The TR-808 Cymbal: a Physically-Informed, Circuit-Bendable, Digital Model

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ABSTRACT
We present an analysis of the cymbal voice circuit from a classic analog drum machine, the Roland TR-808 Rhythm Composer. A digital model based on this analysis (implemented in Cycling ‘74’s Gen’) retains the salient features of the original. Developing physical models of the device’s many sub-circuits allows for accurate emulation of circuit-bent modifications (including component substitution, changes to the device’s architecture, and voltage starve)—complicated behavior that is impossible to capture through black-box modeling or structured sampling. This analysis will support circuit-based musical inquiry into the history of analog drum machines and the design of further mods.

1. INTRODUCTION

Despite significant work that has been done on cloning and emulating the TR-808, there is an almost complete lack of published analyses on the circuit.

We have begun to fill this void with [3], which develops a physically-informed, circuit-bendable, digital model of the TR-808’s bass drum voice circuit. In the following paper, we will use related techniques (well-represented in virtual analog literature) to analyze the 808’s cymbal voice circuit.

The goals of this research are to partition the 808’s cymbal circuit into functional blocks, create a physically-informed analysis of each block, model each block in software, and evaluate the results. Throughout, we will pay special attention to developing an analysis in terms of the electrical values of circuit elements.

It is particularly important to analyze the 808 in this way, because of the extant misinformation surrounding the device.

1 previous work includes [2], which offers a qualitative discussion of [1] in the context of imitating classic synthesized cymbal voices with modular synthesizers.

2 For instance: the user manual for the Novation Drum Station (an early rack-mount TR-808/TR-909 emulator) says the cymbal sound is generated by “multiple noise sources,” which is patently false. Some improbable stories about the machine’s design turn out to be true. Don Lewis, an early pioneer of drum machine modification, worked on the design of the 808 and even developed techniques that influenced its voice design [4].

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4 for instance: adding tuning controls to the oscillators, band pass filters, and high pass filters; additional controls to the static envelope generators; or extending the tone controls.

5 for instance: the user manual for the Novation Drum Station (an early rack-mount TR-808/TR-909 emulator) says the cymbal sound is generated by “multiple noise sources,” which is patently false. Some improbable stories about the machine’s design turn out to be true. Don Lewis, an early pioneer of drum machine modification, worked on the design of the 808 and even developed techniques that influenced its voice design [4]. He relates a story from his visit to Roland’s Tokyo offices in the late 70s, where he worked with chief engineer Tadao Kikumoto. “That day he had a bread board of an 808 and was showing me what was going on inside—he sort of bumped up against the breadboard and spilled some tea in there and all of a sudden he turned it on and got this ‘pssh’ sound—it took them months to figure out how to reproduce it, but that ended up being the crash cymbal in the 808. There was nothing else like it. Nobody could touch it.” [5]
were quick to pull these controls out to the front panel. Intended for factory tuning, early drum machine hackers by opening up the TR-808’s internals. Although this was the last two, which form the basis of the Cowbell voice that tunes the oscillator to a particular frequency. The output of these astable multivibrators, the Schmitt trigger inverter acts as the bistable element, and a passive network of a single resistor and capacitor provides an RC time constant of six of these oscillators, ingeniously implemented with a rectangular waves. In fact, they all share a common bank of six of these multimaster oscillators, producing these clusters. The output of each of the three swing-type VCAs is applied to a corresponding Sallen-Key high pass filter. From there, each band is applied to a tone control stage and a level control, which sums the three bands back together and buffers the output.

3. SCHMITT TRIGGER OSCILLATORS

The TR-808’s Cowbell, Cymbal, Open Hihat, and Closed Hihat voice circuits all work by filtering and enveloping rectangular waves. In fact, they all share a common bank of six of these oscillators, ingeniously implemented with a single HD14584 hex Schmitt trigger inverter chip. In each of these astable multivibrators, the Schmitt trigger inverter acts as the bistable element, and a passive network of a single resistor and capacitor provides an RC time constant that tunes the oscillator to a particular frequency. Oscillators #1—4 are hardwired to a particular frequency. The last two, which form the basis of the Cowbell voice circuit, are tunable via trimpots that are only accessible since the aliased components will be sufficiently filtered out or perceptually suppressed methods for rectangular wave simulation (for instance: [9]), when an input voltage \(V_t\) rises above a positive-going threshold voltage \(V_{T^+}\), the output swings to \(V_{OL}\). When \(V_t\) falls below a negative-going threshold voltage \(V_{T^-}\), the output swings to \(V_{OH}\).

