; and
a complex filter connected to receive the output of said distortion amplifier, said complex filter having a low audio frequency range, a mid audio frequency range including lower and upper portions thereof, and an upper audio frequency range, said complex filter having a roll-off of increased attenuation with increased frequency in the low audio frequency range, a generally flat response in the mid audio frequency range, but having a dip followed by a peak in the upper frequency portion of said mid audio frequency range, and a roll-off of increased attenuation with increased frequency in the upper audio frequency range.

12. An electronic audio signal processor processing signals in the audio frequency range, comprising:
an audio signal compressor circuit for receiving an electrical audio input signal and for producing an output signal having decreased variation in amplitude relative to the variations in the input signal amplitude;
a distortion amplifier connected to receive the output of said compressor circuit for adding audio harmonic signals to said received signal,
a mid band audio filter connected between said compressor circut and said distortion amplifier,
and a complex filter connected after said distortion amplifier, said complex filter having a roll-off of increased attenuation with increased frequency in the low audio frequency range, a generally flat response in the mid audio frequency range, but having a dip followed by a peak in the upper frequency portion of said mid audio frequency range, and a roll-off of increased attenuation with increased frequency in the upper audio frequency range,
said mid band pass audio filter having a pass band extending through a range starting at about 250 Hz to 800 Hz and ending at about 2 KHz to 5 KHz.

13. The electronic audio signal processor according to claim 12 further including:
a high pass audio filtering circuit before said compressor circuit for receiving the audio input signal and for providing the compressor circuit with a filtered signal having a decreased low and mid range audio signal content.

14. The electronic audio signal processor according to claim 13 further including:
an input buffer amplifier connected in front of said high pass audio filtering circuit for receiving the

audio input signal and for providing the high pass filtering circuit with an amplified audio signal.

15. An electronic audio signal processor for processing signals in the audio frequency range, comprising:

a first high pass audio filtering circuit for receiving an electrical audio input signal and having an output;

an audio signal compressor circuit for receiving an electrical audio signal output from said first high pass audio filter and for producing an output signal having decreased variation in amplitude relative to the variations in the input signal amplitude;

a second high pass filter coupled to the output of said compressor circuit,

a low pass filter,

a first switch means associated with said first high pass audio filtering circuit, said first switch means having at least two positions including a first position in which the roll-off of the first high pass audio filter is at a first frequency and a second position in which the roll-off is at a second frequency higher than said first frequency;

and a second switch means which selectively couples an output of the second high pass filter to an input of the low pass filter.

16. An electronic audio signal processor for processing signals in the audio frequency range comprising;

an audio signal compressor circuit for receiving an electrical audio input signal for producing an output signal,

a complex filter connected to receive the output of said compressor circuit,

a mid band pass audio filter having an input and output,

manual switch means having multiple positions including a first circuit interconnecting position in which the compressor circuit, the mid band pass audio filter and the complex filter are interconnected in series,

and a high pass equalization circuit, said switch means having a second circuit interconnecting positions in which the equalization circuit is connected with the compressor circuit while maintaining the mid band pass audio filter coupled between the compressor circuit and complex filter.

17. An electronic audio signal processor according to claim **16** further including a high pass audio filtering circuit, said switch means having a third circuit interconnecting position in which the high pass audio filtering circuit is inntercoupled between the compressor circuit and complex filter.

18. An electronic audio signal processor according to claim **17** further including a low pass audio filtering circuit, said switch means having a fourth circuit interconnecting position in which the output of the high pass audio filtering circuit is coupled to the input of the low pass audio filtering circuit.

19. An electronic audio signal processor according to claim **18** wherein said switch means is a manual slide switch having four positions in which adjacent terminals are successively interconnected.

20. An electronic audio signal processor for processing signals in the audio frequency range comprising;

an audio signal compressor circuit for receiving an electrical audio input signal for producing an output signal,

a complex filter connected to receive the output of said compressor circuit,

a mid band audio filter having an input and output,

manual switch means having multiple positions including a first circuit interconnecting position in which the compressor circuit, the mid band pass audio filter and the complex filter are interconnected in series,

and a distortion amplifier connected to receive the output of said compressor circuit via the mid band pass audio filter for adding audio harmonic signals to said received signal.

21. An electronic audio signal processor for processing signals in the audio frequency range comprising;

an audio signal compressor circuit for receiving an electrical audio input signal for producing an output signal,

a complex filter connected to receive the output of said compressor circuit,

a mid band pass audio filter having an input and output,

and manual switch means having multiple positions including a first circuit interconnecting position in which the compressor circuit, the mid band pass audio filter and the complex filter are interconnected in series,

wherein said complex filter has a roll-off of increased attenuation with increased frequency in the low audio frequency range, a generally flat response in the mid audio frequency range but having a dip followed by a peak in the upper frequency portion of said mid audio frequency range, and a roll-off of increased attenuation with increased frequency in the upper audio frequency range.

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United States Patent [19]

Scholz

[11] Patent Number: 5,133,015

[45] Date of Patent: Jul. 21, 1992

[54] METHOD AND APPARATUS FOR PROCESSING AN AUDIO SIGNAL

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[21] Appl. No.: 467,928

[22] Filed: Jan. 22, 1990

[51] Int. Cl.[5] H03G 3/00

[52] U.S. Cl. 381/61

[58] Field of Search 381/61, 98, 102; 84/621, 691

[56] References Cited

U.S. PATENT DOCUMENTS

4,405,832 9/1983 Sondermeyer 381/61
4,495,640 1/1985 Frey 381/61
4,584,700 4/1986 Scholz 381/61
4,627,094 12/1986 Scholz 381/61
4,683,589 7/1987 Scholz et al. 381/61
4,811,401 3/1989 Brown, Sr. et al. 381/61
4,995,084 2/1991 Pritchard 381/61

Primary Examiner—Forester W. Isen
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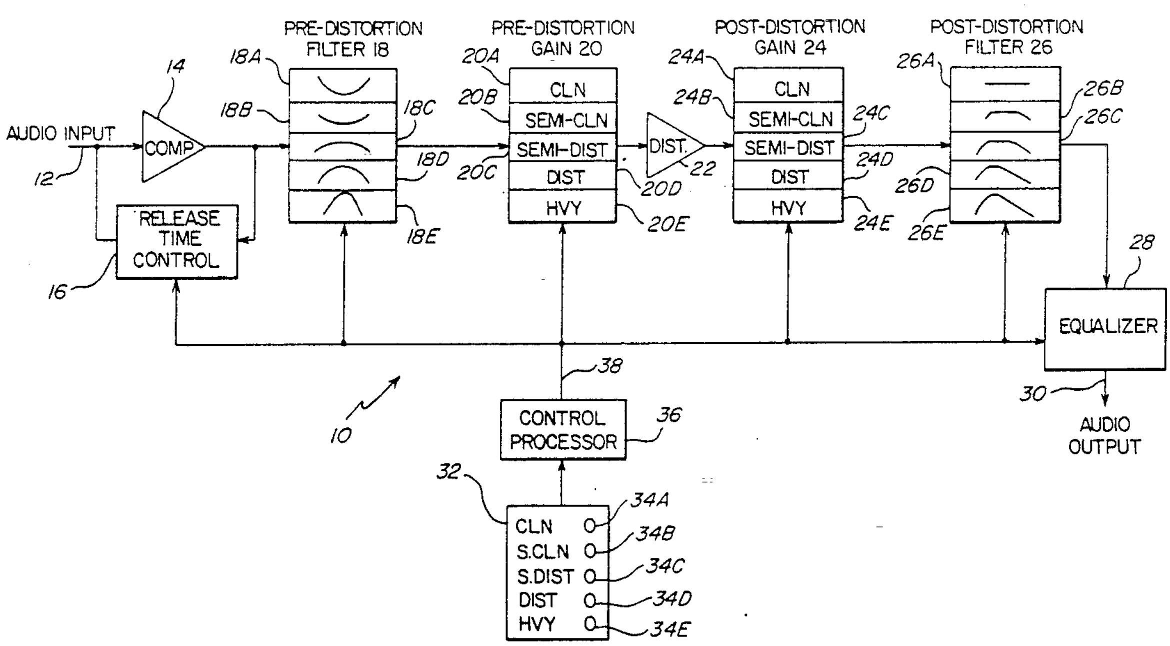
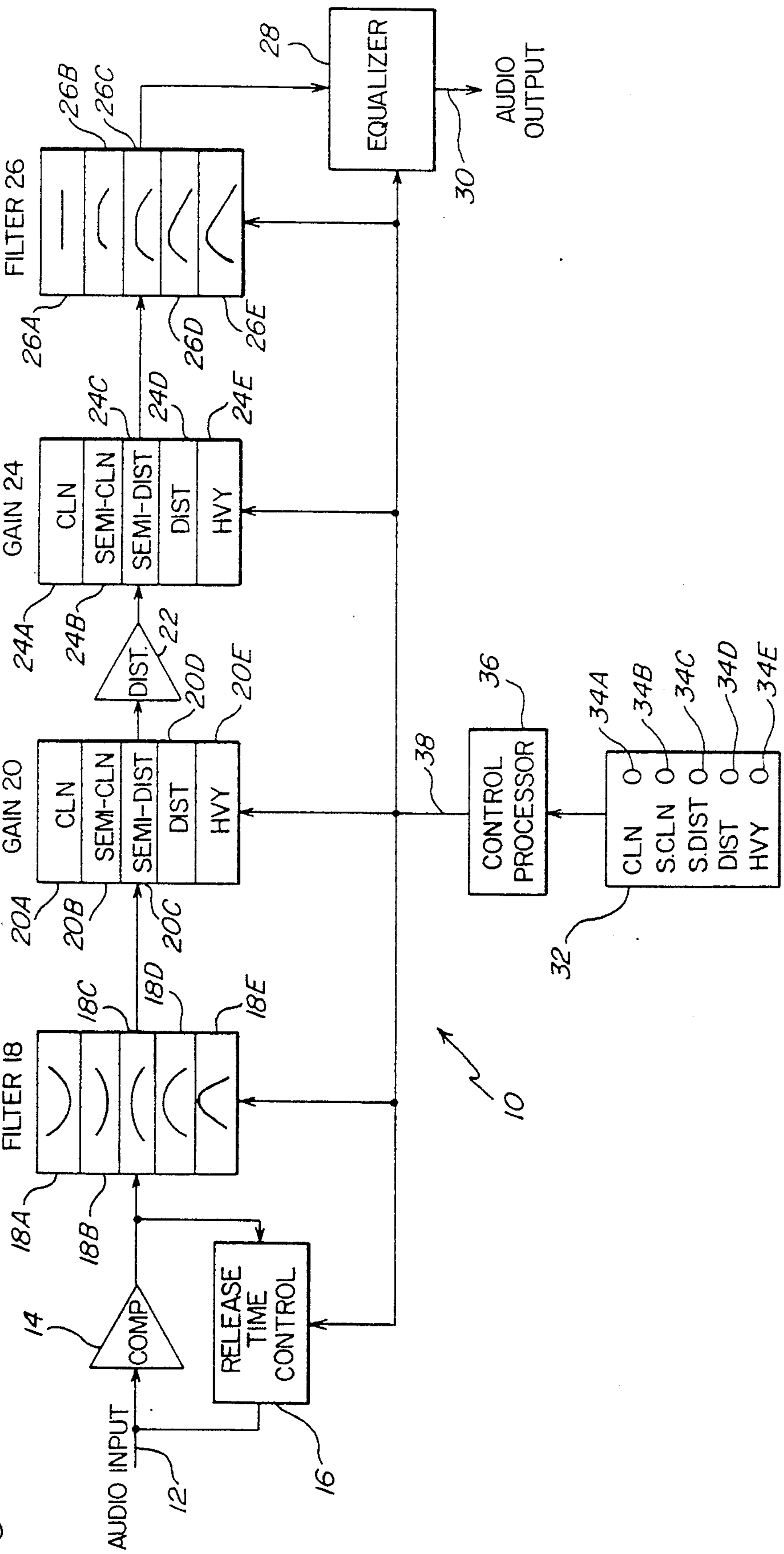
[57] ABSTRACT

An electronic audio signal processor especially suitable for electrical instruments such as electric guitars is provided, including a distortion stage, filtering stages, an output gain stage, and a compressor stage having a varibale release time. These stages are provided in a cascade connection. Further, the characteristics of each of these stages are preset to provide optimum performance at each of several different distortion levels. The correct characteristics of each stage for each distortion level is selected by means of a single user control.

44 Claims, 6 Drawing Sheets




AUDIO INPUT
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COMP
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RELEASE TIME CONTROL
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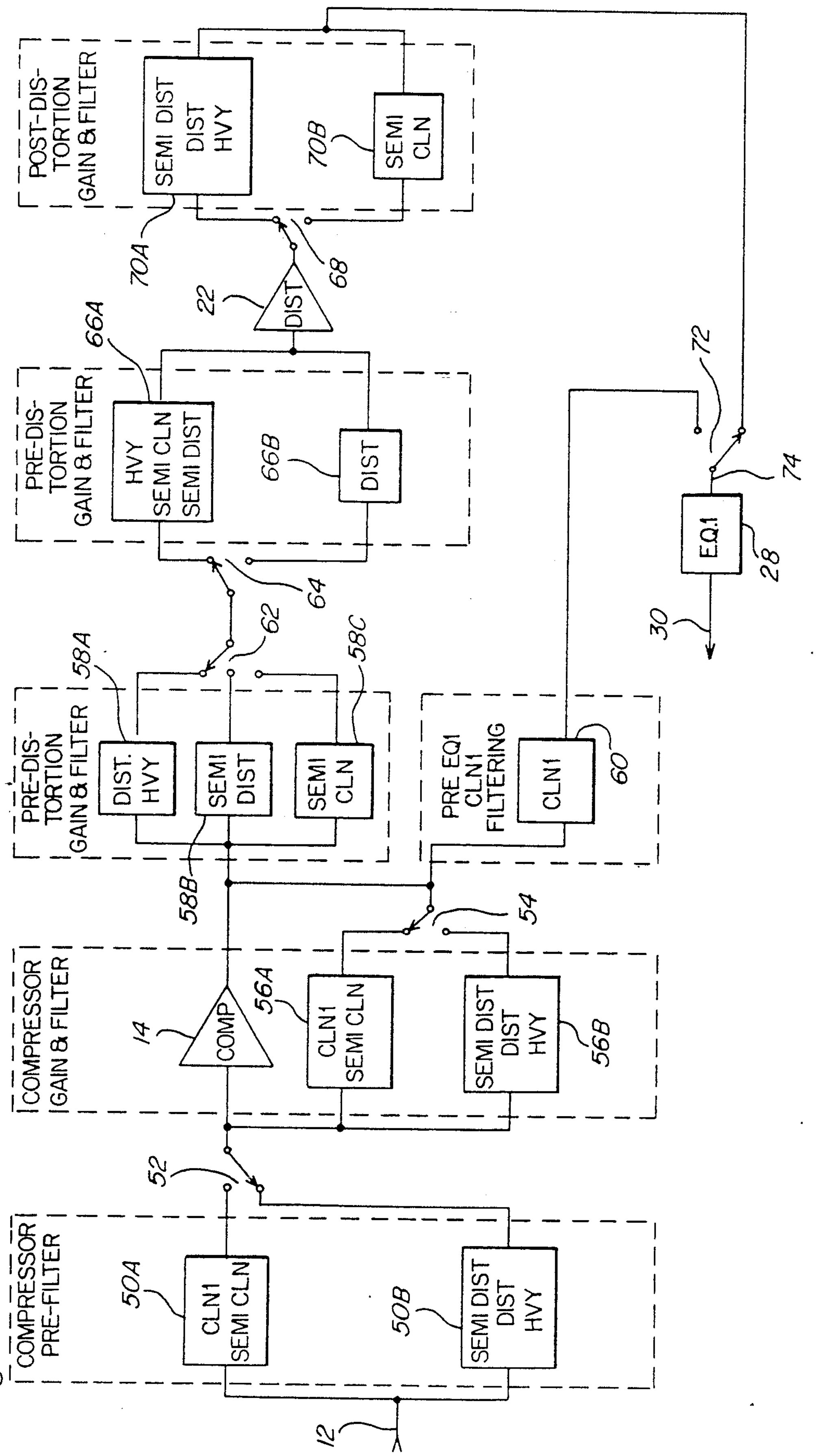
PRE-DISTORTION FILTER 18
18A
18B
18C
18D
18E
PRE-DISTORTION GAIN 20
CLN
SEMI-CLN
SEMI-DIST
DIST
HVY
20A
20B
20C
20D
20E
DIST.
22
POST-DISTORTION GAIN 24
CLN
SEMI-CLN
SEMI-DIST
DIST
HVY
24A
24B
24C
24D
24E
POST-DISTORTION FILTER 26
26A
26B
26C
26D
26E
EQUALIZER
28
AUDIO OUTPUT
30
CONTROL PROCESSOR
36
38
CLN
S.CLN
S.DIST
DIST
HVY
32
34A
34B
34C
34D
34E
10

Fig.1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SEMI DIST DIST HVY
SEMI CLN
70A
70B
68
22
DIST
66A
PRE-DISTORTION GAIN & FILTER
HVY SEMI CLN SEMI DIST
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64
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72
74
E.Q.1
28
30
58A
58C
PRE-DISTORTION GAIN & FILTER
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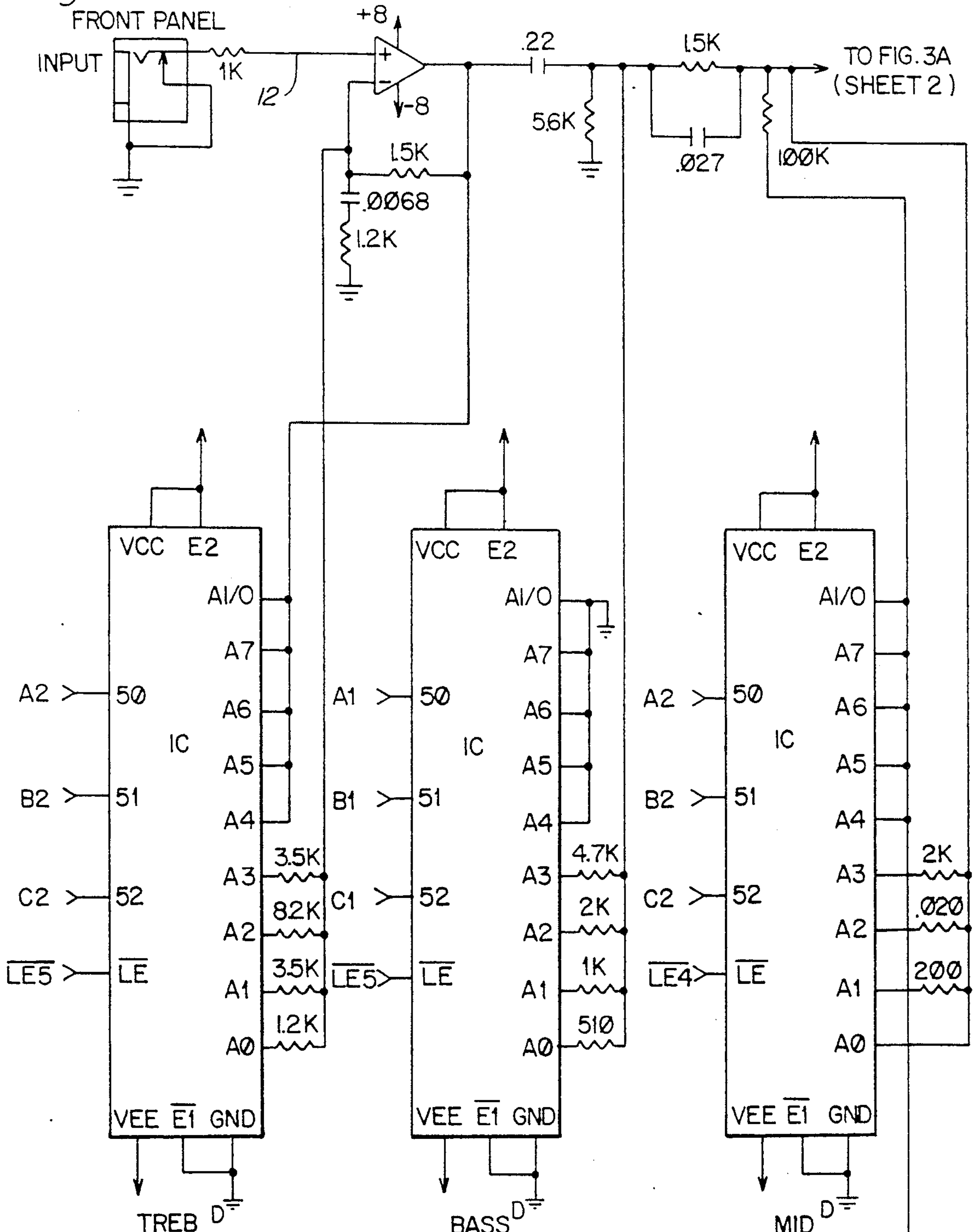
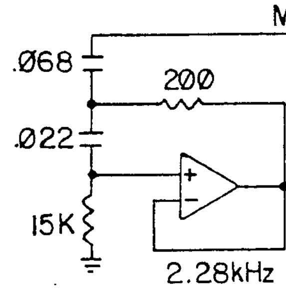
SEMI DIST
SEMI CLN
PRE EQ1 CLN1 FILTERING
CLN1
60
58B
54
COMPRESSOR GAIN & FILTER
14
COMP
56A
CLN1 SEMI CLN
SEMI DIST DIST HVY
56B
52
COMPRESSOR PRE-FILTER
50A
CLN1 SEMI CLN
50B
SEMI DIST DIST HVY
12