Since the positive-going and negative-going threshold voltages for a Schmitt trigger input are different, the circuit will oscillate back and forth as the capacitor is alternatively charged and discharged. Considering the continuous-time cases of a charging and discharging capacitor separately yields the capacitor charge and discharge times:

\[
\begin{align*}
t_{\text{charge}} &= RC \cdot \ln \left( \frac{V_{OH} - V_{T^-}}{V_{OH} - V_{T^+}} \right) \quad (1) \\
t_{\text{discharge}} &= RC \cdot \ln \left( \frac{V_{OL} - V_{T^-}}{V_{OL} - V_{T^+}} \right) \quad (2)
\end{align*}
\]

These times sum to the total period of oscillation \(T = t_{\text{charge}} + t_{\text{discharge}}\). Since the oscillator switches between output states \(V_{OL}\) and \(V_{OH}\), its amplitude is trivially equal to their difference. The duty cycle is, by definition, the proportion of time that the oscillator spends in its high state. Now, the salient features (frequency \(f = \frac{1}{T}\), amplitude \(A = V_{OH} - V_{OL}\), and duty cycle \(D = \frac{t_{\text{charge}}}{T}\)) of the HD14584 are all faster than 250 ns, and the transitions are smooth. Keeping in mind the bandwidth of human hearing (roughly, 20–20,000 hertz, corresponding to a minimum period of 50 \(\mu s\), this transition is treated as instantaneous. We forego alias-suppressed methods for rectangular wave simulation (for instance: [9]), since the aliased components will be sufficiently filtered out or perceptually masked downstream.

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\[\text{Figure 1. TR-808 cymbal schematic (adapted from [1]).}\]
able in terms of passive component values and the device properties of the Schmitt trigger inverter.

Plugging in Equations (1)–(2) to the definition of duty cycle, the time constant RC drops out—the duty cycle depends only on the inverter’s device properties $V_T^-$, $V_T^+$, $V_{OL}$, and $V_{OH}$:

$$D = \frac{\ln \left( \frac{V_{OH} - V_T^+}{V_{OH} - V_T^-} \right)}{\ln \left( \frac{V_{OL} - V_T^+}{V_{OL} - V_T^-} \right)}. \quad (3)$$

The six oscillators have nominal frequencies (or ranges) of 205.3, 369.6, 304.4, 522.7, 359.4–1149.9, and 254.3–627.2 Hz. Oscillators #5–6 have internal trimpots (TM2 and TM1) in series with their resistor, for factory tuning to specific frequencies (800 and 540 Hz, respectively).

By design, the HD14584 is powered with 5 volts, yielding a duty cycle of $D = 47.98\%$ and an amplitude of $A = 5$ volts for each oscillator.

All six oscillators are summed via voltage dividers in a passive mixing network. The output of this network is applied to the input of the band pass filters. Respecting superposition (grounding the other oscillator outputs), an example calculation of this attenuation (for oscillator #1) is:

$$V_{sum} = \frac{V_{O_{LO}}}{R_{O_{L}}} + \frac{R_{S}}{R_{O_{L}} + (R_{37} || R_{39} || R_{40} || R_{48} || R_{50})}. \quad (4)$$

Fig. 3 shows a comparison between the sum of all six oscillators (the signal applied to each band pass filter) and tabulated data from a SPICE simulation of the circuit.

9 Early manufacturing runs (prior to serial #000300) used different values for the resistors and capacitors in oscillators 5 and 6 [1].

10 Using the “typical” values from the device’s datasheet [8]

11 Simulation Program with Integrated Circuit Emphasis, a widely-used family of general purpose analog electronic circuit simulators.

12 No attempt was made to align the phase of the model to the tabulated SPICE data. Fig. 3 just shows, qualitatively, that the model and the reference SPICE simulation have similar features.

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**4. BAND PASS FILTERS**

The two active band pass filters each filter the combination of the six rectangular waves. What follows is a representative analysis and simulation of band pass filter #1. Identical methods were used for #2.

Assuming ideal op-amp behavior, nodal analysis of band pass filter #1 yields a transfer function of the form:

$$H_{bp1}(s) = \frac{V_{bp1}}{V_{sum}} = \frac{\beta_2 s^2 + \beta_1 s}{\alpha_3 s^3 + \alpha_2 s^2 + \alpha_1 s + \alpha_0}, \quad (5)$$

with coefficients:

$$\beta_2 = -R_{56} R_{57} (C_{13} + C_{14}) C_{10}$$
$$\beta_1 = -R_{57} C_{10}$$
$$\alpha_3 = R_{98} R_{97} R_{92} C_{13} C_{14} C_{10}$$
$$\alpha_2 = R_{98} R_{97} C_{13} C_{14} + R_{98} R_{52} (C_{13} + C_{14}) C_{10}$$
$$\alpha_1 = R_{98} C_{13} + R_{56} C_{14} + R_{52} C_{10}$$
$$\alpha_0 = 1 .$$

Fig. 4 shows the magnitude response of each band pass filter. Band pass filter #1 has a center frequency around 3440 Hz and band pass filter #2 has a center frequency around 7100 Hz. These band pass filters strongly accentuate the upper overtones of the square waves, while de-emphasizing their fundamental frequencies.

13 whose topologies are closely related to the so-called “bridged-T network”, a bridged-T Zobel network in the negative-feedback path of an op-amp, that is used in each of the 808’s voice circuits [3]
The circuit comprised of VCAs.

1.4 The attack smoother, envelope generators, and the swing-type filters.

1.8 The effect of the varying accent level is complex, and the mon trigger signal and accent signal, which are combined into a composite trigger signal and accent signal, which are combined into a composite.

2 The 808's sequencer controls the timing and amplitude of each voice circuit via the CPU. The CPU produces a timing signal and accent signal, which are combined into a common trigger signal $V_{trig}$, whose ON voltage is set by $VR_3$, a user-controllable global accent level. In general, instrument timing data (unique to each voice) sequenced by the CPU is “ANDed” with $V_{id}$ to activate individual voice circuits.

5. TRIGGER LOGIC

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In the case of the cymbal, the circuit comprised of $Q_2$, $Q_6$, $R_9$, $R_4$, $R_{10}$, $R_{15}$, and $R_{58}$ “ANDs” the instrument data $V_{id}$ with a buffered version of $V_{id}$ (with its range narrowed to 7–14 V, depending on $VR_3$ [1]). When $V_{id}$ is present (logic high), a 1-ms long pulse is passed to the collector of $Q_6$.

The effect of the varying accent level is complex, and depends heavily on nonlinear downstream subcircuits: the attack smoother, envelope generators, and the swing-type VCAs.

6. ATTACK SMOOTHER

The circuit comprised of $Q_{19}$, $R_{44}$–$R_{46}$, and $C_{45}$ “smooths” the attack of $V_{trig}$, the pulse that the trigger logic produces. Although the physics of this sub-circuit are difficult to work out in closed form, a one pole filter of the following form approximates the behavior of this circuit:

$$H_{as}(s) = \frac{V_{smooth}}{V_{trig} - V_{BE}} = \frac{1}{1 + \tau s} ,$$

where $V_{BE}$ is the voltage drop from the base of $Q_{19}$ to its emitter, and $\tau$ is an RC time constant. The values of these constants are found by least-squares minimization to be:

$$V_{BE} = 0.7258, \quad \tau = 1.0244 \cdot 10^{-4} .$$

Fig. 6 shows good agreement between this simplified model and tabulated data from SPICE.

7. ENVELOPE GENERATORS

The attack smoother applies a smoothed trigger $V_{smooth}$ to three envelope generators via $D_6$–$D_8$. These envelopes serve as amplitude control for each swing-type VCA. What follows is a representative analysis and simulation of envelope generator #1. Identical methods were used for #2–3.

In envelope generator #1, an emitter follower buffers the attack of $Q_1$, whose ON voltage is set by $VR_3$, and $C_{41}$, an identical diode model [14] is leveraged to separate the attack and release phases, allowing simulation by numerical solution of a switched first-order ordinary differential equation (ODE).