Fig. 2

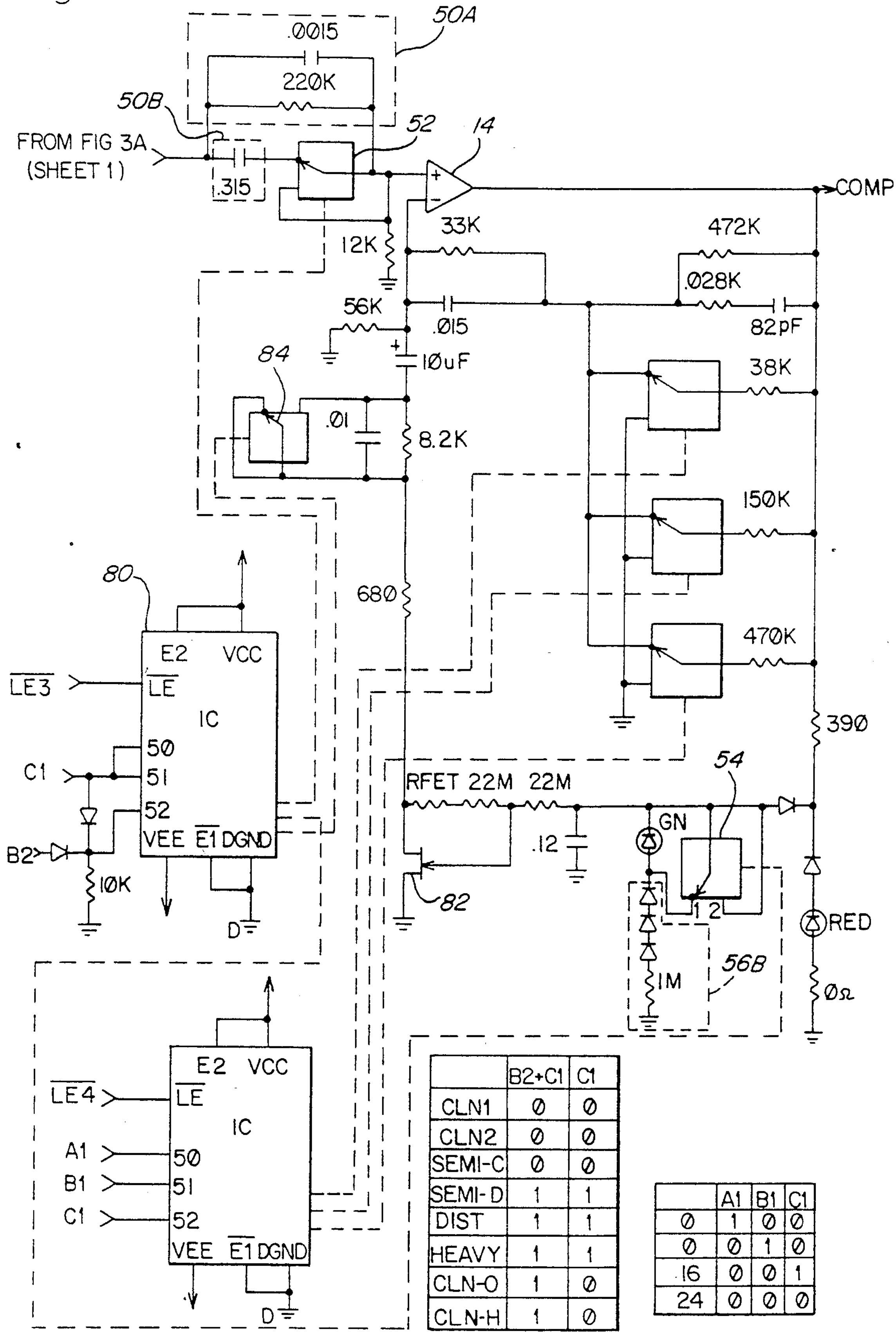
Fig.3A (Sheet 1)





	A	B	C
10	0	0	1
5	1	1	0
0	0	1	0
-5	1	0	0
-10	0	0	0

Fig.3A (Sheet 2)



	B2+C1	C1
CLN1	0	0
CLN2	0	0
SEMI-C	0	0
SEMI-D	1	1
DIST	1	1
HEAVY	1	1
CLN-O	1	0
CLN-H	1	0

	A1	B1	C1
0	1	0	0
0	0	1	0
.16	0	0	1
24	0	0	0

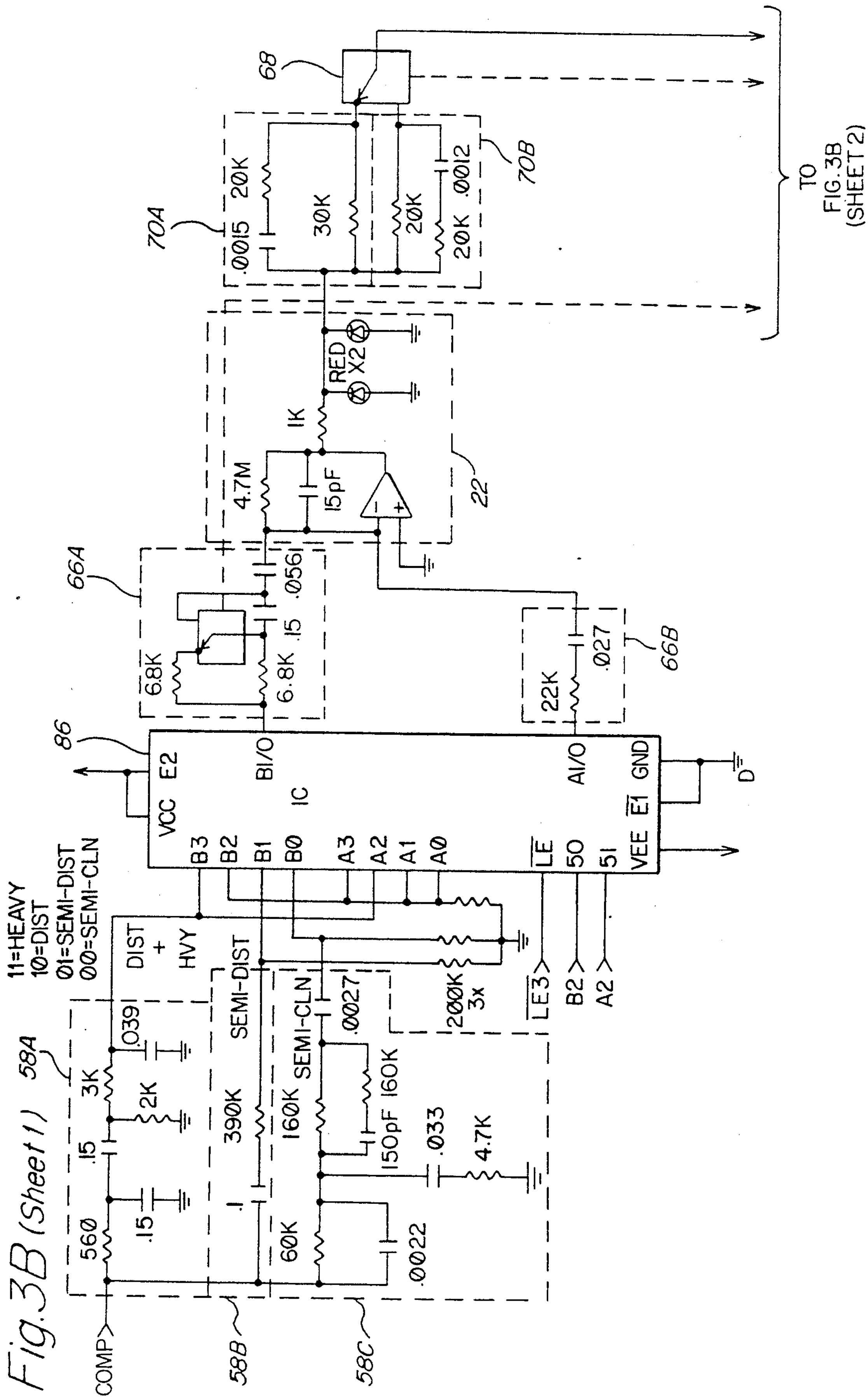
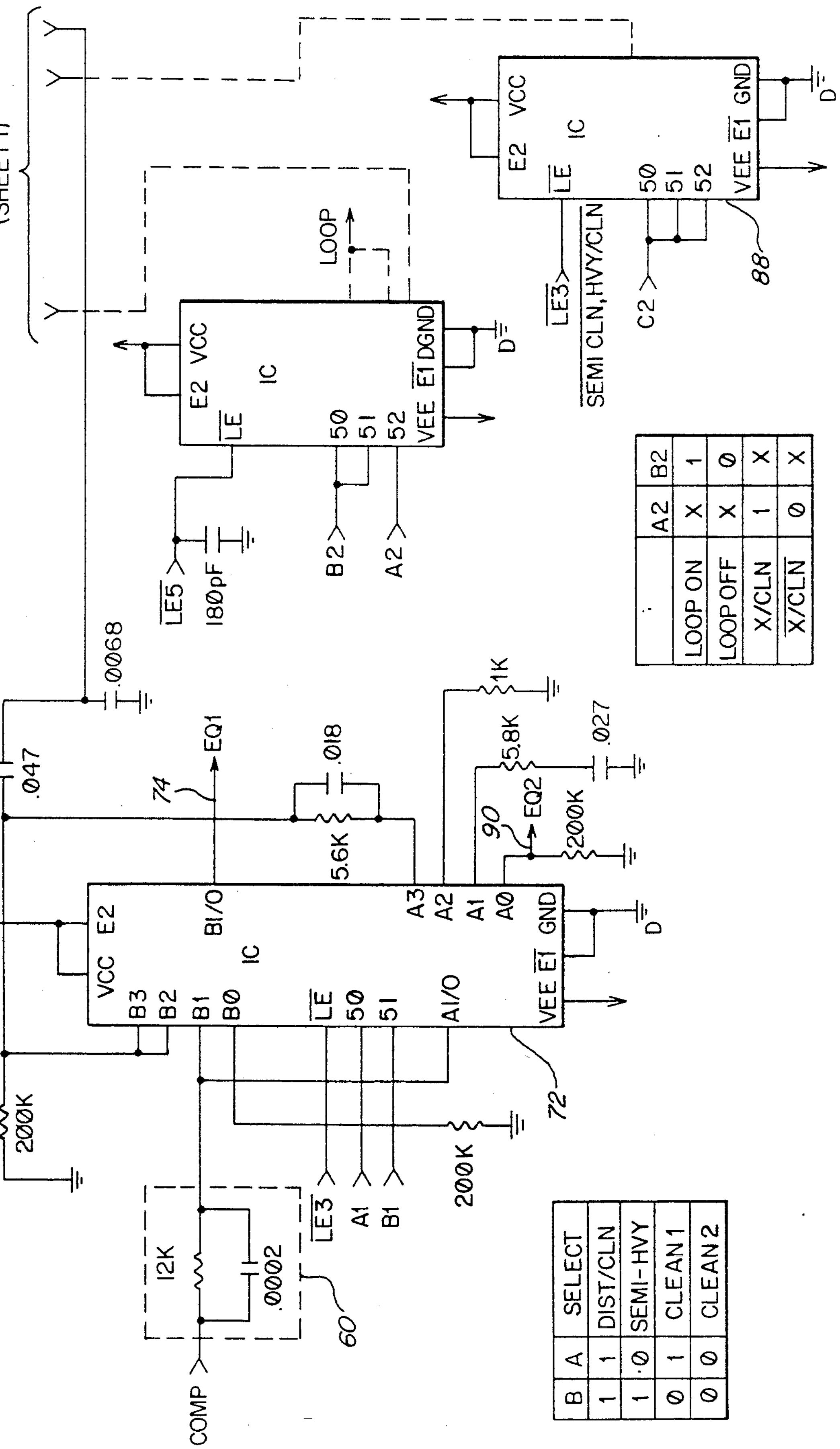


Fig. 3B (Sheet 1)



	A2	B2
LOOP ON	X	1
LOOP OFF	X	Ø
X/CLN	1	X
$\overline{\text{X/CLN}}$	Ø	X

B	A	SELECT
1	1	DIST/CLN
1	Ø	SEMI-HVY
Ø	1	CLEAN1
Ø	Ø	CLEAN2

Fig 3B (Sheet 2)

METHOD AND APPARATUS FOR PROCESSING AN AUDIO SIGNAL

BACKGROUND OF THE INVENTION

1. Field of the Invention

This invention relates to a method and apparatus for processing an audio signal and more particularly to a technique for permitting varying degrees of distortion to the audio output for a guitar or other musical instrument without causing any substantial change in the volume of the output signal and in the treble and bass content of this signal.

In playing various types of electronic musical instruments, such as for example electric guitars, a desirable acoustic effect can be achieved for some types of music by the controlled distortion of the instrument output. An undistorted signal is generally referred to as a "clean signal". Distortion is may be achieved by increasing the gain of the signal as the signal is applied to a distortion amplifier, overdriving the amplifier so that a portion of the wave-form is clipped, or by clipping in other ways, some of which may result in a reduced volume output with increasing distortion. Depending on the desired musical effect, distortion may vary from very little clipping or distortion to heavy distortion where most of the wave form is clipped.

While distortion devices have been on the market for many years, existing devices of this type have a number of limitations. First, because of the manner in which distortion is achieved through use of either a distortion amplifier or other forms of clipping, as distortion on a signal is increased, the output volume of the signal is variably increased. This variation in volume is generally undesirable and is particularly undesirable when the instrument is being played as part of a band or as a backup for a singer where it can adversely affect the balance of the group. In the past, a musician might compensate for this volume change by manually adjusting the output volume when he changed the degree of distortion. However, having to adjust two controls during a live or studio performance is difficult and it is even more difficult under these circumstances to achieve anything resembling a uniform output volume level. While devices exist which permit a particular preset volume level for a particular distortion setting, and U.S. Pat. No. 4,752,962, entitled "Audio Processing Circuit" issued Jun. 21, 1988, teaches a circuit which detects output volume and uses the detected output volume to perform compensations, there is no system currently on the market which automatically compensates for volume changes solely as a function of a selected distortion level over the full distortion range of the distortion device. Operating solely in response to the selected distortion is simpler and less expensive than the circuit shown in the patent and may eliminate noise and distortion caused by spurious volume changes or to rapid response thereto.

Another problem with distortion devices is that the distortion tends to alter the harmonic content of the output signal, and in particular to make the signal brassier or more treble. There may also be some increase in the perceived bass content as distortion increases. Such changes in harmonic content and spectral density may also adversely affect balance in a band setting and are thus also undesirable; and such variations are also very difficult to compensate for during a live performance. Again, while a preset may be possible for a single distortion setting, and U.S. Pat. No. 4,752,960 teaches some tone compensation based on a detected volume level, no technique or apparatus is currently available which automatically compensates for such spectral and harmonic variations over the full range of the distortion device solely as a function of the selected distortion level.

Finally, it is common to use a compressor circuit at the input to the distortion device. A compressor is basically an amplifier, the gain of which varies as a function of the amplitude of the signal applied thereto. Since rapid changes in this gain can cause distortion of a clean audio output, the release time, which is the time required for a change in gain to occur as a result of an increase in audio input volume level, it is generally relatively long for such compressors. Typically, the release time is in the range of one half to one second for a olean signal. However, the distortion caused by a shorter release time is not a problem for a signal which is undergoing relatively heavy distortion, and it is desirable that the release time of the input compressor be reduced when the distortion device is operating with a relatively high degree of distortion. Existing distortion devices do not alter the release time of the input compressor based on the degree of distortion. In most systems it would be difficult for the musician to achieve this desirable affect.

It is therefore desirable that an improved method and apparatus be provided for distorting an audio signal such as that generated by an electronic musical instrument which permits such distortion to be achieved without any substantial change in the volume, base content or treble content of the audio output and in particular that these capabilities be automatically achievable in response only to the musician selecting a desired level of distortion.

SUMMARY OF THE INVENTION

In accordance with the above, this invention provides a method and apparatus for processing an audio signal which includes distorting the audio signal by a controlled degree. The distortion causes variations in the volume of the output as the degree of distortion increases and also causes increases in the bass and treble content of the output with increasing distortion. The volume variation is compensated for by automatically maintaining the volume of the audio signal outputted from the distortion element substantially uniform regardless of the degree of distortion, the compensation occuring solely as a function of the selected degree of distortion. Where compression is provided prior to distortion, a capability is provided for adjusting the compressor release time so that the release time is less for higher degrees of distortion. Adjustable filtering may also be applied at the input to the distortion element, the filtering being adjustable in response to the selected degree of distortion to at least partially compensate for spectral and harmonic variations caused by the distortion. Preferably, the filtering causes decreasing emphasis on treble frequencies and bass frequencies as the degree of distortion increases. The output from the distortion element may also be adjustably filtered for evening out spectral density and harmonic content. Preferably, the output filtering increasingly rolls off at the high frequency end and at the low frequency end as the degree of distortion increases. For the preferred embodiment, output gain control, input and output filtering, and adjustment on compressor release time are all provided.

The foregoing and other objects, features and advantages of the invention will be apparent from the following more particular description of a preferred embodiment of the invention as illustrated in the accompanying drawings.

IN THE DRAWINGS

FIG. **1** is a block diagram of an idealized circuit incorporating the teachings of this invention.

FIG. **2** is a block diagram of a circuit implementing the teachings of this invention.

FIGS. 3A and 3B, when combined, form a more detail circuit diagram in semi block form of the embodiment of the invention shown in FIG. **2**.

DETAILED DESCRIPTION

Referring to FIG. **1**, a block diagram is shown of an idealized circuit **10** incorporating the teachings of this invention. In this circuit, an audio input signal is received on line **12** from an audio source which would typically be an electronic musical instrument such as an electric guitar. The audio input is passed through a compression circuit **14** which has a controlled gain which varies as a function of the input amplitude. The release time for compressor **14**, which is the time required for increasing gain transitions, is controlled by a release time control circuit **16**. Compressor **14** may for example be the compression circuit shown in the before mentioned patent or other suitable circuits for performing this function.

The output from compressor **14** is applied as a signal input to a pre-distortion filter **18**. Filter **18** is preferably formed of a number of separate filter sections **18A–18E** connected in parallel. Each filter has a particular filter characteristic which is appropriate for a particular degree of distortion, with only one of the filters **18** being connected in the circuit at any given time. Exemplary filter characteristics as a function of frequency are shown in FIG. **1** for each of the filters **18**, with filter **18A** being for a "clean" setting and filter **18E** being for the maximum or "Heavy" distortion setting. From FIG. **1**, it is seen that as the level of distortion increases, there is less and less emphasis at the treble and base ends of the spectrum, with maximum roll off at the treble and base ends occuring in filter **18E** for heavy distortion.

While in the discussion of filter **18**, and in the discussion to follow, five discrete distortion levels are assumed which, for purposes of illustration, are labeled as clean (i.e. no distortion), semi-clean, semi distorted, distorted, and heavy (i.e. heavy distortion), these five settings are for purpose of illustration only. It is to be understood that the invention may be practiced with a greater or lesser number of discrete distortion settings, and that it is also possible that the degree of distortion may be continuously variable, while still practising the teachings of the invention. Further, while exemplary filter characteristics have been illustrated for the filters **18A–18E**, it is to be understood that these characteristics are by way of illustration only and that the exact characteristics will depend on a number of factors including the distortion circuit utilized, and any preprocessing on the audio input prior to being applied to the filter **18**. Finally, while a single filter **18** is shown in FIG. **1**, the audio input may, in fact, be passed through a series of filters to achieve the desired filter characteristic,, and some of the filters may be located prior to compressor circuit **14** or after pre-distortion gain circuit **20**.

The output from the filter **18** utilized is applied as the audio signal input to pre-distortion gain circuit **20**. Again, gain circuit **20** may be formed of a plurality of separate gain circuits **20A–20E** connected in parallel with the appropriate one of the gain circuits **20** being switched into operation based on the selected distortion. The gain circuit may also be located before the filter. Some gain change may also be made before the compressor which will affect the amplitude of lower level signal more than the amplitude of higher level signal at the input of the distortion circuit.

The audio output from the selected gain circuit **20** is applied as the audio input to distortion circuit **22**. Where the distortion circuit is a distortion amplifier, depending on the gain applied to the audio signal by circuit **20**, distortion circuit **22** is overdriven by a predetermined amount resulting in the desired degree of distortion.

The output from distortion circuit **22** is connected as the audio input to post-distortion gain circuit **24**.

Post-distortion gain circuit **24** is operative to compensate for the gain in the audio signal caused by circuits **20** and **22** so that the perceived audio output from circuit **10** for a typical musical instrument input signal remains substantially uniform regardless of the distortion setting. Circuit **24** may also be formed of five separate gain circuits **24A–24E** connected in parallel, with only one of the circuits **24** being switched in depending on the distortion setting. If the gain introduced by circuits **20** and **22** increases as the distortion increases, circuits **24** provide decreasing gain as the degree of distortion increases. Thus, circuit **24A** may provide no attenuation, while circuit **24E** provides the greatest attenuation. However, because of losses in distortion amplifier **22**, the amount of attenuation required in circuit **24** for a given distortion setting is substantially less than the degree of gain required from circuit **20** to achieve the desired level of distortion. Alternatively, circuit **24** may be a post amplifier which provides greater amplification as the degree of selected distortion decreases. If a distortion circuit is utilized which results in reduced volume with increasing distortion circuits **24** might provide increasing gain for increased distortion. The objective is that circuits **24** provide appropriate compensation to maintain a substantially uniform preceived output volume.

The output from the selected circuit **24** is applied as the audio input to post distortion filter circuit **26**. Again, for purposes of illustration, the circuit **26** is shown as five separate filter circuits **26A–26E** which are connected in parallel, with only one of the filter circuits being switched into the circuit for any selected distortion level. As for the other circuits, circuit **26A** is a filter utilized with a "clean" setting while filter **26E** is for heavy distortion. Exemplary filter characteristics ar shown for each of the filter segments, with the filter characteristic for "clean" filter **26A** being substantially flat, and with the base and treble roll-offs on the filters becoming increasingly great, particularly the treble roll off, as the degree of distortion increases. The filter is thus designed to compensate for the increases in bass and treble harmonic content in the audio signal caused by distortion circuit **22**. Again, as previously indicated, the exact filter characteristic for each distortion setting will vary depending on a number of factors including the particular distortion circuit being utilized and the characteristics shown in FIG. 1 are thus for purposes of illustration only. However, for typical distortion circuits 22 the bass and treble characteristics of the filter will exhibit increasing roll-off as the distortion level increases.

The output from the selected one of filters 26 is connected as the audio input to an equalizer 28. Equalizer 28 may be of the type described in the aforementioned U.S. patent or other suitable equalizer circuit, the function of this circuit not forming part of the present invention. The output from equalizer 28 on line 30 is the audio output from circuit 10.

An input device 32 is provided which may be a control panel with push buttons 34 (as shown), a foot switch which may be stepped to the desired distortion level, a dial which may be set to the desired distortion level, or other suitable control. The output from device 32, which output may be either analog or digital, is applied as a control input to processor 36. Processor 36 may, for example, be a standard microprocessor which is programmed to control the operation of the circuit of this invention, or it may be a special purpose control circuit designed for this function. Control processor 36 recognizes the distortion level selected by the musician o the user on control device 32 and generates suitable outputs on lines 38 to control switch settings for release time control 16, pre-distortion filter 18, pre-distortion filter gain 20, post distortion gain 24, post-distortion filter 26, and equalizer 28. Typically, each of the control devices would include an electronic switch which operates in response to a digital input from the processor to switch the appropriate element into the circuit depending on the distortion level selected at device 32. Thus, if the "dist" button 34D is operated, processor 36 would generate outputs on lines 18 causing release time control 16 to operate in a reduced release time mode, and to switch pre-distortion filter 18D, pre-distortion gain circuit 20D, post-distortion circuit 24D and post-distortion circuit 26D into the circuit. With the circuit, the volume and harmonic content of audio output 30 would be perceived by a listener to be substantially uniform for a typical instrumentation over the full range of distortion settings.

While in FIG. 1, separate elements have been shown for filtering and gain control, as will be seen in the discussion to follow, these functions may in some cases be performed by common elements. Also, in some applications, commercially acceptable results may be achieved by utilizing the same gain control and/or filter for several distortion levels or settings rather than requiring a separate circuit for each setting. Thus, while FIG. 1 shows a conceptual implementation of the circuit of this invention in somewhat idealized form, FIG. 2 illustrates at a practical implementation of the circuit. The same numbers have been used for common elements in the two circuits.

Referring to FIG. 2, it is seen that the audio input on line 12 is initially applied in parallel to a pair of compressor pre filters 50A and 50B. To reduce cost, only two filters 50 are utilized in this in circuit with filter 50A being utilized if a clean or semi-clean distortion setting is selected and filter 50B being utilized if a semi-distorted, distorted or heavy distortion setting is selected. The characteristics of the filters 50A and 50B ar selected such that, in conjunction with the other pre-distortion filtering circuits in the circuit of FIG. 2, the filter characteristics for the various distortion settings are substantially as shown for the pre-distortion filter 18 in FIG. 1.

The outputs from filters 50A and 50B are applied as inputs to an electronic switch 52 which is set to one or the other of its settings in response to a suitable control signal from central processor 36 (FIG. 1). The output from switch 52 is applied to compressor circuit 14.

The output from compressor circuit 14 is fed back through an electronic switch 54 to either a circuit 56A or a circuit 56B, which circuits perform, among other things, the release time control function. Switch 54 is set to direct the feedback signal to circuit 56A if a clean or semi-clean distortion setting is selected and to direct the feedback signal to circuit 56B if a semi distorted, distorted, or heavy distortion setting is selected. The outputs from circuits 56A and 56B are fed back to an input of compressor 14.

The output from compressor 14 is also applied to pre-distortion filter gain circuits 58A, 58B and 58C and to pre-equalizer filter circuit 60. Circuit 58A is utilized if the distortion setting is either distortion or heavy, circuit 58B is utilized for a semi-distortion setting and circuit 58C is used for a semi-clean distortion setting. The characteristics of the filters portion of circuits 58 and filter 60 are selected such that, in conjunction with the characteristics of filters 50, and any other pre-distortion filters in the system, the combined pre-distortion characteristic for the filters through which an audio signal passes for a given distortion setting are substantially as shown for the pre-distortion filters 18 in FIG. 1.

The outputs from circuits 58 are applied as the inputs to an electronic switch 62 which is set to the appropriate one of the circuit outputs under control of processor 36 in response to the selected distortion setting. The output from switch 62 is applied as an input to electronic switch 64, the outputs from which are applied as the inputs to pre-distortion gain and filter circuits 66A and 66B. Processor 36 causes switch 64 to be set to circuit 66A if the distortion setting is semi-clean, semi-distorted, or heavy and to circuit 66B if the distortion setting is "distorted." The circuits 66 in conjunction with the circuits 58 provide the gain control required to achieve the desired distortion level and also perform additional pre-distortion filtering to assist in achieving the desired pre-distortion filter characteristics shown for the filters 18 in FIG. 1.

The outputs from the selected circuit 66 are applied as the audio inputs to distortion circuit 22. The outputs from distortion circuit 22 are applied to an electronic switch 68 controlled from processor 36. Switch 68 applies the distortion output to either post distortion gain and filter circuit 70A or 70B. Switch 68 is set to circuit 70A if a semi distorted, distorted, or heavy distorted setting is selected and to circuit 70B if a semi-clean distortion level is selected. Circuits 70 perform the functions of the post distortion gain circuit 24 and post distortion filter circuit 26 shown in FIG. 1 to compensate for the increased gain resulting from the distortion operation and to perform treble and base filtering to compensate for increases in harmonic content in these ranges as a result of the distortion operation. Circuit 70A can perform the desired function for the three distortion levels only if there is little difference in the volume and harmonic input to the circuit for these settings. It is preferable that additional circuits 70 be provided.

The outputs from circuit 70 are applied as one input to electronic switch 72, the other input to this switch being the output from filter **60**. Switch **72** is set to the output from filter **60** if a "clean" setting has been selected and is otherwise set to receive the outputs from a circuit **70**. The output from switch **72** is applied through equalizer circuit **28** to audio output line **30**.

FIGS. 3A and 3B are more detailed diagrams of the circuit shown in FIG. 2. To the extent possible, common reference numerals have been utilized in the various figures. It will, however, be noted that some minor differences in detail exist between the general circuit diagram of FIG. 2 and the more detailed circuit diagram of FIGS. 3A and 3B. Further, the circuit diagrams of FIGS. 3A and 3B contain a number of elements which perform various functions not directly related to the current invention. These elements will not be mentioned in the discussion to follow which is limited to a discussion of the elements utilized in performing the functions of this invention.

Referring first to FIG. 3A, it is seen that the input signal **12** is passed through a number of components which are involved in performing an equalization function which is not part of the present invention to the compressor pre filters **50A** and **50B**. Switch **52**, which is part of an electronically controlled switch chip **80**, passes the output from the appropriate one of the filters **50** to one input of compressor amplifier **14**. The feedback circuit for the compressor amplifier includes several parallel paths to ground, one of which is reached through switch **54**. When switch **54** is in position **1**, the path **56B** is added into the circuit resulting in a reduction in the release time for the compressor circuit. The compressor circuit is controlled by an FET **82**, the output from which is applied through varies components as a control input to amplifier **14**. Switch **84**, which is also part of the switch circuit **80**, is in the position shown for the distortion or heavy distortion settings, and is transferred to a short circuit mode to short out the resistor and capacitor in other distortion modes to reduce noise in the circuit at low signal level.

The output from the compressor **14** is applied at two points in the circuit of FIG. 3B, namely as an input to circuits **58A 58B** and **58C**, and is an input to filter **60**. The circuits **58** perform the functions previously indicated and the outputs from these circuits are applied to a switch circuit **86** which performs the functions of the switches **62** and **64** in FIG. 2. Switch circuit **86** applies outputs to pre-distortion gain and filter circuits **66A** and **66B** which are the inputs to the distortion circuit **22**. The distortion circuit **22** functions in well known manner to perform controlled clipping on the input signal applied thereto depending on the gain level of its input signal.

The output from distortion circuit **22** is applied to post distortion gain and filter circuits **70A** and **70B**, with the outputs from these circuits being applied as the inputs to switch **68**. Switch **68** is part of a switch circuit **88** controlled from processor **36**.

The output from switch **68** is applied as one of the inputs to switch circuit **72**, a second input to this switch circuit being the output from filter circuit **60**. Line **74** leading to equalizer **28** is one of the outputs from switch **72**. A second output from this switch is line **90** leading to a second equalizer circuit.

While specific components are shown in FIG. s 3A and 3B for performing the various functions, it is to be understood that these components and values, while utilized for a preferred embodiment of the invention, are for purposes of illustration only, and that these components and values might be different for an input having different characteristics, for a different distortion circuit, for a different desired output, or for other variations which might come within the contemplation of this invention.

While for the preferred embodiment, the "clean" output from filter **60** has bypassed the gain and filter circuits **66**, distortion circuit **22** and post distortion gain and filter circuits **70**, as illustrated by FIG. **1**, this is not a limitation on the invention. However, the "clean" setting is the only one which can bypass these elements. Further, while three specific functions have been automatically compensated for in connection with a distortion circuit, it is apparent that other variations resulting from the use of a distortion circuit might also be automatically compensated should a user so desire. Thus, while the invention has been particularly shown and described above with reference to a generalized and a preferred embodiment, the foregoing and other changes in form in detail may be made therein by one skilled in the art without departing from the spirit and scope of the invention.

What is claimed is:

1. A circuit for processing an audio signal comprising:

means for distorting said audio signal;

means for indicating a desired degree of distortion having a single user input element;

means responsive to said indicating means for controlling the degree of distortion caused by said distortion means, said distortion means having an output volume which varies in a predetermined way as the degree of distortion increases; and

output means responsive to said indicating means for automatically, and without further user input, maintaining the volume of the audio signal output from said circuit substantially at a desired level regardless of the degree of distortion caused by said distortion means.

2. A circuit as claimed in claim **1** wherein said distortion means has a predetermined number of distortion states, each providing a different degree of distortion; and wherein said output means comprises a predetermined number of fixed gain elements, each one of said gain elements corresponding to one of said distortion states, and having a gain value which causes the overall gain of the distortion means and output means, combined, to remain substantially constant for each of said states.

3. A circuit as claimed in claim **2** wherein said distortion means includes a distortion amplifier; and

wherein said means for controlling includes means responsive to said indicating means for controlling the gain of the audio signal inputted to said distortion means.

4. A circuit as claimed in claim **1** including compressor means through which said audio signal is passed before being applied to said distortion means, said compressor means having a release time; and

means responsive to said indicating means for adjusting said compressor means release time, the release time being less for higher degrees of distortion.

5. A circuit as claimed in claim **1** including adjustable filter means through which said audio signal is passed before being applied to said distortion means, said filter means being adjusted in response to said indicating means to at least partially compensate for spectral and harmonic variations caused by said distortion means which vary with changes in the degree of distortion.

6. A circuit as claimed in claim 5 wherein said filter means has decreasing emphasis at high frequencies and low frequencies as the degree of distortion increases.

7. A circuit as claimed in claim 1 wherein said output means includes adjustable output filter means responsive to said indicating means for evening out spectral density and harmonic content of said outputted audio signal.

8. A circuit as claimed in claim 7 wherein said output filter means increasingly rolls off at the high frequency end and at the low frequency end as the degree of distortion of said distortion means is increased.

9. A circuit as claimed in claim 7 including compressor means through which said audio signal is passed before being applied to said distortion means, said compressor means having a release time; and

means responsive to said indicating means for adjusting said compressor means release time, the release time being less for higher degrees of distortion.

10. A circuit as claimed in claim 1 wherein said output means includes adjustable output filter means responsive to said indicating means for evening out spectral density and harmonic content of said outputted audio signal.

11. A circuit as claimed in claim 10 wherein said output filter means increasingly rolls off at the high frequency end and at the low frequency end as the degree of distortion of said distortion means is increased.

12. A circuit for processing an audio signal comprising:

means for distorting said audio signal;

means for indicating a desired degree of distortion;

means responsive to said indicating means for controlling the degree of distortion caused by said distortion means, said distortion means causing spectral and harmonic variations in said audio signal which vary with changes in the degree of distortion;

output means responsive to said indicating means for automatically, and without further user input, maintaining the volume of the audio signal output from said circuit substantially at a desired level regardless of the degree of distortion caused by said distortion means; and

adjustable filter means through which said audio signal is passed before being applied in said distortion means, said filter means being adjusted in response to said indicating means to at least partially compensate for said spectral and harmonic variations.

13. A circuit as claimed in claim 12 wherein said filter means has decreasing emphasis at high frequencies and low frequencies as the degree of distortion increases.

14. A circuit as claimed in claim 12 including adjustable output filter means at the output from said distortion means, said output means being adjusted in response to said indicating means to even out spectral density and harmonic content of said outputted audio signal.

15. A circuit as claimed in claim 14 wherein said output filter means increasingly rolls off at the high frequency end and at the low frequency end as the degree of distortion of said distortion means is increased.

16. A circuit for processing an audio signal comprising:

means for distorting said audio signal;

means for indicating a desired degree of distortion;

means responsive to said indicating means for controlling the degree of distortion caused by said distortion means, said distortion means causing spectral and harmonic variations in said audio signal which vary with changes in the degree of distortion;

output means responsive to said indicating means for automatically, and without further user input, maintaining the volume of the audio signal output from said circuit substantially at a desired level regardless of the degree of distortion caused by said distortion means; and

output filter means at the output from said distortion means, said output filter means being adjustable in response to said indicating means to even out the spectral density and harmonic content of said outputted audio signal.

17. A circuit as claimed in claim 16 wherein said output filter means increasingly rolls off at the high frequency end and at the low frequency end as the degree of distortion of said distortion means is increased.

18. A circuit for processing an audio signal comprising:

means for distorting said audio signal;

means for indicating a desired degree of distortion;

means responsive to said indicating means for controlling the degree of distortion caused by said distortion means;

compressor means through which said audio signal is passed before being applied to said distortion means, said compressor means having a release time; and

means responsive to said indicating means for adjusting said compressor means release time, the release time being less for higher degrees of distortion.

19. A method for processing an audio signal comprising the steps of:

providing a predetermined degree of distortion to said audio signal, said distortion step varying the volume of said audio signal in a predetermined way as the degree of distortion increases; and

automatically maintaining the volume of the audio signal outputted from said distortion step substantially at a desired level regardless of the degree of distortion during said distortion step, said maintaining step being performed in response to a selected distortion level input from a user.

20. A method as claimed in claim 19 including the step performed before said distortion step of compressing said audio signal, and adjusting the release time of the audio signal during said compression step so that said release time is less for higher degrees of distortion.

21. A method as claimed in claim 19 including the step performed before said distorting step of adjustably filtering said audio signal to at least partially compensate for spectral and harmonic variations caused by the distortion step which variations vary with changes in the degree of distortion, said adjustable filtering being performed in response to said selected distortion level input.

22. A method as claimed in claim 21 wherein said filtering step includes the step of decreasing emphasis

on high frequencies and low frequencies as the degree of distortion increases.

23. A method as claimed in claim 22 including the step of adjustably filtering the output from said distortion step to even out spectral density and harmonic content of said outputted audio signal, said output filtering step being performed in response to said selected distortion level input.

24. A method as claimed in claim 23 wherein said output filtering step includes the step of increasingly rolling off at the high frequency and at the low frequency end as the degree of distortion of said distortion means is increased.

25. A method as claimed in claim 23 including the step performed before said distortion step of compressing said audio signal, and adjusting the release time of the audio signal during said compression step so that said release time is less for higher degrees of distortion.

26. A method as claimed in claim 19 including the step of adjustably filtering the output from said distortion step to even out spectral density and harmonic content of said outputted audio signal, said output filtering step being performed in response to said selected distortion level input.

27. A method as claimed in claim 26 wherein said output filtering step includes the step of increasingly rolling off at the high frequency end at the low frequency end as the degree of distortion of said distortion means is increased.

28. A method for processing an audio signal comprising the steps of:

providing a predetermined degree of distortion to said audio signal in response to a selected distortion level input from a user, said distortion step causing spectral and harmonic variations in said audio signal which vary with changes in the degree of distortion;

adjustably filtering the audio signal inputted to said distortion step in response to said selected distortion level input to at least partially compensate for said spectral and harmonic variations; and

automatically maintaining the volume of the audio signal output from said distortion step substantially at a desired level regardless of the degree of distortion during said distortion step, said maintaining step being performed in response to a selected distortion level input from a user.

29. A method as claimed in claim 28 wherein said filtering step includes the step of decreasing emphasis on high frequencies and low frequencies as the degree of distortion increases.

30. A method as claimed in claim 28 including the step of adjustably filtering the output from said distortion step in response to said selected distortion level input to even out the spectral density and harmonic content of said outputted audio signal.

31. A method as claimed in claim 30 wherein said output filtering step includes the step of increasingly rolling off at the high frequency and at the low frequency end as the degree of distortion of said distortion means is increased.

32. A method of processing an audio signal comprising the steps of:

providing a predetermined degree of distortion to said audio signal in response to a selected distortion level input from a user, said distortion step causing spectral and harmonic variations in said audio signal which vary with changes in the degree of distortion;

adjustably filtering the output from said distortion step in response to said selected distortion level input to even out spectral density and harmonic content of said outputted audio signal; and

automatically maintaining the volume of the audio signal output from said distortion step substantially at a desired level regardless of the degree of distortion during said distortion step, said maintaining step being performed in response to a selected distortion level input from a user.

33. A method as claimed in claim 32 wherein said output filtering step includes the step of increasingly rolling off at the high frequency and at the low frequency end as the degree of distortion of said distortion means is increased.

34. A circuit for processing an audio signal comprising:

means for distorting said audio signal;

means for indicating a desired degree of distortion;

means responsive to said indicating means for controlling the degree of distortion caused by said distortion means, said distortion means having an output volume which varies in a predetermined way as the degree of distortion increases;

output means responsive to said indicating means maintaining the volume of the audio signal output from said circuit substantially at a desired level regardless of the degree of distortion caused by said distortion means;

compressor means through which said audio signal is passed before being applied to said distortion means, said compressor means having a release time; and

means responsive to said indicating means for adjusting said compressor means release time, the release time being less for higher degrees of distortion.

35. A circuit for processing an audio signal comprising:

means for distorting said audio signal;

means for indicating a desired degree of distortion;

means responsive to said indicating means for controlling the degree of distortion caused by said distortion means, said distortion means having an output volume which varies in a predetermined way as the degree of distortion increases;

output means responsive to said indicating means maintaining the volume of the audio signal output from said circuit substantially at a desired level regardless of the degree of distortion caused by said distortion means;

said output means further including adjustable filter means responsive to said indicating means for evening out spectral density and harmonic content of said outputted audio signal;

compressor means through which said audio signal is passed before being applied to said distortion means, said compressor means having a release time; and

means responsive to said indicating means for adjusting said compressor means release time, the release time being less for higher degrees of distortion.