During the attack phase (while $V_B < V_{smooth} - V_{on}$, and $D_6$ acts as a short), an equation that approximates $V_B$ is:

$$V_B = V_{in} - V_{on} .$$

During the release phase (while $V_B > V_{smooth}$, and $D_6$ acts as an open circuit), nodal analysis yields:

$$\frac{dV_B}{dt} = -\frac{V_B}{VR_2 k R_{53} C_{41}} ,$$

where $VR_2 k$ is the resistance of the decay control with maximum resistance $VR_2$ and knob position $k \in [0,0.1,1]$. At higher decay settings, the RC time constant of decay increases, and the decay for this band is longer.

$V_{on}$ is estimated in the least-squares sense by a fit to tabulated data from a SPICE simulation during the attack portion on the envelope. This yields $V_{on} = 0.5899$.

14 One that assumes the diode spends little time around the corner of its I–V characteristic and can be treated as a switch:

$$\frac{R_{diode}}{\infty} \text{ if } V_{diode} < V_{on}$$

$$0 \text{ if } V_{diode} \geq V_{on}$$

†
8. SWING-TYPE VCAS

So-called swing-type VCAs perform nonlinear voltage-controlled amplification. They resemble common-emitter amplifiers, despite their non-standard biasing scheme, extra diode, and the fact the $V_{CC}$ has been replaced by an applied envelope voltage. What follows is a representative analysis and simulation of swing-type VCA #1. Similar methods were used for #2–3. An analysis and close approximation to the behavior of a swing-type VCA is found by framing it as a modified common-emitter amplifier, working through its biasing details, performing a large-signal DC analysis when the transistor is in forward-active operation, and estimating a memoryless nonlinearity to approximate the behavior of the diode and the transistor’s saturation region operation.

8.1 VCA biasing

$R_{102}$–$R_{103}$ and $C_{46}$ form a DC blocking high-pass filter, with an offset $V_B$ (which is estimated like before).

This kind of biasing scheme is inherently unstable to temperature change, and depends strongly on the Bipolar Junction Transistor’s (BJT) $\beta$ parameter, which varies significantly from component to component. For this reason, this sort of biasing is usually discouraged [10]. Here, it leads to clipping—a wild swing of the output between $V_{EG\#1.2}$ and $V_{lower\,Edge}$, apparent intended behavior.

8.2 Large-Signal DC Analysis

A BJT in forward-active operation is partially described by an equation (an approximation to the Ebers-Moll model) relating the emitter current $i_E$ to the voltage drop from the base to the emitter, $V_{BE}$:

$$i_E = I_{ES} \left( \frac{V_{BE}}{V_T} - 1 \right),$$

where $I_{ES}$ is the reverse saturation current of the base–emitter p–n junction and $V_T$ is the thermal voltage. Since $i_E$ is much smaller than the others, it is reasonable to approximate $i_C = i_E$ via KCL. Now, ignoring output loading, we can get a naive expression for $V_{out}$ by finding the voltage drop across the collector resistor $R_{104}$:

$$V_{estimated\#1} = V_{EG\#1.2} - R_{104} I_{ES} \left( \frac{V_{BE}}{V_T} - 1 \right).$$

8.3 VCA Nonlinearity

The diode gates the output when $V_{EG\#1.2}$ is low enough. Without the diode, the output of the VCA would never shut all the way off. The approximation in Equation (13) is fairly accurate for negative values of $V_{BE}$. However, it takes neither the gating action of the diode nor the fact that the transistor can operate in the saturation region into account. These two behaviors are approximated by estimating the lower edge of the output as a function of $V_{EG\#1.2}$ and calculating the output voltage $V_{VCA\#1}$ as a nonlinear function of $V_{EG\#1.2}, V_{lower\,Edge}$, and $V_{estimated\#1}$.