36. A method for processing an audio signal comprising the steps of:

compressing said audio signal, and adjusting the release time of the audio signal during said compression step so that said release time is less for higher degrees of distortion;

providing a predetermined degree of distortion to said audio signal, said distortion step varying the volume of said audio signal in a predetermined way as the degree of distortion increases; and

automatically maintaining the volume of the audio signal output from said distortion step substantially at a desired level regardless of the degree of distortion during said distortion step, said maintaining step being performed in response to a selected distortion level input from a user.

37. A method for processing an audio signal comprising the steps of:

adjustably filtering audio signal before said distorting step to at least partially compensate for spectral and harmonic variations caused by the distortion step which variations vary with changes in the degree of distortion, said adjustable filtering being performed in response to said selected distortion level input;

said filtering step further including the step of decreasing emphasis on high frequencies and low frequencies as the degree of distortion increases;

compressing said audio signal, and adjusting the release time of the audio signal during said compression step so that said release time is less for higher degrees of distortion;

providing a predetermined degree of distortion to said audio signal, said distortion steps varying the volume of said audio signal in a predetermined way as the degree of distortion increases;

adjustably filtering the audio signal inputted to said distortion step to even out spectral density and harmonic content of said outputted audio signal, said output filtering step being performed in response to said selected distortion level input;

automatically maintaining the volume of the audio signal output from said distortion step substantially at a desired level regardless of the degree of distortion during said distortion step, said maintaining step being performed in response to a selected distortion level input from a user.

38. A circuit for processing an audio signal comprising:

only one user input device for receiving user input specifying distortion level;

a distortion element, responsive to said user input device, for providing a distorted audio signal having said distortion level specified, and having a gain which varies in a known way corresponding with said distortion level; and

and output element, responsive to said user input device, for providing said distorted audio signal amplified or attenuated by a gain corresponding with said distortion element and output element is fixed for all distortion levels.

39. A circuit as claimed in claim 38, further comprising:

a compressor, responsive to said user input device, for compressing said audio signal prior to said distortion element, and having a release time of said compression, said release time corresponding with said distortion level specified such that said release time is lower at higher distortion levels.

40. A circuit as claimed in claim 38, further comprising:

a prefilter, responsive to said user input device, for filtering said audio signal prior to said distortion element having settings to at least partially compensate for spectral and harmonic variations caused by said distortion element, and operating over at least upper, middle and lower bands, said settings corresponding with said distortion level specified, such that said upper and lower bands are emphasized less at higher distortion levels relative to said middle band.

41. A circuit as claimed in claim 38 further comprising:

an equalizer, responsive to said user input device, for filtering said audio signal after said distortion element, reducing spectral and harmonic variations caused by said distortion element, and operating over at least upper, middle and lower bands said reduction of spectral and harmonic variations corresponding with said distortion level specified such that said upper and lower bands are emphasized less at higher distortion levels relative to said middle band.

42. A circuit for processing an audio signal comprising:

means for distorting said audio signal;

means for indicating a desired degree of distortion including only a single input device;

means responsive to said indicating means for controlling the degree of distortion caused by said distortion means, said distortion means having an output volume which varies in a predetermined way as the degree of distortion increases; and

output means responsive to said single input device for automatically maintaining the volume of the audio signal output from said circuit substantially at a desired level regardless of the degree of distortion caused by said distortion means.

43. A circuit as claimed in claim 42 wherein said distortion means includes only a predetermined number of distortion states, each providing a different degree of distortion; and wherein said output means includes only a predetermined number of fixed gain elements, each one of said elements corresponding to one of said distortion states and having a gain value which causes the overall gain of the distortion means and output means, combined, to remain substantially constant for each of said states.

44. A circuit for processing an audio signal comprising:

a user input device for receiving user input specifying distortion level;

a distortion element, responsive to said user input device, for providing a distorted audio signal having said distortion level specified, and having a gain which varies in a known way corresponding with said distortion level; and

and output element, responsive to said user input device, for providing said distorted audio signal amplified or attenuated by a gain corresponding with said distortion level specified, such that a total gain through said distortion element and output element is fixed for all distortion levels.

* * * * *

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Poletti

(54) NONLINEAR PROCESSOR FOR AUDIO SIGNALS

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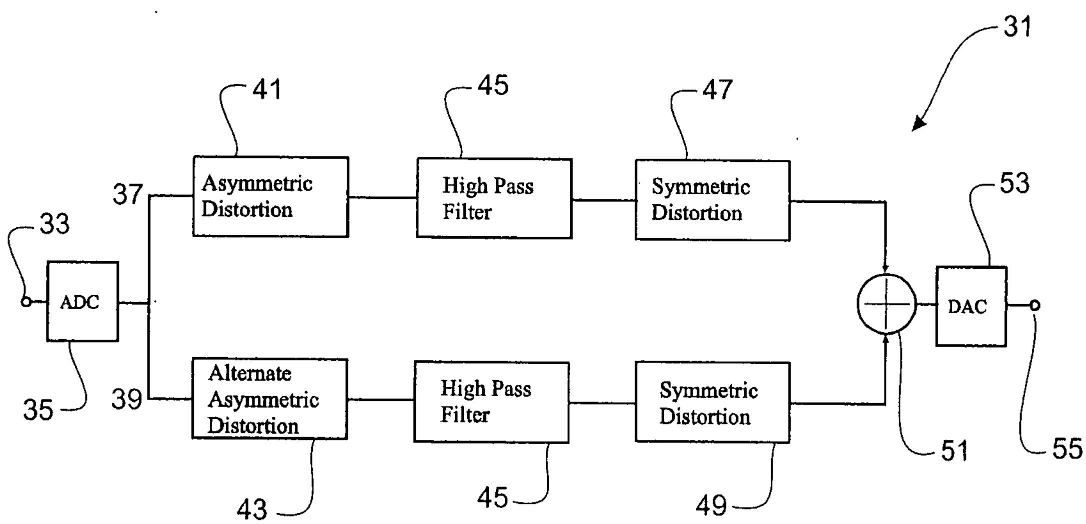
(57) ABSTRACT

A nonlinear processor for distorting audio signals having an input stage (**15**) that is arranged to split an audio input signal (**13**) into two signal paths and then a pair of asymmetric distortion stages (**17, 19**), one in each signal path, with non-equal negative and positive saturation limits, so as to produce opposite polarity mean signal levels at their outputs in each signal path, and which produce a smooth transition from linear to nonlinear behaviour. Following the asymmetric distortion stages (**17, 19**) is a pair of AC-coupled symmetric distortion stages (**21, 23**), one in each signal path, and an output stage (**25**) that is arranged to add the two nonlinearly distorted signals from the symmetric distortion stages to generate an audio output signal (**27**) that demonstrates a smooth transition from linear behaviour to the production of crossover-like artifacts.