Unsurprisingly (due to the presence of two p–n junctions), $V_{lower\,Edge}$ is well-approximated by the sum of two stretched exponential functions and an offset:

$$V_{lower\,Edge} = \alpha_0 e^{-\beta_0 V_{EG\#1} z^{\gamma_0}} + \alpha_1 e^{-\beta_1 V_{EG\#2} z^{\gamma_1}} + \alpha_2.$$

Curve-fitting yields the coefficients shown in Table 1. The clipping behavior of the diode and the transistor’s behavior in its saturation region are approximated by a nonlinear equation of $V_{env}, V_{lower\,Edge}$ and $V_{estimated\#1}$:

$$V_{VCA\#1} = \frac{V_{estimated\#1} - V_{env}}{\left( 1 + \frac{V_{estimated\#1} - V_{env}}{V_{lower\,Edge}} \right)^{\frac{1}{\alpha}}} + V_{env}.$$

15 $x (n+1) = x (n) + T \cdot f (x (n))$

16 This odd name is given by [1]. [8.1] should offer a clue to its meaning.

17 For swing-type VCA #2, there are two envelope sources (EG #1.2 and EG #2). For swing-type VCAs #2–3, there are additional capacitors to ground around the diodes, so analysis is complicated slightly, requiring the use of ODEs.
This is a modified version of the clipping equation from [11]. The parameter $\alpha$ controls the “sharpness” of the function’s corners. A value of $\alpha = 3.5^{21}$ gives a good visual fit to a tabulated SPICE simulation. Fig. 8 shows VCA #1’s behavior in the time domain.

9. HIGH PASS FILTERS

What follows is a representative analysis and simulation of high pass filter #1. Identical methods were used for #2–3.

High pass filter #1 ($H_{h1}$) is a 2nd-order high pass Sallen-Key filter [14] with an emitter follower ($Q_{25}$) as its buffer amplifier element. Assuming the emitter follower acts as a perfect buffer, this yields a continuous-time transfer function [15]:

$$H_{h1}(s) = \frac{V_{h1}}{V_{VCA#1}} = \frac{\beta_2s^2}{\alpha_2s^2 + \alpha_1s + \alpha_0}, \quad (16)$$

with coefficients:

$$\beta_2 = R_{124}R_{127}C_{48}C_78$$
$$\alpha_2 = R_{124}R_{127}C_{48}C_78$$
$$\alpha_1 = R_{124}C_{48} + R_{124}C_{50}$$
$$\alpha_0 = 1.$$

High pass filter #2 ($H_{h2}$) is a 2nd-order, high pass, non-unity-gain Sallen-Key filter with an op-amp as its buffer amplifier element.

High pass filter #3 ($H_{h3}$) is a 3rd-order high pass Sallen-Key filter with an op-amp as its buffer amplifier element. Using an op-amp in place of an emitter-follower allows for the design of non-unity-gain buffering. This manifests as resonance at the corner frequency (around 10500 Hz) in the magnitude response.

Fig. 4 shows the magnitude responses of the three high pass filters.

10. TONE STAGE

The tone stage is a highly-interconnected passive network of resistors and capacitors, which can be described by three transfer functions, $H_{t1} = V_{tone}/V_{h1}$, $H_{t2} = V_{tone}/V_{h2}$, and $H_{t3} = V_{tone}/V_{h3}$. These transfer functions are fifth order, and the expressions for their filter coefficients are far too lengthy to print in this work or to provide much insight by visual inspection. They are reported on this work’s companion site. 23

Fig. 9 shows a family of magnitude responses for each of the three bands. The primary effect of the tone control is to change the amount of attenuation in the third band. However, the center frequencies of all bands and the attenuation of the first and second bands are also somewhat affected. The tone control of the TR-808 cymbal voice shares the property of weakly-separated, non-orthogonal controls with guitar amplifier tone stacks [16].

11. OUTPUT BUFFER AND LEVEL CONTROL

The output buffer sums together the three cymbal bands, through the tone stage, and offers level control. Assuming an ideal op-amp, there will be no current through, and therefore no voltage across, $R_{118}$. Nodal analysis yields a transfer function of the form:

$$H_{le}(s) = \frac{V_{out}}{V_{r1}} = \frac{\beta_2s^2 + \beta_1s}{\alpha_2s^2 + \alpha_1s + \alpha_0}, \quad (17)$$

with coefficients:

$$\beta_2 = -R_{128}VR_{0}kC_{90}C_78$$
$$\beta_1 = -R_{128}C_{90}$$
$$\alpha_2 = R_{128}VR_{0}kC_{77}C_78$$
$$\alpha_1 = (R_{128} + VR_{0})C_78$$
$$\alpha_0 = 1.$$
the output level control buffer also acts as a differentiator in the audio band—it has a frequency response with a 6 dB/octave rising slope.