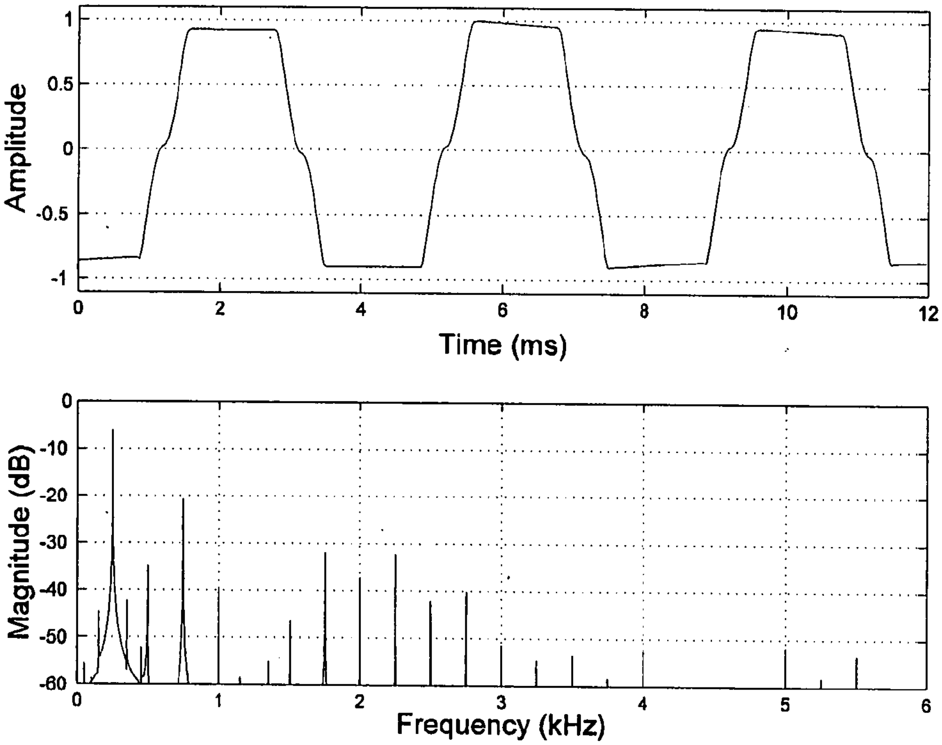


FIGURE 1

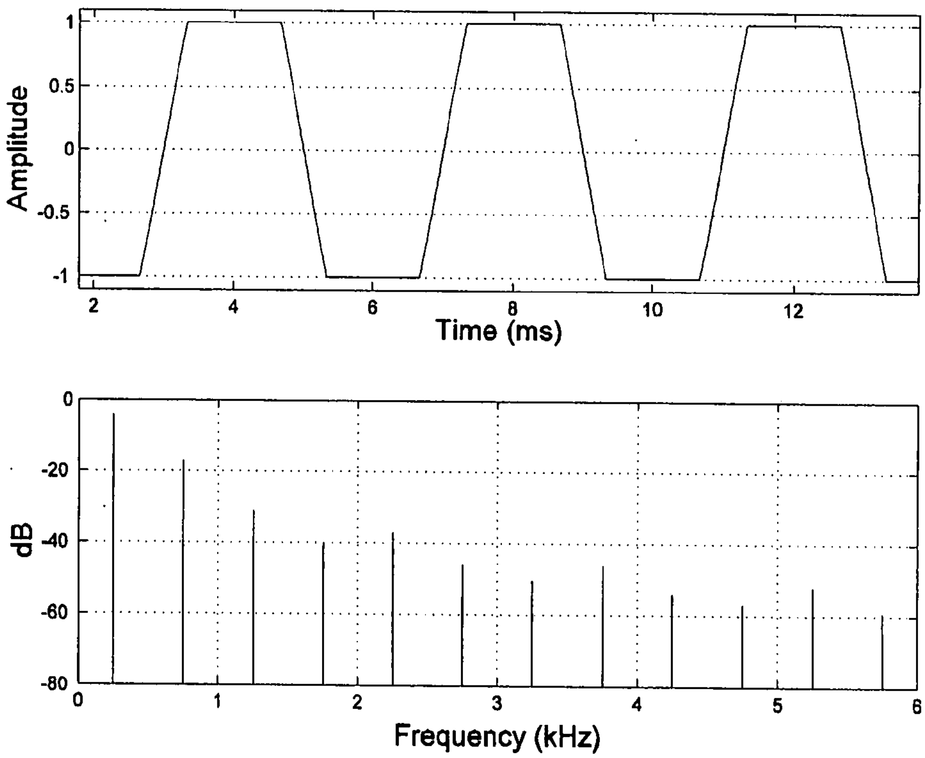


FIGURE 2

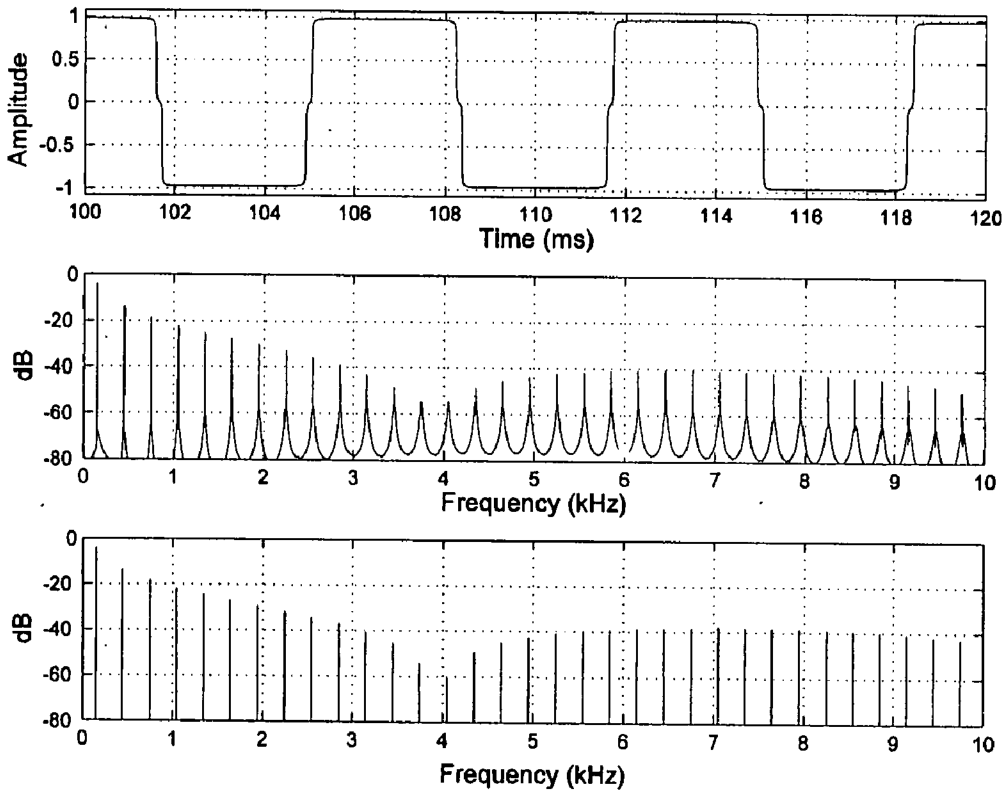


FIGURE 3

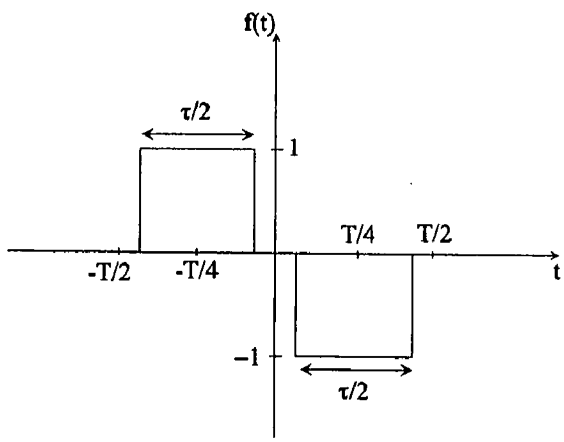


FIGURE 4

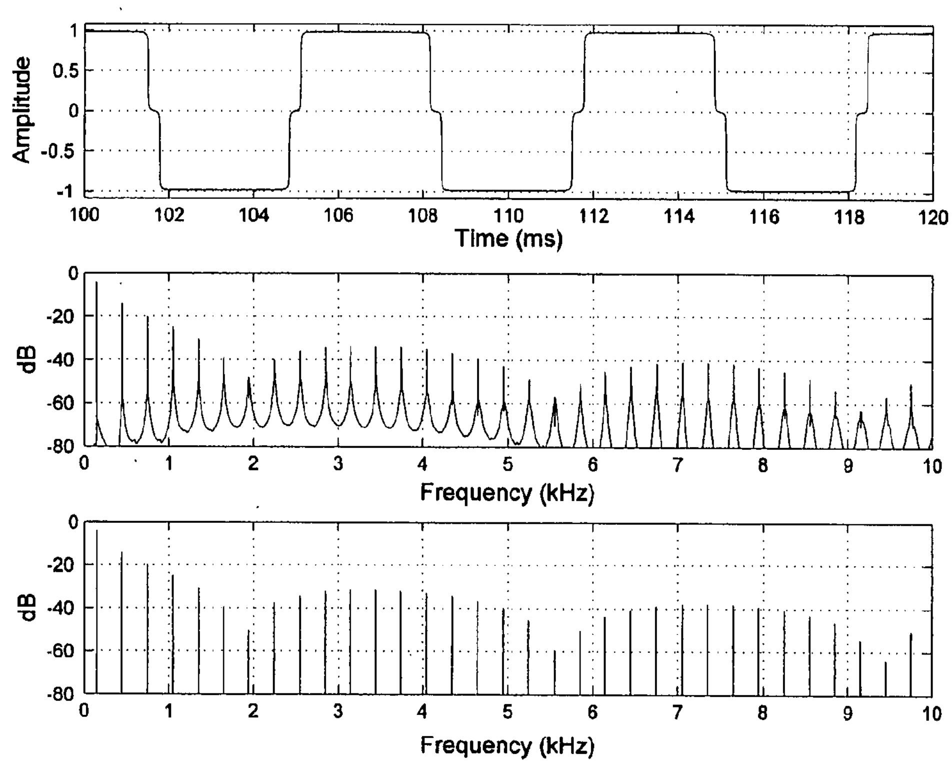


FIGURE 5

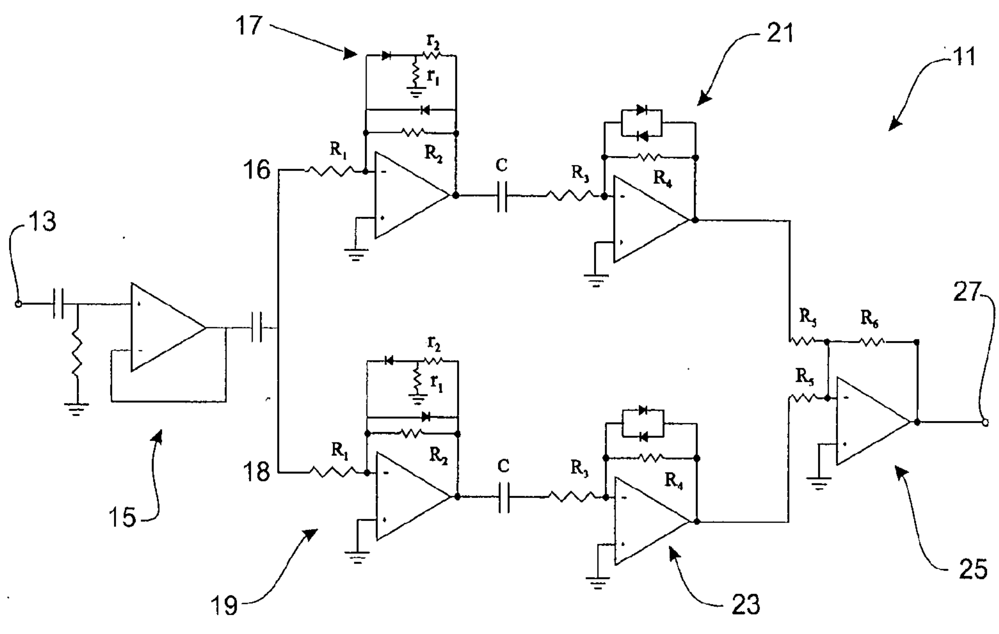


FIGURE 6

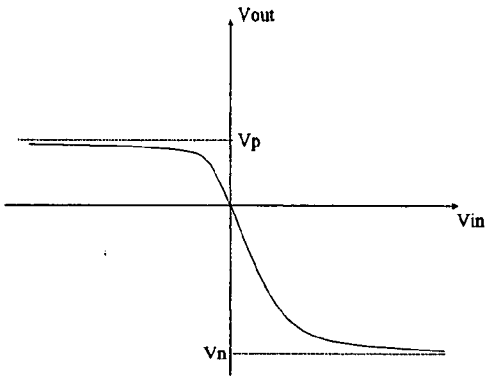


FIGURE 7a

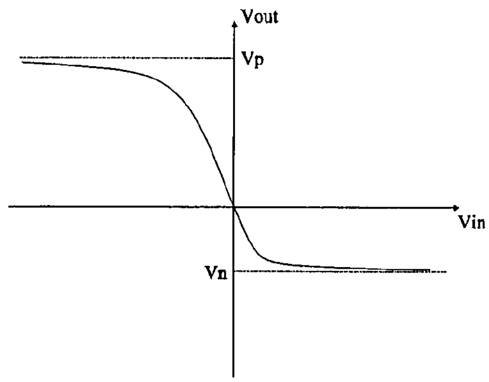


FIGURE 7b

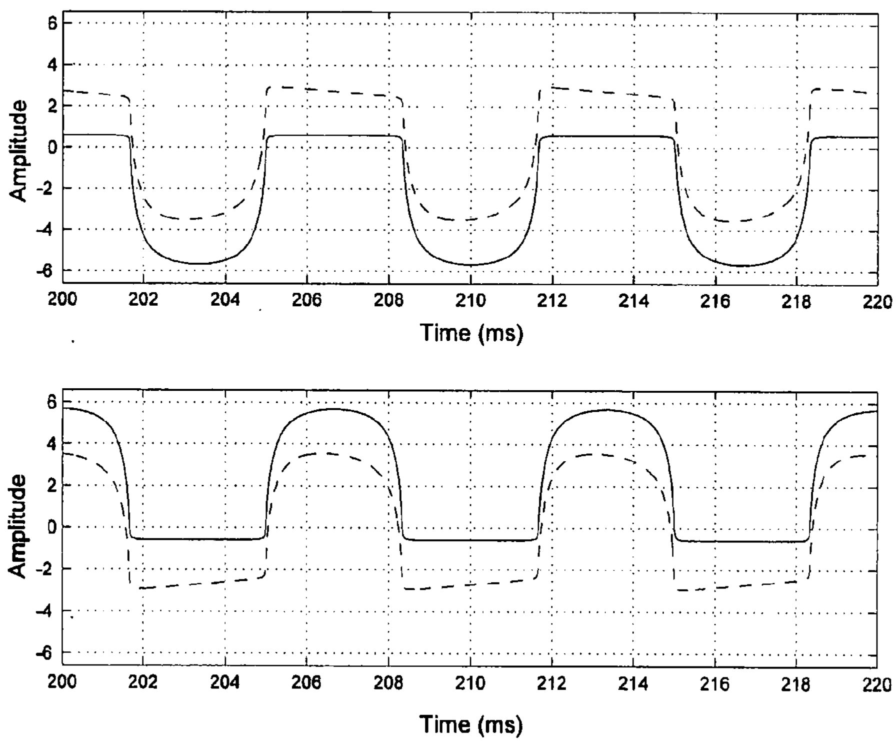


FIGURE 8

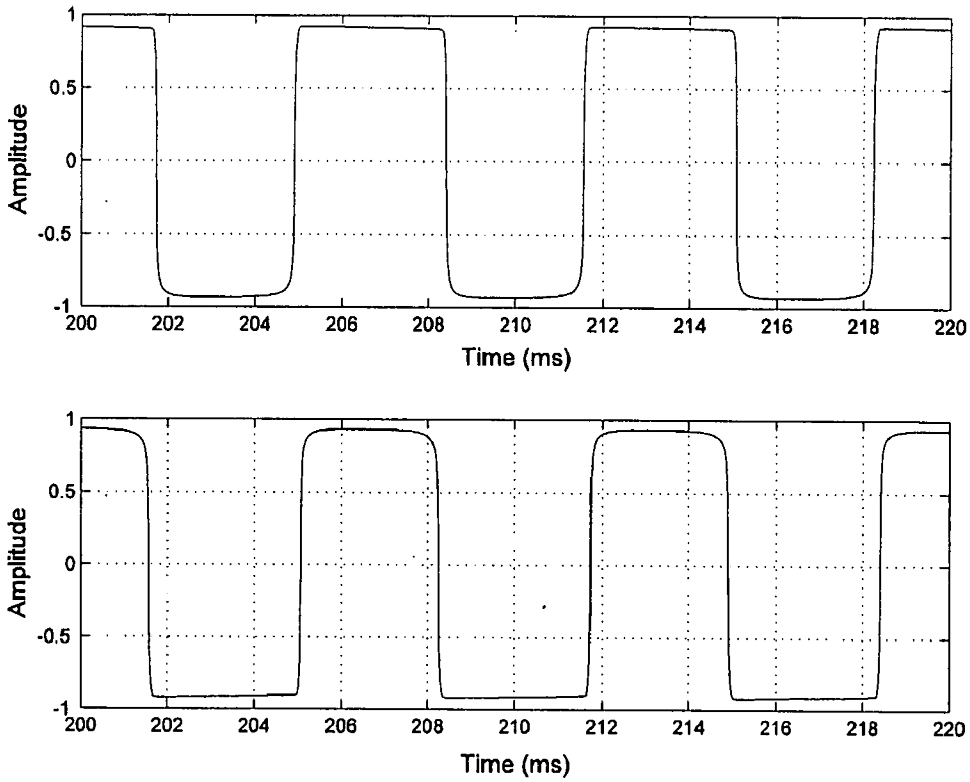


FIGURE 9

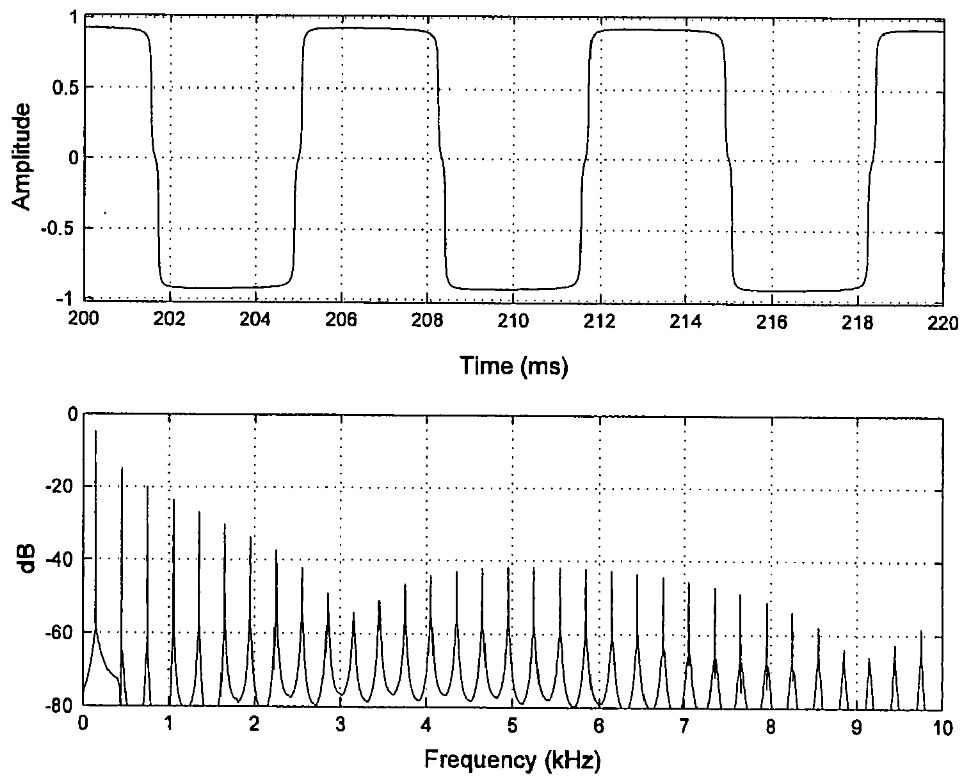


FIGURE 10

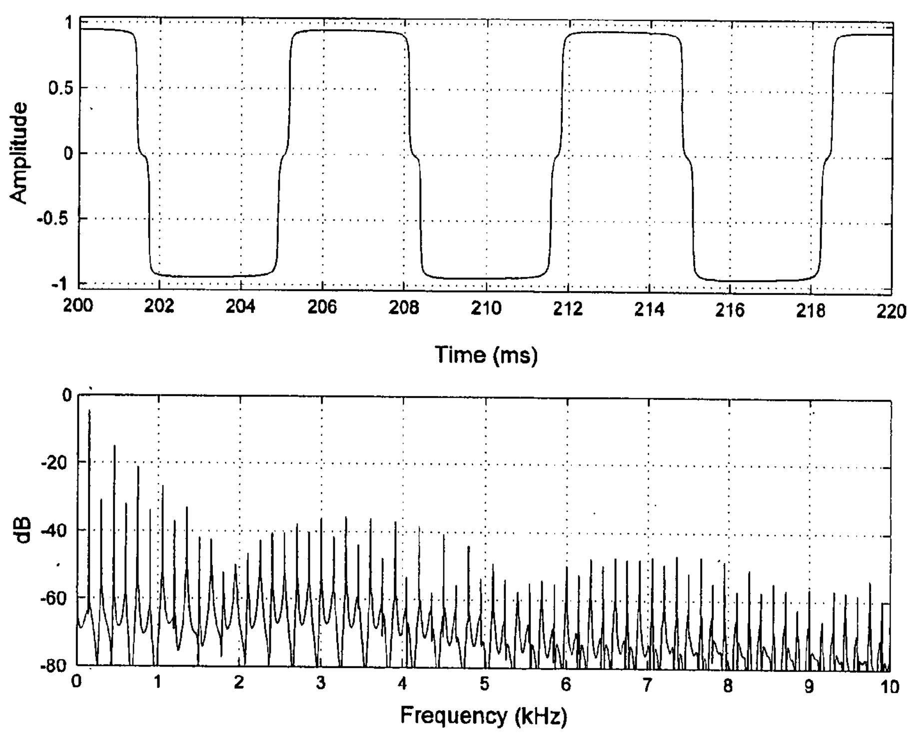


FIGURE 11

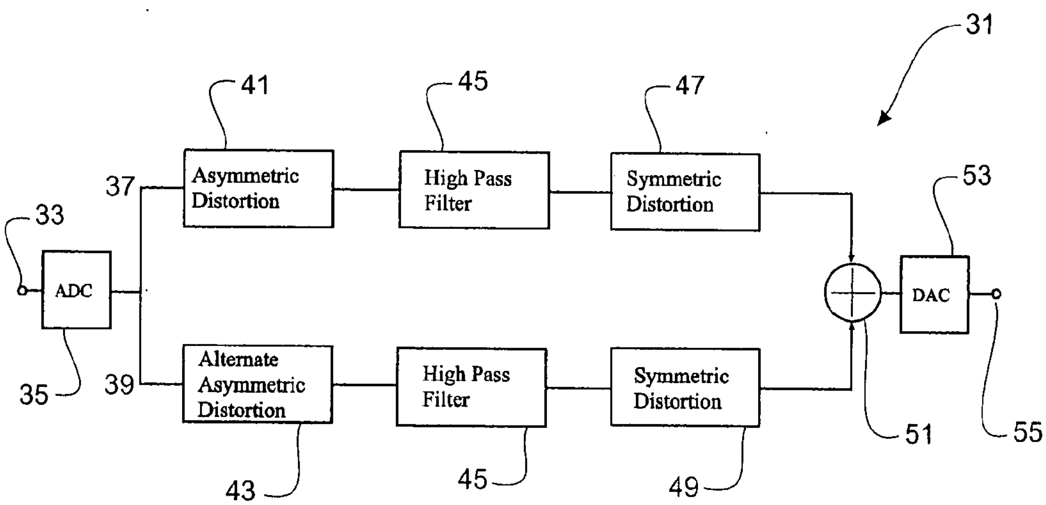


FIGURE 12

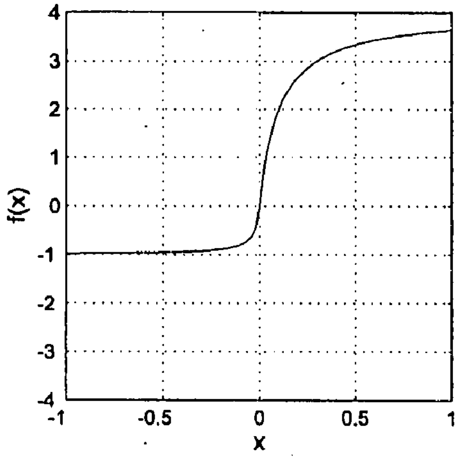


FIGURE 13a

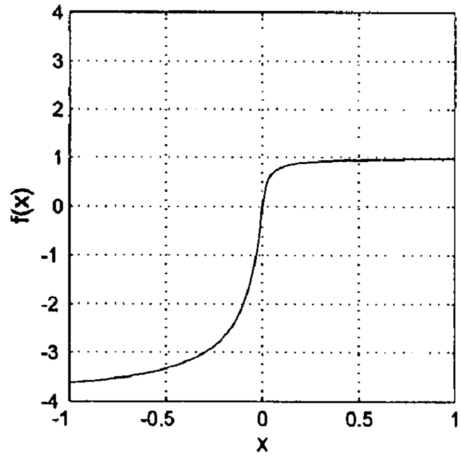


FIGURE 13b

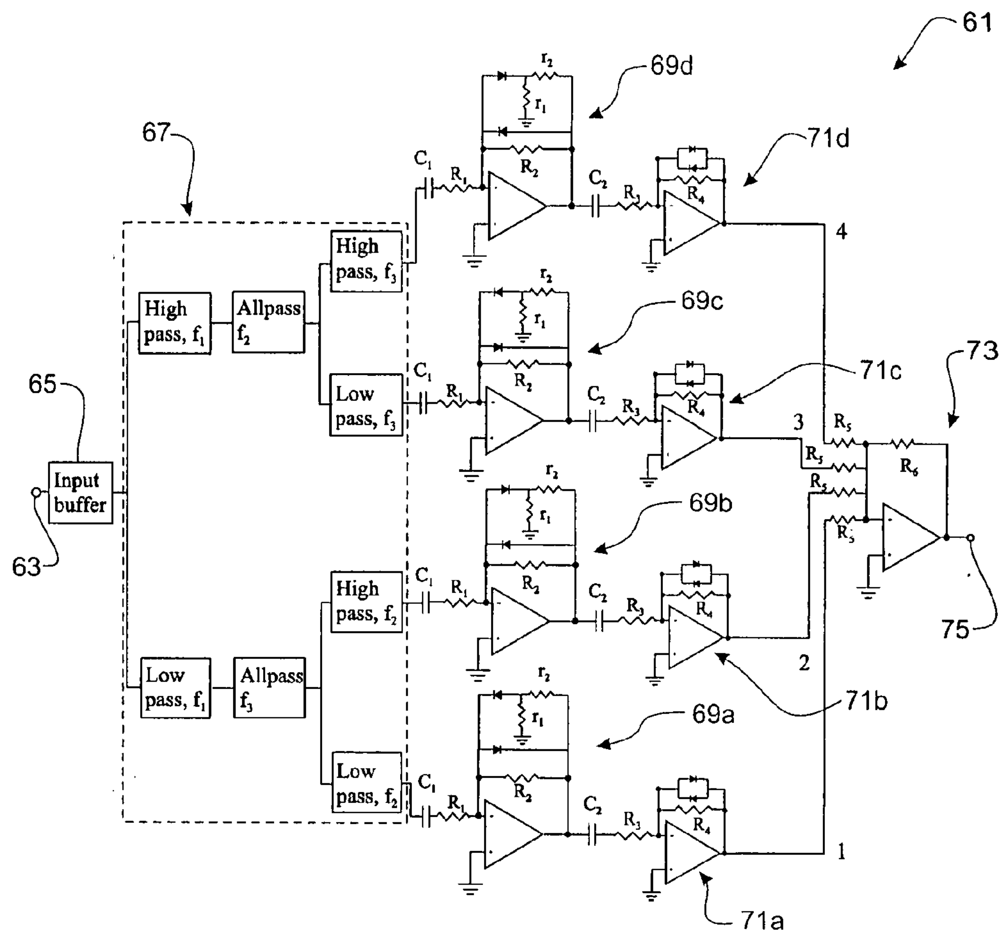


FIGURE 14

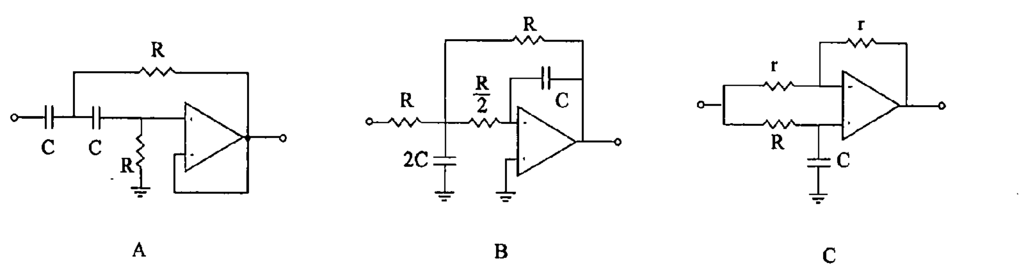


FIGURE 15a FIGURE 15b FIGURE 15c

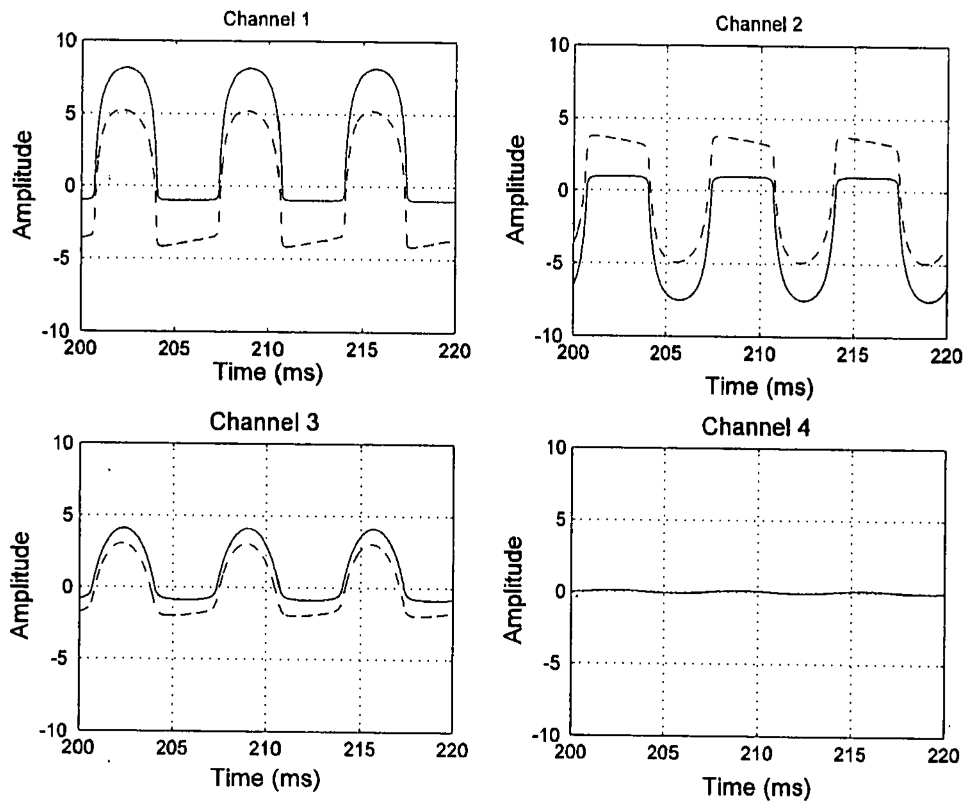


FIGURE 16

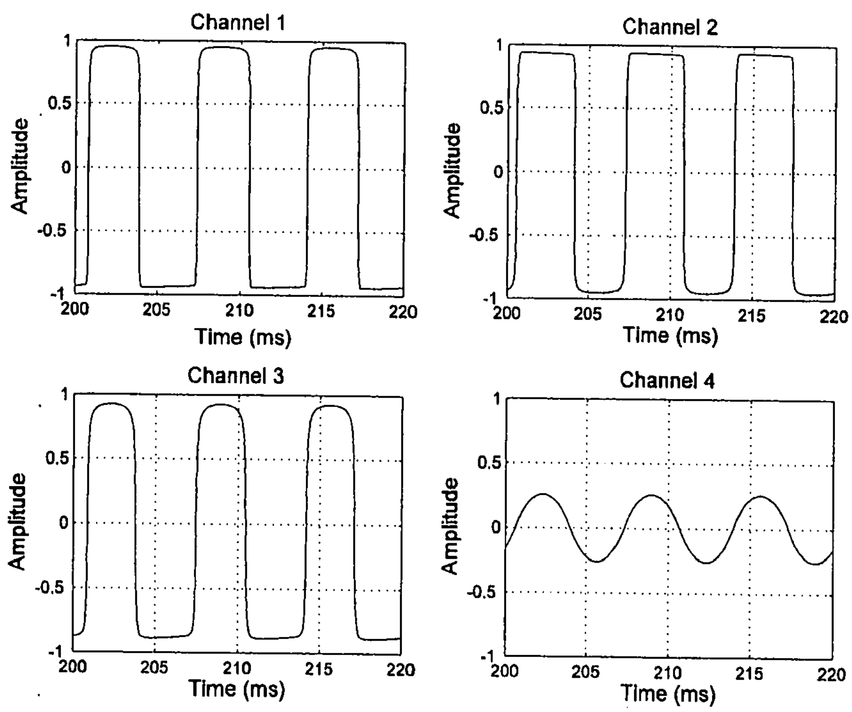


FIGURE 17

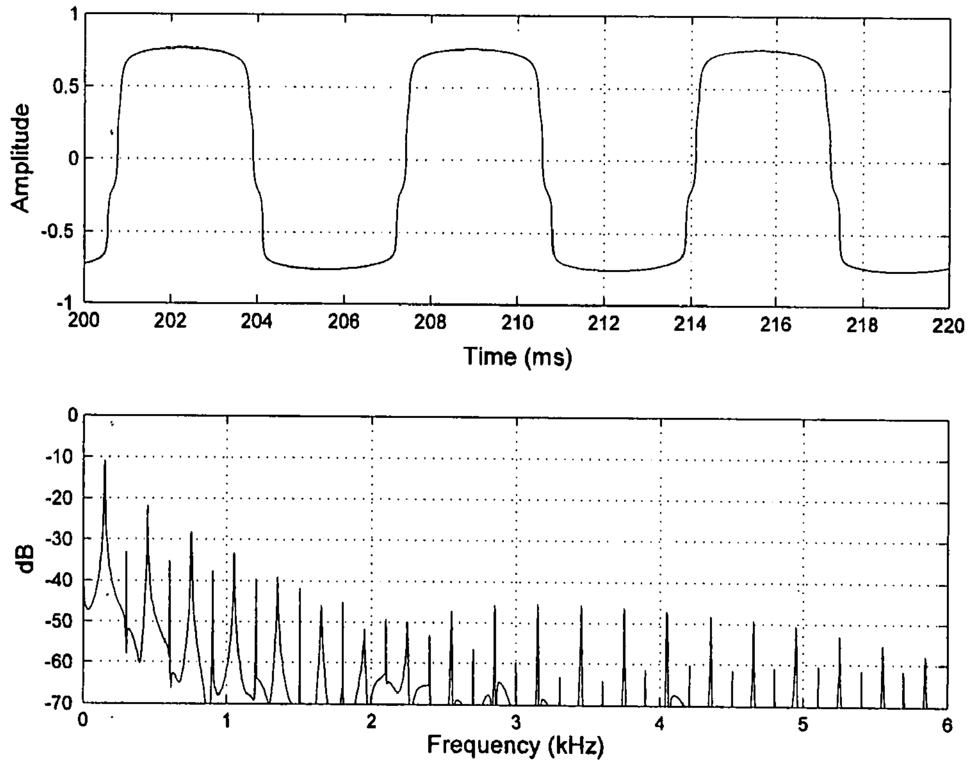


FIGURE 18

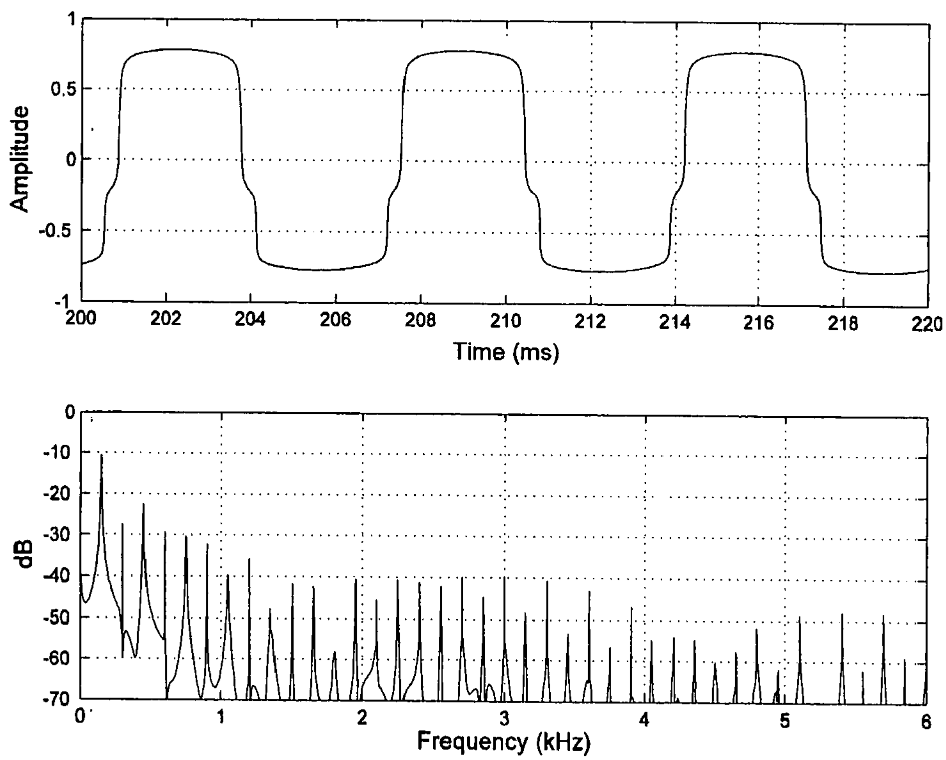


FIGURE 19

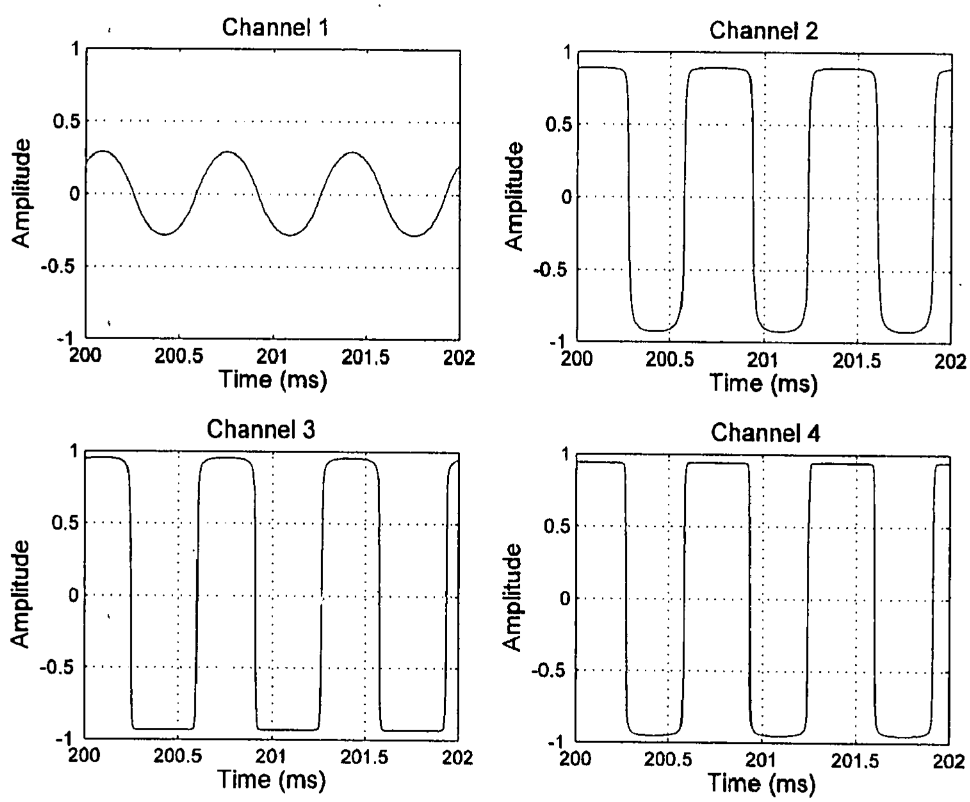


FIGURE 20

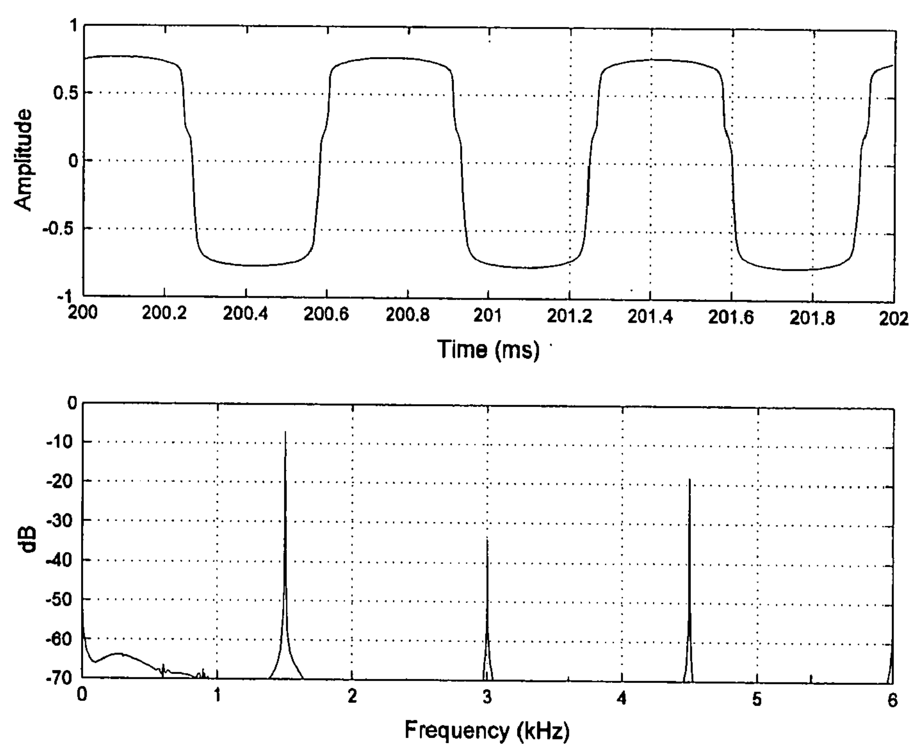


FIGURE 21

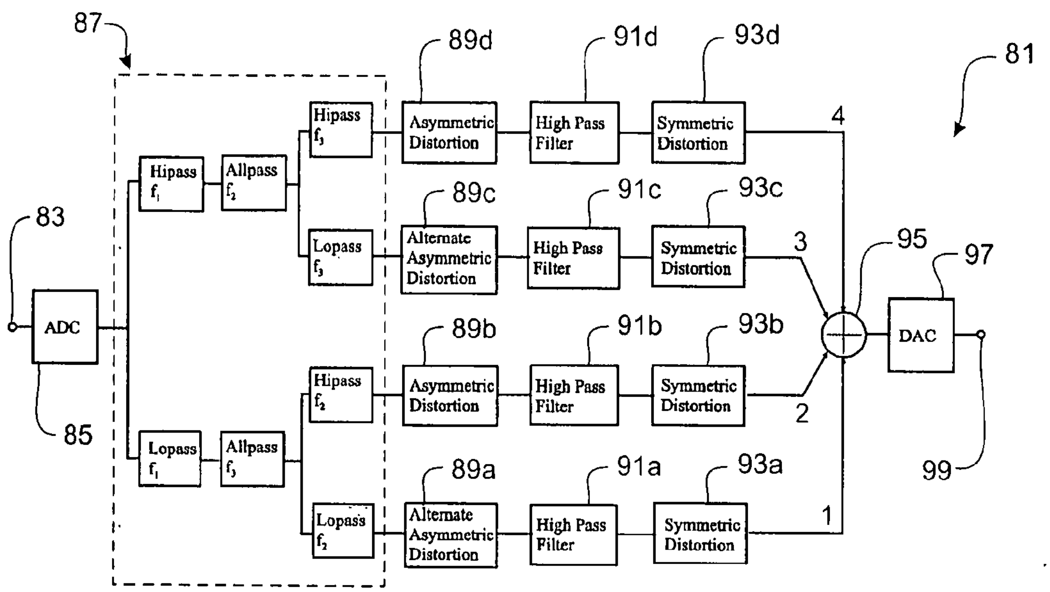


FIGURE 22

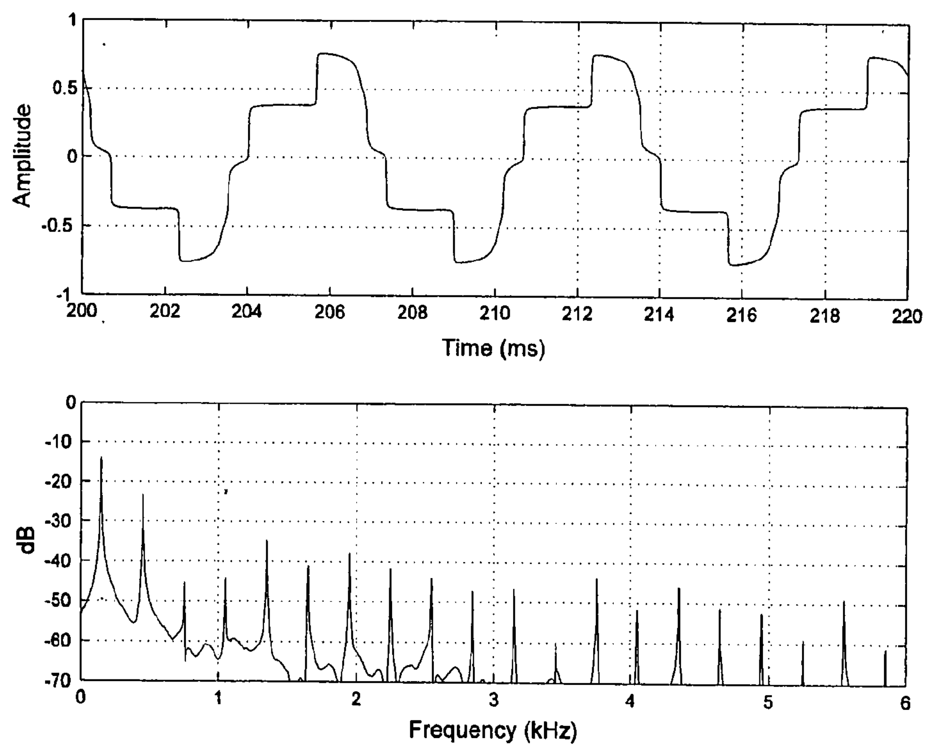


FIGURE 23

NONLINEAR PROCESSOR FOR AUDIO SIGNALS

FIELD OF THE INVENTION

[0001] The present invention relates to a nonlinear processor for musical signals that are generated by electronic instruments such as guitars and keyboards and musical signals from recorded acoustic instruments. More particularly, although not exclusively, the invention relates to the distortion of electric guitar signals to produce musically desirable sounds.

BACKGROUND TO THE INVENTION

[0002] The sound of the electric guitar is significantly dependent on the properties of the guitar amplifier. Guitar amplifiers typically have a non-flat frequency response aimed to enhance the sound of the guitar signal, such as by compensating for the guitar pickups or providing enhanced high frequencies for other subjective reasons. In addition, guitar amplifiers often operate in a highly nonlinear manner, distorting the guitar signal to produce harmonics and intermodulation frequency components which provides increased sustain and a more interesting and complex interaction between notes which is commonly used in pop, rock or heavy metal genres. In addition, the distortion produces output waveforms with high average power, particularly where the power amplifier saturates, so that the loudness of the amplifier for a given power rating is maximized.

[0003] Many of the properties of the electric guitar sound are related to the nonlinear behaviour of vacuum tube (valve) amplifiers, which were predominant when electric guitars were first developed. The majority of amplifiers built using modern technology seek to emulate the properties of tube amplifiers. See for example [E. Barbour: "The Cool Sound of Tubes", *IEEE Spectrum,* pp 24-35, August 1998, E. K. Pritchard: "The Tube sound and Tube Emulators," dB, pp 22-30, July/August 1994].

[0004] Many patents disclose devices that claim to emulate the operation of tube preamplifiers, which operate in class-A mode. Tube preamplifier stages produce bias-shifting when overdriven due to the grid conduction that occurs when the grid voltage exceeds the cathode voltage, in conjunction with the AC coupling between preamplifier stages. At high gains bias-shifting produces clipped waveforms resembling square waves with uneven mark-space ratios which include even harmonics. For example Sondermeyer [U.S. Pat. No. 5,619,578] discloses a multistage preamplifier using FETs with diode clipping to emulate grid conduction between stages.

[0005] Other patents disclose means for simulating one or more properties of tube power amplifiers, which typically operate in class AB or class B mode, having one or more symmetric pairs of output tubes coupled to the loudspeaker via an output transformer. Power amplifiers produce different characteristics to preamplifier tubes when overdriven. For example, symmetric-pair power stages produce crossover distortion when overdriven because grid conduction alters the input bias of the tubes. For example, Butler [U.S. Pat. No. 4,987,381] discloses a symmetric Mosfet output stage which claims to emulate the characteristics of vacuum tubes. Pritchard [U.S. Pat. Nos. 5,636,284 and 5,761,316] discloses means for emulating vacuum tube power amplifiers, including power supply compression effects, bias shifting due to grid conduction and variable output impedance. Sondermeyer [U.S. Pat. No. 5,524,055] also discloses a method for emulating the bias-shift due to grid conduction.

[0006] A feature of this form of crossover distortion is that as the input signal amplitude is reduced, the grid conduction ceases, and the crossover distortion disappears, so that the crossover artifacts only occur at high signal levels or high gains. This contrasts with crossover distortion in many solid state amplifiers, which is always present and so becomes objectionable at small signal levels.

[0007] A limitation of the emulation approach is that higher quality sound might in principle be achievable by modifying emulation circuitry so that it no longer precisely emulates a tube amplifier. For example, in the crossover distortion emulation circuits in U.S. Pat. Nos. 5,524,055 and 5,734,725, crossover distortion effects are obtained using diode clamping, which is highly nonlinear. This is reasonable for the emulation of the grid conduction that occurs in tubes when the input voltage rises above the bias voltage, but could be modified.

[0008] High quality guitar sound may also be achieved using circuitry that is significantly different to tube amplifiers. For example, one such technique is to filter the guitar signal into two or more frequency bands, to distort each band, and then to add the distorted bands together to produce a single output signal. Since notes with widely different frequencies fall within different frequency bands, the intermodulation distortion between those notes is reduced by this technique. The filter bands have sufficient and gradual overlap to ensure that some intermodulation occurs, and this produces a sound which is desirable for many music genres such as rock and heavy metal. This technique is discussed in [C. Anderton, "Four fuzzes in one with active EQ, Guitar Player, pp 37-46, June 1984], which discloses a four band system using standard bandpass filters.

[0009] An improvement to the bandpass filtering operation is to use equi-phase crossover networks to separate the signals into two or more bands as discussed in [M. Poletti, "An improved guitar preamplifier system with controllable distortion", NZ Patent 329119], which is incorporated herein by reference. Equi-phase networks are commonly applied to multi-way loudspeaker systems [see for example S. H. Linkwitz, "Active crossover networks for noncoincident drivers,"*J. Audio Eng. Soc., Vol.* 24, No. 1, pp 2-8, January/February 1976] and have the advantage that the sum of the bands produces a flat frequency response, and so the band-splitting and recombination operation does not alter the pre-existing frequency spectrum of the signal input to the bandsplitting network. When applied to nonlinear distortion of guitar signals, the output of the equi-phase system has a lower crest factor and a higher rms level than non-equi-phase systems and therefore produces a greater loudness for a fixed power amplifier rating, allowing it to better compete with tube amplifiers in which the power amplifier saturates.

The Effect of Crossover Distortion in Valve Power Amplifiers

[0010] An interesting characteristic of tube amplifiers is the crossover distortion that occurs in the power amplifier when overloaded. This process is discussed by Sondermeyer in [U.S. Pat. No. 5,524,055], where it is stated that when grid conduction occurs the output tubes become overbiased, causing crossover distortion, and that this reduces the peak clipping of the waveform. However, this reduction of peak clipping does not explain the spectrum of the output waveform, as will now be demonstrated.

[0011] FIG. 1 shows the output of a tube power amplifier driven into overload for a 250 Hz sinewave input, with a resistive load, with the recorded waveform normalized to a peak amplitude of one. The limiting of the peaks of the sinewave and the crossover distortion due to grid conduction are clear. The spectrum shows a modulated envelope, with both even and odd harmonics, and with a minimum in the envelope in the region of 1 kHz. This contrasts with the spectrum of a sinewave clipped to a similar level, as shown in FIG. 2, which has only odd harmonics, and an envelope which decays in a more monotonic manner with frequency and with only slight variations in magnitude. At higher gains the clipped sinewave becomes close to that of a square wave, and the spectrum consists of the fundamental plus all odd mth harmonics, with amplitudes 1/mth of that of the fundamental. The envelope of the spectrum then falls monotonically with frequency. However, with crossover distortion, the spectrum at higher gains maintains its modulated envelope. For example, FIG. 3 shows a heavily distorted sinewave with crossover distortion. The spectrum—shown in the middle plot of FIG. 3—shows a similar characteristic modulation of the spectrum to FIG. 1, with a first null at 4 kHz. Since most guitar amplifier loudspeakers roll off above 4 kHz, the reduction in the spectrum at 4 kHz will produce a reduction of high frequencies and an improvement in subjective sound quality compared to the spectrum without crossover distortion.