12. MODELING

We implemented a digital model with Cycling 74’s Gen−, a low-level DSP environment in Max/MSP. The model contains stock controls for the cymbal decay, tone, and level. Additionally, the model contains bends for tuning and muting individual Schmitt trigger oscillators, tuning or bypassing each band pass and high pass filter, and controlling the release of envelope generators #1 and #3. One bend that has a surprising effect on the sound is replacing the level control buffer also acts as a differentiator in the audio band—it has a frequency response with a 6 dB/octave rising slope. The level control magnitude responses with level control $k \in [0.01, 1.0]$. The $k = 1.0$ response is marked with an asterisk ($\ast$).

The family of level control magnitude responses is shown in Figure 10.

![Figure 10](https://ccrma.stanford.edu/~kwerner/papers/icmcsmc2014.html)

**Figure 10.** Family of level control magnitude responses with level control $k \in [0.01, 1.0]$. The $k = 1.0$ response is marked with an asterisk ($\ast$).

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13. RESULTS AND CONCLUSIONS

Fig. 11(a) is a spectrogram/waveform pair of a reference cymbal note, produced via SPICE simulation. Fig. 11(b) is the same output from the physical model, which shows good agreement with the SPICE simulation.

Figs. 11(c)–11(f) show some of the bends that are available in the model. These bends can dramatically alter the cymbal’s timbre and texture.

Fig. 11(d) shows the effect of lowering the values of resistors $R_{56}$ and $R_{58}$ by a factor of one half. This raises their center frequencies, affects their Qs, and generally eliminates much of the low-frequency content.

Fig. 11(e) shows the effect of lowering $R_{53}$. This affects the level of the signal going into the band pass filters, and causes the signal to only sporadically clip in the VCAs—this has an effect on the texture of the output.

Fig. 11(f) shows the effect of a 50% voltage starve. This technique is common in circuit bending and guitar pedal modification but remains unexplored in the context of the TR-808. Lowering the voltage supplied to the TR-808’s cymbal voice circuit have response that are high in frequency with respect to normal audio sampling rates (i.e. $f_s = 44100$ Hz), this frequency warping must be addressed. All examples in this work are rendered at 4× oversampling ($f_s = 44100 \times 4$ Hz), so they suffer only negligibly from frequency warping.

There will be some bandwidth expansion in the VCA stage, due to its nonlinearity. Although this creates the potential for aliasing, the cymbal circuit creates tightly-packed, inharmonically-related partials by design. As well, some of the aliasing will be mitigated by the oversampling. Unsurprisingly, the presence of low amplitude aliased frequency components seem to be perceptually insignificant.

Some parts of the model are oversimplified. The VCAs and envelope generators, in particular, are modeled with significant simplifications, and some of their nonlinearities ignored. Non-idealities in high pass filter #1’s emitter-follower buffer are ignored. To break the delay-free loops arising from mutual interaction between the VCAs and envelope generators, fictitious unit delays are inserted. Since the envelope signals change very slowly with respect to the sampling rate, the error that is introduced by the unit delays is not very significant. However, in general, inserting fictitious unit delays into systems with feedback can have negative effects on accuracy and stability.

Acknowledgments

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25 see [19] for a DIY example, and Danelectro’s Danelectrode for a commercial example.
26 although it is hinted at in [7]
27 A reasonable estimate of these properties as a function of supply voltages in the range 3–15 V is obtained by extrapolation of the electrical characteristics at 5, 10, and 15 volts [8].

4 https://ccrma.stanford.edu/~jos/tp/Frequency_Warping.html
Figure 11. Waveform/spectrogram pairs of cymbal voice circuit simulations: (a) a baseline SPICE simulation for comparison, (b) a baseline emulation with the physically-informed model, (c)muting 3 of the rectangular wave oscillators and tuning the remaining 3 to an A major chord, (d) altering the band pass filter responses by lowering $R_{56}$ and $R_{58}$, (e) effect of lowering $R_{53}$, and (f) voltage starve. All are rendered at $4\times$ oversampling, with the decay knob at 25%, tone knob at 50%, and level knob at 50%.

14. REFERENCES