[0012] The characteristic modulation of the spectrum for heavily clipped sinewaves with crossover distortion may be explained by a Fourier analysis. The waveform is similar to a single period of a square wave with a "dead-zone" crossover region, as shown in FIG. 4. A single cycle of this waveform consists of two pulse signals, $p_{\tau/2}(t)$, of width $\tau/2$, delayed by $-T/4$ and $T/4$, and with the second pulse inverted. For $\tau=T$ the crossover region is zero and the signal becomes one period of a square wave. The time signal can be written

$$s(t)=p_{\tau/2}(t+T/4)-p_{\tau/2}(t-T/4) \tag{1}$$

The Fourier transform is

$$S(f)=2j\frac{\sin(\pi f\tau/2)\sin(\pi fT/2)}{\pi f} \tag{2}$$

When the signal is repeated periodically, the spectrum is sampled at f=m/T, and scaled by 1/T, yielding the discrete spectrum of the periodic signal

$$S(m)=\frac{2j}{m\pi}\sin\left(\frac{m\pi}{2}\frac{\tau}{T}\right)\sin\left(\frac{m\pi}{2}\right) \tag{3}$$

For $\tau=T$ the sine terms become one and the spectrum reduces to

$$S(m)=\frac{2j}{m\pi},\ m\text{ odd} \tag{4}$$

which is the spectrum of a square wave. For $\tau<T$ the product of the two sine terms produces a slowly varying envelope whose rate increases as τ reduces. The theoretical spectrum according to equation 3 is shown in the lower plot in FIG. 3 for $\tau/T=0.962$, and is a reasonable match to the measured spectrum of the signal.

[0013] The modulation of the envelope increases as the degree of crossover distortion increases. FIG. 5 shows a sinewave distorted with a greater degree of crossover distortion. The first null in the envelope of the spectrum has reduced from 4 kHz to 2 kHz and the magnitude at 4 kHz is increased. The theoretical spectrum is shown with $\tau/T=0.92$ and is a good match. Since 4 kHz is the typical upper limit of guitar loudspeakers, the increase in signal energy near 4 kHz increases the upper harmonics of the perceived waveform, which is likely to reduce the subjective sound quality.

[0014] Hence, the crossover distortion which occurs in tube amplifiers can produce a subjective improvement to the sound of distorted guitar signals, provided that the crossover effect is limited so that a reduction in spectral components occurs at the maximum frequencies which are transmitted by the guitar loudspeaker.

[0015] In this specification where reference has been made to patent specifications, other external documents, or other sources of information, this is generally for the purpose of providing a context for discussing the features of the invention. Unless specifically stated otherwise, reference to such external documents is not to be construed as an admission that such documents, or such sources of information, in any jurisdiction, are prior art, or form part of the common general knowledge in the art.

[0016] It is an object of the present invention to provide a nonlinear processor for audio signals that is capable of producing controllable crossover-like distortion, or to at least provide the public with a useful choice.

SUMMARY OF THE INVENTION

[0017] In a first aspect, the present invention broadly consists in a nonlinear processor for distorting audio signals, comprising: an input stage that is arranged to split an audio input signal into two signal paths; a pair of asymmetric distortion stages following the input stage such that there is one asymmetric distortion stage in each signal path, each asymmetric distortion stage having non-equal negative and positive saturation limits and a smooth transition between linear and nonlinear behaviour, and being arranged to produce a distorted output signal that has a mean signal level that is opposite in polarity to the other asymmetric distortion stage; a pair of AC-coupled symmetric distortion stages following the asymmetric distortion stages such that there is one symmetric distortion stage in each signal path, each symmetric distortion stage being arranged to nonlinearly limit the distorted signals in each signal path; and an output stage following the symmetric distortion stages that is arranged to add the two nonlinearly distorted signals from the symmetric distortion stages to generate an audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artifacts.

[0018] In one form, the processor may be implemented in an analogue circuit wherein the input stage may be arranged to receive an analogue audio input signal, buffer the input signal, and split the input signal into two signal paths, and wherein the output stage may be arranged as a summer for adding the two analogue nonlinearly distorted signals from the symmetric distortion stages to generate a single analogue audio output signal.

[0019] In an alternative form, the processor may be implemented in a digital system wherein the input stage comprises an analogue-to-digital converter that may be arranged to receive an analogue audio input signal, convert the analogue input signal into a digital input signal, and split the digital input signal into two digital signal paths, and wherein the output stage may comprise: a summer that may be arranged to add the two digital nonlinearly distorted signals from the symmetric distortion stages to generate a single digital audio output signal; and a digital-to-analogue converter that may be arranged to convert the single digital audio output signal into a single analogue audio output signal.

[0020] In one form, the magnitude of the positive and negative saturation limits for one of the asymmetric distortion stages may be substantially equal to the magnitude of the negative and positive saturation limits respectively for the other asymmetric distortion stage so as to produce an audio output signal at the output stage that demonstrates a smooth transition from linear behaviour to the production of crossover-like artefacts.

[0021] In an alternative form, the magnitude of one or both of the positive and negative saturation limits for one of the asymmetric distortion stages may be different to the magnitude of the negative and positive saturation limits respectively for the other asymmetric distortion stage so as to produce an audio output signal at the output stage that demonstrates a smooth transition from linear behaviour to the production of crossover-like artefacts, with a spectrum which includes even harmonics of input frequencies of the audio input signal. Preferably, the magnitude of the positive saturation limit for one of the asymmetric distortion stages may be substantially higher than the magnitude of the negative saturation limit for the other asymmetric distortion stage.

[0022] Preferably, the symmetric distortion stages may each comprise a low-pass filter to provide a reduction of harmonic energy when nonlinearly limiting the distorted signals from the asymmetric distortion stages.

[0023] Preferably, the audio input signal may be from an electric or electronic musical instrument.

[0024] In a second aspect, the present invention broadly consists in a multiband nonlinear processor for distorting audio signals, comprising: an input stage that is arranged to receive an audio input signal: an equi-phase crossover network that is arranged to split the input signal into two or more frequency bands with finite overlap between the frequency bands, and equal phase responses in each band, and in each frequency band: an asymmetric distortion stage having non-equal negative and positive saturation limits and a smooth transition from linear to nonlinear behaviour, and where the saturation limits alternate across the frequency bands so as to produce distorted output signals having alternating polarity mean signal levels across the frequency bands; and an AC-coupled symmetric distortion stage following the asymmetric distortion stage that is arranged to nonlinearly limit the distorted output signal from the asymmetric distortion stage; and an output stage that is arranged to add the nonlinearly distorted signals from the symmetric distortion stages of all frequency bands to generate an audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artifacts, with a reduction of intermodulation distortion.

[0025] In one form, the processor may be implemented in an analogue circuit wherein the input stage may be arranged to receive an analogue audio input signal and buffer it into the equi-phase crossover network, and wherein the output stage may be arranged as a summer for adding the analogue output signals from all the frequency bands to generate a single analogue audio output signal.

[0026] In another form, the processor may be implemented in a digital system, and wherein the input stage may comprise an analogue-to-digital converter that may be arranged to receive an analogue audio input signal and convert it into a digital input signal for the equi-phase crossover network, and wherein the output stage may comprise: a summer that may be arranged to add the digital output signals from all frequency bands to generate a single digital audio output signal; and a digital-to-analogue converter that may be arranged to convert the single digital audio output signal into a single analogue audio output signal.

[0027] In one form, the magnitude of the positive and negative saturation limits of each asymmetric distortion stage may be substantially equal to the magnitude of the negative and positive saturation limits respectively of adjacent asymmetric distortion stages of adjacent frequency bands so as to produce an audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artifacts, with a reduction of intermodulation distortion.

[0028] In an alternative form, one or both of the positive and negative saturation limits of each asymmetric distortion stage may be different to the magnitude of the negative and positive saturation limits respectively of adjacent asymmetric distortion stages of adjacent frequency bands so as to produce an audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artifacts, with a reduction of intermodulation distortion, and with a spectrum which includes even harmonics of the input frequencies of the audio input signal.

[0029] Preferably, the symmetric distortion stages may each comprise a low-pass filter to provide a reduction of harmonic energy when nonlinearly limiting the distorted signals from the asymmetric distortion stages.

[0030] Preferably, the multiband nonlinear processor may further comprise cross-coupling between the frequency bands before the distortion stages to allow the controlled increase of intermodulation distortion.

[0031] Preferably, the audio input signal may be from an electric or electronic musical instrument.

[0032] In a third aspect, the present invention broadly consists in a nonlinear audio distortion circuit for distorting audio signals from musical instruments, comprising: an input stage that is arranged to split an audio input signal into two signal paths; a pair of asymmetric distortion stages, one in each signal path, with non-equal negative and positive saturation limits, so as to produce opposite polarity mean signal levels at their outputs in each signal path, and which produce a smooth transition from linear to nonlinear behaviour; a pair of AC-coupled symmetric distortion stages, one in each signal path, following the asymmetric distortion stages; and an output stage that is arranged to add the two nonlinearly distorted signals from the symmetric distortion stages to generate an audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artifacts.

[0033] In one form, the saturation limits in the two asymmetric distortion stages may be the opposite of each other so as to produce an audio output signal at the output stage that demonstrates a smooth transition from linear behaviour to the production of crossover-like artefacts.

[0034] In another form, the saturation limits of the two asymmetric distortion stages may be different to each other so as to produce a final audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artefacts, with a spectrum which includes even harmonics of the input frequencies of the audio input signal.

[0035] Preferably, the symmetric distortion stages may each comprise an amplifier with a feedback loop that may be arranged to nonlinearly limit the signal of its signal path and a low-pass filter in the feedback loop that is arranged to provide a reduction of harmonic energy when limiting the signal.

[0036] The phrase "mean signal level(s)" in relation to the outputs of the asymmetric distortion stages, and in the context of polarity, is intended to cover the polarity of the time-average of the analogue outputs over a time equal to one or more periods of the input fundamental frequency in terms of voltage for the analogue implementation of the nonlinear processor and the sign of the time-average of the digital outputs over a time equal to one or more periods of the input fundamental frequency in terms of digital signal values for the digital implementation of the nonlinear processor.

[0037] The term 'comprising' as used in this specification means 'consisting at least in part of', that is to say when interpreting statements in this specification which include that term, the features, prefaced by that term in each statement, all need to be present but other features can also be present.

[0038] The invention consists in the foregoing and also envisages constructions of which the following gives examples only.

BRIEF DESCRIPTION OF THE DRAWINGS

[0039] Preferred embodiments of the invention will be described by way of example only and with reference to the drawings, in which:

[0040] FIG. **1** shows temporal (top) and spectral (bottom) graphs of an output signal from a tube power amplifier that is driven into overload for a 250 Hz sinewave input;

[0041] FIG. **2** shows temporal (top) and spectral (bottom) graphs of a clipped sinewave;

[0042] FIG. **3** shows a temporal graph (top) of a heavily distorted sinewave with crossover distortion, a spectral graph (middle) calculated from the FFT of the heavily distorted sinewave, and a spectral graph (bottom) of the heavily distorted sinewave derived from the theoretical model of FIG. **4** with $\tau/T=0.962$;

[0043] FIG. **4** shows a theoretical crossover distortion model;

[0044] FIG. **5** shows a temporal graph (top) of the heavily distorted sinewave of FIG. **3** with a greater degree of crossover distortion, a spectral graph (middle) calculated from the FFT of the heavily distorted sinewave with greater crossover distortion, and a spectral graph (bottom) of the heavily distorted sinewave with greater crossover distortion derived from the theoretical model of FIG. **4** with $\tau/T=0.92$;

[0045] FIG. **6** shows a first preferred embodiment of the nonlinear processor of the present invention in the form of an analogue circuit for producing crossover-like artifacts;

[0046] FIGS. **7***a* and **7***b* show the transfer characteristics of the upper asymmetric and lower alternate asymmetric distortion amplifiers respectively of the analogue circuit of FIG. **6**;

[0047] FIG. **8** shows modelled temporal graphs of the output waveforms from the upper asymmetric (top) and lower alternate asymmetric (bottom) distortion amplifiers of the analogue circuit of FIG. **6**, with the output waveforms after AC-coupling shown dashed;

[0048] FIG. **9** shows modelled temporal graphs of the output waveforms from the upper (top) and lower (bottom) symmetric distortion amplifiers of the circuit of FIG. **6**;

[0049] FIG. **10** shows modelled temporal (top) and spectral (bottom) graphs of the output waveform from the summer operational amplifier of the analogue circuit of FIG. **6**;

[0050] FIG. **11** shows the modelled temporal (top) and spectral (bottom) graphs of the output waveform from FIG. **10**, but where the saturation levels of the asymmetric distortion amplifiers are different to each other, resulting in a reduced width in the positive cycle and additional even harmonics;

[0051] FIG. **12** shows a schematic block diagram of a second preferred embodiment of the nonlinear processor of the present invention in the form of a digital system for generating signal limiting and crossover-like artifacts;

[0052] FIGS. **13***a* and **13***b* show the transfer characteristics of the upper asymmetric and lower alternate asymmetric distortion stages respectively of the digital system of FIG. **12**;

[0053] FIG. **14** shows a third preferred embodiment of the nonlinear processor of the present invention in the form of a four-band analogue circuit for producing crossover-like artifacts with controllable intermodulation distortion;

[0054] FIGS. **15***a*-**15***c* show examples of typical high-pass, low-pass and all-pass filters, respectively, that may be implemented in the analogue circuit of FIG. **14**;

[0055] FIG. **16** shows modelled temporal graphs of the output waveforms from the asymmetric distortion stages of channels **1**-**4** of the four-band analogue circuit of FIG. **14** for an input of 150 Hz;

[0056] FIG. **17** shows modelled temporal graphs of the output waveforms from the symmetric distortion stages of channels **1**-**4** of the four-band analogue circuit of FIG. **14** for an input of 150 Hz;

[0057] FIG. **18** shows modelled temporal (top) and spectral (lower) graphs of the output waveform from the summer operational amplifier of the output stage of the four-band analogue circuit of FIG. **14** for an input of 150 Hz;

[0058] FIG. **19** shows modelled temporal (top) and spectral (bottom) graphs of the output waveform from FIG. **18**, but where the saturation levels of each of the asymmetric distortion stages are different or unmatched relative to those of the adjacent asymmetric distortion stages, resulting in a reduced width in the positive cycle and increased even harmonics;

[0059] FIG. **20** shows modelled temporal graphs of the output waveforms from the asymmetric distortion stages of channels **1**-**4** of the four-band analogue circuit of FIG. **14** for an input of 1.5 kHz;

[0060] FIG. **21** shows modelled temporal (top) and spectral (bottom) graphs of the output waveform from the summer operational amplifier of the output stage of the four-band analogue circuit of FIG. **14** for an input of 1.5 kHz;

[0061] FIG. **22** shows a schematic block diagram of a fourth preferred embodiment of the nonlinear processor of the present invention in the form of four-band digital system for producing crossover-like artifacts with controllable intermodulation distortion; and

[0062] FIG. **23** shows modelled temporal (top) and spectral (bottom) graphs of the output waveform of the output stage of the digital system of FIG. **22** with a non-equi-phase bandsplitter for an input of 150 Hz.

DETAILED DESCRIPTION OF PREFERRED EMBODIMENTS

[0063] The present invention is directed at a nonlinear processor for audio signals that is capable of producing controllable crossover-distortion-like effects without requiring the use of D.C. biasing, and which can produce a more gradual transition into crossover distortion than obtained by tube emulation. The nonlinear processor can be implemented in analogue or digital form as will be described, by way of example, with reference to the first and second preferred embodiments of FIGS. **6** and **12** respectively.

[0064] The present invention may also enable the incorporation of controllable crossover-like effects into a multiband nonlinear processor to reduce harmonic distortion while also offering control of intermodulation distortion. The multiband nonlinear processor may also be implemented in analogue or digital form as will be explained with reference to the third and fourth preferred embodiments of FIGS. **14** and **22** respectively.

[0065] Referring to FIG. **6**, the first preferred embodiment of the nonlinear processor is shown in the form of a solid-state analogue circuit **11**. The analogue circuit **11** will now be explained in more detail below. The analogue circuit **11** is capable of producing amplitude-dependent crossover distortion where the distortion is minimal for small signal amplitudes and the transition into crossover effects is gradual.

[0066] The input **13** is connected to an input stage **15**, for example a unity gain buffer circuit, whose output is connected to two class A circuits which operate in parallel upper 16 and lower 18 channels. The first amplifier circuit **17**, **19** in each channel has an asymmetric, nonlinear transfer characteristic. The gain of each amplifier circuit **17**, **19** for small input voltages is $-R_2/R_1$. At larger voltages the gain reduces due to the conduction of the diodes in the feedback network in parallel with R_2 and r_1 and r_2, which are typically smaller than R_2. Since the circuits **17**, **19** use diodes in the feedback loop of the operational amplifiers, the transfer characteristic is smoother than can be obtained using a diode clipper with diodes connected to ground. Furthermore, the negative output voltage saturation limit of the asymmetric distortion stage is different to the positive output voltage saturation limit. For example, for the lower channel **18** the negative limit is the diode voltage, V_d, required to maintain the virtual earth condition, which would typically be of the order of −0.6 volts. The positive limit is $(1+ r_2/r_1)V_d$ for example with r_1=100 Ohms and r_2=1 kOhm the positive limit would be about 6.6 volts. The transfer characteristic therefore typically has the form of FIG. **7***b*, where V_n=−0.6 and V_p=6.6. Note that the transfer characteristic includes the inversion of the input voltage due to the inverting configuration of the operational amplifier circuit **19**.

[0067] The upper channel **16** uses the same circuit, but the asymmetry has the opposite polarity to the lower channel **18**. With the same values of r_1 and r_2 the negative saturation limit would be −6.6 volts and the positive voltage saturation limit 0.6 volts, and the transfer characteristic would have the alternate asymmetry, as shown in FIG. **7***a*, including inversion of the input voltage.

[0068] Due to the non-equal clipping voltages, the output waveforms from the asymmetric amplifier circuits **17**, **19** of the two channels **16**, **18** have non-zero average voltages with opposite polarities, a representative waveform of which is shown in FIG. **8**. These signal voltages are preferably AC coupled into the next stages, which removes the DC offsets from the two signals. The AC coupled waveforms are shown dashed in FIG. **8**.

[0069] The following nonlinear amplifier stages **21**, **23** are arranged to nonlinearly limit the waveforms in each of the channels **16**, **18** symmetrically with respect to each other. The gain for small signal voltages is $-R_4/R_3$, and this reduces for large input voltages, and the reduction in gain is equal for positive or negative input voltages. Because of the asymmetry of the input waveforms, the output waveforms from the symmetric amplifier circuits **21**, **23** produce distorted waveforms with unequal durations of negative and positive going excursions (an unequal "mark-space" ratio), as shown in FIG. **9**.

[0070] The two symmetric distortion outputs are added in an output stage with equal gains $-R_6/R_5$ in the final summer operational amplifier circuit **25**, producing an output **27** with characteristics similar to those of crossover distortion, as shown in FIG. **10**, although it is produced by a different mechanism to that which occurs in a tube amplifier. Furthermore, the transition into crossover distortion is smoother than prior art methods, because the diodes in the asymmetric distortion stages **17**, **19** are in the feedback loops of the operational amplifiers. This produces a gradual, rounded clipping of the signals which does not occur in tube grid conduction, and this helps to produce a slower transition into asymmetry.

[0071] If the two asymmetric stages **17**, **19** have different saturation levels but still produce opposite polarity mean voltages at their outputs, crossover distortion will still occur, but the width of the positive and negative halves of the waveform will differ. This introduces even harmonics into the spectrum. For example, if the asymmetric amplifier circuit **19** of the lower channel **18** stage has voltage saturation limits of $V_n=-6.6$ and $V_p=0.6$ and the alternate asymmetric amplifier circuit **17** of the upper stage has saturation limits $V_n=0.6$ and $V_p=26.4$, then the output **27** in FIG. **11** is produced. The positive half of the waveform exhibits a narrower width than the negative half, and the spectrum shows odd harmonics, and even harmonics at a lower level relative to the adjacent odd harmonics. The degree of crossover is also increased, altering the modulation of the spectrum. The addition of even harmonics creates a subjectively different sound quality and this is a desirable option which can be implemented as required. This feature is easily implemented using the analogue circuit **11** of FIG. **6**, but does not occur under normal operation in a tube power amplifier, giving the nonlinear processor a flexibility which exceeds that of the tube power amplifier.

[0072] The nonlinear processor, shown in FIG. **6** as analogue circuit **11**, may also be implemented digitally as will be described with reference to the second preferred embodiment of the nonlinear processor, in particular the digital system **31** of FIG. **12**.

[0073] The analogue input signal **33** is first sampled at the input stage in an analogue-to-digital converter (ADC) **35** at a rate sufficiently high to accommodate the distortion products generated by the subsequent nonlinear processing. The sampled signal is then split into upper 37 and lower 39 channels. An asymmetric distortion stage **41** is applied to the upper channel **37**, and an alternate asymmetric distortion stage **43** is applied to the lower channel **39**. The outputs from the asymmetric distortion stages **41**, **43** are then preferably high-pass filtered **45** (AC coupled) to remove the DC component. Each AC coupled sampled waveform is then applied to symmetric distortion stages **47**, **49** provided in the upper 37 and lower 39 channels. The outputs from the symmetric distortion stages **47**, **49** are then added together at the output stage by summer **51**. The output of the summer **51** is then applied to a digital-to-analogue converter (DAC) **53** that provides a single analogue output **55** demonstrating crossover-like artifacts.

[0074] A method of producing an asymmetric, nonlinear transfer characteristic for the asymmetric distortion stages **41**, **43** of the digital system **31** is

$$f(x) = \begin{cases} \dfrac{gx}{1 - gx/L_n}, & x \le 0 \\ \dfrac{gx}{1 + gx/L_p}, & x > 0 \end{cases} \quad (5)$$

which is a simplification and modification of the function given in [M. C. Jeruchim, P. Balaban and K. S. Shanmugan, *Simulation of Communication Systems*, Plenum Press, 1992]. This produces a gain g for x=0, a negative limit of $f(x)=-L_n$ for x<<0 and a positive limit of $f(x)=L_p$ for x>>0. For example, a transfer characteristic for the asymmetric distortion stage **41** of the upper channel **37** is shown in FIG. **13***a* for g=40, a negative limit of −1 and a positive limit of 4. FIG. **13***b* shows the alternate transfer characteristic for the alternate asymmetric distortion stage **43** of the lower channel **39** with the same gain, a negative limit of −4 and a positive limit of 1. These transfer characteristic curves are similar in form to those shown in FIGS. **7***a* and **7***b* in relation to analogue circuit **11**, but do not include inversion of the input signal.

[0075] The high-pass filter stages **45** may be implemented using standard first order filter designs such as a digital Butterworth filter or any other type of suitable filters. Higher order filters may also be utilised if desired. The symmetric distortion stages **47** may be obtained using equation 5, with $L_n=L_p$.

[0076] The modelled waveforms shown in FIGS. **8** to **10** were obtained using equation 5 with $L_n=-6.6$ and $L_p=0.6$ for the upper channel asymmetric distortion stage **41** and $L_n=-0.6$ and $L_p=6.6$ for the lower channel alternate asymmetric distortion stage **43**, and are essentially similar in form to the analogue voltages waveforms produced by the analogue circuit **11** in FIG. **6**. Both L_n and L_p were set to one for the symmetric distortion stages **47**, **49**. In FIG. **11** the upper channel asymmetric distortion stage **41** used $L_p=24.6$ to produce additional even harmonics of the input frequency. A sample rate of 176400 Hz was used, and digital high-pass filters **45** each with a 10 Hz cut off were utilised (with feedforward coefficients 0.9998 and −0.9998, and feedback coefficient −0.9996). The input sinewave had a frequency of 150 Hz and amplitude 1 and the asymmetric stage gains were 40 and the symmetric stage gains were 4.

[0077] As mentioned, the nonlinear processor may be implemented in a multiband form to reduce harmonic distortion and to provide controllable crossover-like artifacts and reduced intermodulation distortion. Referring to FIG. **14**, a third preferred embodiment of the nonlinear processor in the form of a solid-state multiband analogue circuit **61** is shown. This embodiment will be explained in more detail below.

[0078] The analogue input signal **63** is first buffered at an input stage by input buffer **65** in a similar manner to analogue circuit **11** described with reference to FIG. **6**. The buffered output is then split into four frequency bands using an equi-phase bandsplitter **67**, an example of which is as discussed in NZ Patent 329119. The low-pass, high-pass and all-pass filters of the bandsplitter **67** may be implemented, for example, as shown in FIG. **15***a* (second order high-pass), **15***b* (inverting, second order low-pass) and **15***c* (first order all-pass). The four outputs or frequency bands from the bandsplitter are fed into four asymmetric distortion stages **69***a*-**69***d*. The small-signal gains of these stages are $-R_2/R_1$, and the gain reduces at higher signal levels due to the conduction of the diodes in series with R_2 and the feedback resistors r_1 and r_2, which are typically smaller than R_2. The asymmetry in each band is of the opposite polarity to that in the adjacent band or channel. Specifically, the asymmetric distortion stage **69***a* in channel one has a large positive output voltage saturation limit of approximately $(1+r_2/r_1)V_d$ and a small negative voltage saturation limit of $-V_d$. The asymmetric distortion stage **69***b* in channel two has the opposite polarity to that in channel one, with small positive output voltage limit V_d and large negative voltage limit $-(1+r_2/r_1)V_d$. The asymmetric distortion stage **69***c* of channel three has the opposite polarity to that in channel two, and the same polarity as that in channel one. Finally, the asymmetric distortion stage **69***d* of channel four has the opposite polarity to that in channel three, but the same polarity to that in channel two. In summary, the asymmetry alternates across the four channels or frequency bands.

[**0079**] Representative waveforms at the outputs of the asymmetric distortion stages **69***a*-**69***d*, and their dashed AC-coupled forms, are shown in FIG. **16**, for an input **63** having a signal frequency of 150 Hz. The signal energy is predominantly in channels one and two (since adjacent channels overlap due to the finite roll-off of the bandsplitting filters). The asymmetry alternates across the channels, but the signal amplitude is reduced in the upper two channels.

[**0080**] The outputs from the asymmetric distortion stages **69***a*-**69***d* are AC-coupled into symmetric distortion stages **71***a*-**71***d*. These have gains of $-R_4/R_3$ for small voltages, and the gain reduces for large input voltages, and the reduction in gain is approximately equal for positive or negative input voltages. In those channels where the signal energy is sufficiently large, this produces waveforms with non-even mark-space ratios. A representative example is shown in FIG. **17** for a 150 Hz sinewave input **63**. The first and second channels produce distorted waveforms which are similar to square waves, and which have non-equal mark-space ratios. The sum of these waveforms generated by summing circuit **73** of the output stage produces an output **75** with crossover effects reminiscent of standard crossover distortion, the waveform and spectrum of which are shown in FIG. **18**.

[**0081**] In tube amplifiers, the crossover distortion in the output waveform occurs at zero volts for a symmetrical output stage. The crossover effect in FIG. **18** occurs at a voltage which depends on the input sinewave and the bandsplitting frequencies, and may not be at zero volts. The spectrum of the waveform shows the typical characteristic of crossover distortion, with a modulation of the spectral envelope, as shown in the lower graph of FIG. **18**. The waveform is non-symmetric for a non-zero crossover voltage, and as a result the spectrum includes even harmonics of the input frequency. The inclusion of even harmonics due to this waveform asymmetry can be subjectively desirable. The effect can be increased or decreased by altering the saturation limits as discussed in relation to FIG. **6**, and shown in FIG. **11**. FIG. **19** shows the waveform of the output **75** for negative saturation limits of 1, 10, 1 and 10, and positive saturation limits of 40, 1, 40 and 1. The positive half of the waveform has a reduced width, and this further enhances the even harmonics compared to FIG. **18**. If the positive saturation limits were instead decreased to less than 10, then the even harmonics would be reduced.

[**0082**] In addition to producing crossover distortion effects, the analogue circuit **61** of FIG. **14** also produces reduced intermodulation distortion between frequencies which are sufficiently separate to fall predominantly into different channels. For example, FIG. **20** shows the output of the symmetric distortion stages **71***a*-**71***d* for a 1.5 kHz sinewave input **63**. The energy in the signal now resides predominantly in the third and fourth channels, as opposed to the first and second as in FIG. **17**. FIG. **21** shows the combined waveform at the output **75** which produces crossover-like artifacts at a positive voltage. Since the 150 Hz and a 1.5 kHz signals occur predominantly in different channels, the intermodulation between these two frequencies will be significantly reduced, and the output of the circuit for an input consisting of the sum of the two sinewaves will be predominantly the sum of the waveforms in FIGS. **18** and **21**.

[**0083**] The multiband nonlinear processor, shown in FIG. **14** as analogue circuit **61**, may also be implemented digitally as will be described with reference to the fourth preferred embodiment of the nonlinear processor, in particular the digital system **81** of FIG. **22**.

[**0084**] The analogue input signal **83** is first sampled at the input stage by ADC **85** at a rate sufficiently high to accommodate the distortion products generated by the subsequent nonlinear processing. The sampled signal is split into four channels or frequency bands by an equi-phase bandsplitter **87** that, for example, utilises digital filters obtained from the bilinear transform of the filters in FIGS. **15***a*-**15***c*. Asymmetric distortion stages **89***a*-**89***d* are provided in each channel, for example using the nonlinear function in equation 5. These asymmetric distortion stages **89***a*-**89***d* alternate across the four channels in a similar manner to that described in relation to the analogue circuit **61** of FIG. **14**, with opposite polarities between even and odd channels. The outputs of the asymmetric distortion stages **89***a*-**89***d* are AC-coupled using high-pass digital filters **91***a*-**91***d* and fed into symmetric distortion stages **93***a*-**93***d*, using for example equation 5 with equal negative and positive limits. The outputs of the symmetric distortion stages **93***a*-**93***d* are then added together at the output stage by summer **95** to produce, after being fed through DAC **97**, an analogue output **99** with similar properties to the output of the analogue circuit **61** of FIG. **14**. The control of even harmonics can be implemented in similar form to FIG. **14** by adjusting the relative saturation limits of the asymmetric distortion stages **89***a*-**89***d*, whilst maintaining opposite polarities of their mean output waveforms between adjacent channels.

[**0085**] It will be appreciated that the multiband nonlinear processor may be arranged to split the input signal into two or more frequency bands or channels, and that the four-band embodiments are provided by way of example only.

[**0086**] A distinction should be made between the effects on sound quality of using a prior-art, non-equi-phase, band-pass-filter-based bandsplitter with different phase responses between bands and symmetric distortion, as in [C. Anderton, "Four fuzzes in one with active EQ, Guitar Player, pp 37-46, June 1984], and the method disclosed here. The use of non-equi-phase bandsplitting produces waveforms in each band with widely different phase responses. This occurs because each bandpass filter must be positioned at a different frequency, and so the phase responses must be different between filters. This means that, when the bands are combined, the degree of crossover distortion is significant, and is frequency-dependent. Severe crossover artifacts occur at most frequencies within the range of interest which—as shown in FIGS. **3** and **5**—does not produce a reduction of high frequency harmonics near the bandlimit of the guitar loudspeaker, and hence produces no benefit. In addition, the waveforms produced in the non-equi-phase case can have high crest factors.

[0087] For example, FIG. **23** shows the output of a multiband nonlinear processor using four bandpass filters with non-equi-phase responses (with center frequencies 100, 300, 900 and 2700 Hz), for a 150 Hz input signal. At 150 Hz the phase difference between bands one and two is about 90 degrees. The output waveform therefore produces maximal forms of crossover distortion, as shown, and the crest factor of the output is 4 dB as opposed to 1.5 dB in FIG. **18**. This means that the non-equi-phase waveform will not be as loud as the equi-phase waveform when transmitted from a power amplifier with limited headroom.

[0088] Further, prior art bandsplitters will always produce the most extreme crossover in the region where bandsplitting is applied, since this is where the phase differences are maximum, so the problem is difficult to avoid without employing equi-phase bandsplitting as described in NZ Patent 329119. Furthermore, the crossover distortion caused by non-equi-phase networks occurs at all signal levels, since it is not the result of asymmetric distortion as used in the present invention, or bias shift as in the tube amplifier case. Therefore non-equi-phase bandsplitting will produce significant effects at lower signal amplitudes, whereas in the method disclosed here crossover distortion disappears at small signal levels, which is more desirable. Lastly, due to the symmetric distortion in each stage, the prior art circuit produces only odd harmonics, with no control of even harmonics. The use of equi-phase bandsplitting and controlled alternating asymmetry as described herein thus provides for output waveforms with controllable crossover distortion artifacts at all frequencies which remain subjectively desirable for all input signals, which are signal-level-dependent, and the output waveform always exhibits a low crest factor which maximizes loudness.

[0089] It will be understood that various modifications can be made to the analogue circuits of FIGS. **6** and **14** without substantially altering their operation, or which further enhance the subjective sound quality. For example, input gain and equalisation may be applied to the signal before nonlinear processing, and equalization (tone controls) may be applied to the output of the nonlinear processor. Low-pass filters may be placed after the symmetric distortion stages, or symmetric distortion stages used which incorporate low-pass filters as discussed in NZ Patent 329119. The asymmetric distortion circuits may be simplified by removing r_1. Alternative forms of asymmetric distortion stages may use transistors to provide continuously variable voltage limits, or diodes with different on-voltages, such as zener or light emitting diodes. Different forms of asymmetric distortion may be used in each channel to produce crossover-like artifacts, the spectrum of which includes even harmonics of the input signal. Symmetric distortion stages with different nonlinear elements in the feedback loop may also be used such as light emitting diodes, zener diodes or transistors, or circuits without nonlinear elements in the feedback loop such as a resistor and pairs of diodes to ground may be used to produce increased harmonic energy for more extreme sounds. Lastly, deliberate cross coupling between the bandsplitter outputs before nonlinear distortion may be introduced to allow the controlled increase of intermodulation distortion for musical purposes, or alternatively, controlled nonlinear distortion of the combined output may be added for similar reasons. Similarly, it will be appreciated that various modifications may be made to the digital systems of FIGS. **12** and **22** if desired. For example, input gain and equalisation may be applied to the signal after sampling by the ADC and before nonlinear processing, and equalization (tone controls) may be applied to the output of the nonlinear processor before conversion to an analogue signal by the DAC. Low-pass filters may be placed after the symmetric distortion stages, or symmetric distortion stages used which incorporate low-pass filters as discussed in NZ Patent 329119.

[0090] The nonlinear processor is primarily designed for distorting audio signals from electric and electronic instruments such as guitars and keyboards, and other recorded acoustic instruments. However, it will be appreciated that the nonlinear processor may be arranged to distort audio signals generated by any number of different types of sources.

[0091] The foregoing description of the invention includes preferred forms thereof. Modifications may be made thereto without departing from the scope of the invention as defined by the accompanying claims.

1. A nonlinear processor for distorting audio signals, comprising:

an input stage that is arranged to split an audio input signal into two signal paths;

a pair of asymmetric distortion stages following the input stage such that there is one asymmetric distortion stage in each signal path, each asymmetric distortion stage having non-equal negative and positive saturation limits and a smooth transition between linear and nonlinear behaviour, and being arranged to produce a distorted output signal that has a mean signal level that is opposite in polarity to the other asymmetric distortion stage;

a pair of AC-coupled symmetric distortion stages following the asymmetric distortion stages such that there is one symmetric distortion stage in each signal path, each symmetric distortion stage being arranged to nonlinearly limit the distorted signals in each signal path; and

an output stage following the symmetric distortion stages that is arranged to add the two nonlinearly distorted signals from the symmetric distortion stages to generate an audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artifacts.

2. A nonlinear processor according to claim 1 in which the processor is implemented in an analogue circuit wherein the input stage is arranged to receive an analogue audio input signal, buffer the input signal, and split the input signal into two signal paths, and wherein the output stage is arranged as a summer for adding the two analogue nonlinearly distorted signals from the symmetric distortion stages to generate a single analogue audio output signal.

3. A nonlinear processor according to claim 1 in which the processor is implemented in a digital system wherein the input stage comprises an analogue-to-digital converter that is arranged to receive an analogue audio input signal, convert the analogue input signal into a digital input signal, and split the digital input signal into two digital signal paths, and wherein the output stage comprises: a summer that is arranged to add the two digital nonlinearly distorted signals from the symmetric distortion stages to generate a single digital audio output signal; and a digital-to-analogue converter that is arranged to convert the single digital audio output signal into a single analogue audio output signal.

4. A nonlinear processor according to claim 1 wherein the magnitude of the positive and negative saturation limits for one of the asymmetric distortion stages is substantially equal to the magnitude of the negative and positive saturation limits respectively for the other asymmetric distortion stage so as to produce an audio output signal at the output stage that demonstrates a smooth transition from linear behaviour to the production of crossover-like artefacts.

5. A nonlinear processor according to claim 1 wherein the magnitude of one or both of the positive and negative saturation limits for one of the asymmetric distortion stages is different to the magnitude of the negative and positive saturation limits respectively for the other asymmetric distortion stage so as to produce an audio output signal at the output stage that demonstrates a smooth transition from linear behaviour to the production of crossover-like artefacts, with a spectrum which includes even harmonics of input frequencies of the audio input signal.

6. A nonlinear processor according to claim 5 wherein the magnitude of the positive saturation limit for one of the asymmetric distortion stages is substantially higher than the magnitude of the negative saturation limit for the other asymmetric distortion stage.

7. A nonlinear processor according to claim 1 wherein the symmetric distortion stages each comprise a low-pass filter to provide a reduction of harmonic energy when nonlinearly limiting the distorted signals from the asymmetric distortion stages.

8. A nonlinear processor according to claim 1 wherein the audio input signal is from an electric or electronic musical instrument.

9. A multiband nonlinear processor for distorting audio signals, comprising:

an input stage that is arranged to receive an audio input signal:

an equi-phase crossover network that is arranged to split the input signal into two or more frequency bands with finite overlap between the frequency bands, and equal phase responses in each band, and in each frequency band:

an asymmetric distortion stage having non-equal negative and positive saturation limits and a smooth transition from linear to nonlinear behaviour, and where the saturation limits alternate across the frequency bands so as to produce distorted output signals having alternating polarity mean signal levels across the frequency bands; and

an AC-coupled symmetric distortion stage following the asymmetric distortion stage that is arranged to nonlinearly limit the distorted output signal from the asymmetric distortion stage; and

an output stage that is arranged to add the nonlinearly distorted signals from the symmetric distortion stages of all frequency bands to generate an audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artifacts, with a reduction of intermodulation distortion.

10. A multiband nonlinear processor according to claim 9 in which the processor is implemented in an analogue circuit wherein the input stage is arranged to receive an analogue audio input signal and buffer it into the equi-phase crossover network, and wherein the output stage is arranged as a summer for adding the analogue output signals from all the frequency bands to generate a single analogue audio output signal.

11. A multiband nonlinear processor according to claim 9 in which the processor is implemented in a digital system, and wherein the input stage comprises an analogue-to-digital converter that is arranged to receive an analogue audio input signal and convert it into a digital input signal for the equi-phase crossover network, and wherein the output stage comprises: a summer that is arranged to add the digital output signals from all frequency bands to generate a single digital audio output signal; and a digital-to-analogue converter that is arranged to convert the single digital audio output signal into a single analogue audio output signal.

12. A multiband nonlinear processor according to claim 9 wherein the magnitude of the positive and negative saturation limits of each asymmetric distortion stage is substantially equal to the magnitude of the negative and positive saturation limits respectively of adjacent asymmetric distortion stages of adjacent frequency bands so as to produce an audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artifacts, with a reduction of intermodulation distortion.

13. A multiband nonlinear processor according to claim 9 wherein one or both of the positive and negative saturation limits of each asymmetric distortion stage is different to the magnitude of the negative and positive saturation limits respectively of adjacent asymmetric distortion stages of adjacent frequency bands so as to produce an audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artifacts, with a reduction of intermodulation distortion, and with a spectrum which includes even harmonics of the input frequencies of the audio input signal.

14. A multiband nonlinear processor according to claim 9 wherein the symmetric distortion stages each comprise a low-pass filter to provide a reduction of harmonic energy when nonlinearly limiting the distorted signals from the asymmetric distortion stages.

15. A multiband nonlinear processor according to claim 9 further comprising cross-coupling between the frequency bands before the distortion stages to allow the controlled increase of intermodulation distortion.

16. A multiband nonlinear processor according to claim 9 wherein the audio input signal is from an electric or electronic musical instrument.

17. A nonlinear audio distortion circuit for distorting audio signals from musical instruments, comprising:

an input stage that is arranged to split an audio input signal into two signal paths;

a pair of asymmetric distortion stages, one in each signal path, with non-equal negative and positive saturation limits, so as to produce opposite polarity mean signal levels at their outputs in each signal path, and which produce a smooth transition from linear to nonlinear behaviour;

a pair of AC-coupled symmetric distortion stages, one in each signal path, following the asymmetric distortion stages; and

an output stage that is arranged to add the two nonlinearly distorted signals from the symmetric distortion stages to generate an audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artifacts.

18. A nonlinear audio distortion circuit according to claim 17 wherein the saturation limits in the two asymmetric distortion stages are the opposite of each other so as to produce an audio output signal at the output stage that demonstrates a smooth transition from linear behaviour to the production of crossover-like artefacts.

19. A nonlinear audio distortion circuit according to claim 17 wherein the saturation limits of the two asymmetric distortion stages are different to each other so as to produce a final audio output signal that demonstrates a smooth transition from linear behaviour to the production of crossover-like artefacts, with a spectrum which includes even harmonics of the input frequencies of the audio input signal.

20. A nonlinear audio distortion circuit according to claim 17 wherein the symmetric distortion stages each comprise an amplifier with a feedback loop that is arranged to nonlinearly limit the signal of its signal path and a low-pass filter in the feedback loop that is arranged to provide a reduction of harmonic energy when limiting the signal.

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