Build Advanced Effects for Electric Guitar & Bass

the StompBox Cookbook

2.733"   1.513"

Funk   Punk   Jazz

Rock   Roll

OUTPUT   INPUT

OUTPUT

4.075"

1.879"

2.196"

1.963"

Stomp • Matic
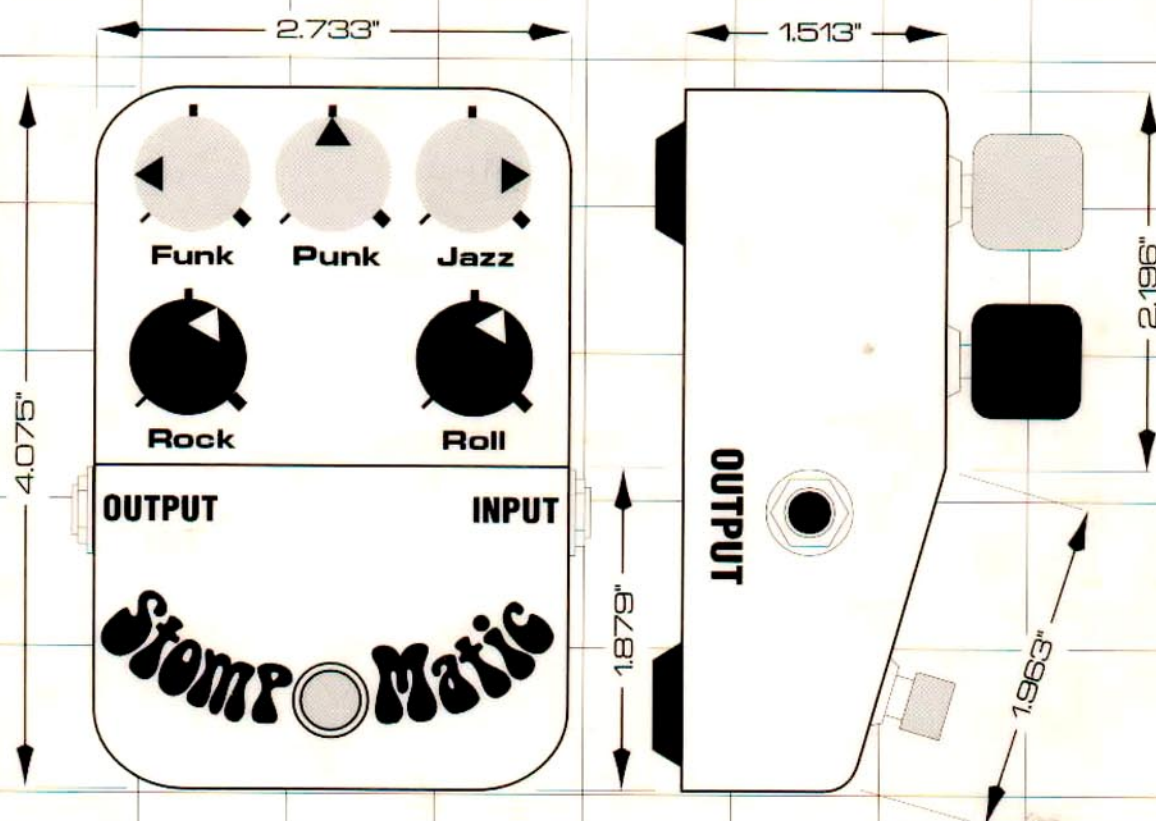
Nicholas Boscorelli

the StompBox Cookbook

2nd Edition

The Stomp Box Cookbook, Second Edition
Build Advanced Effects for Electric Guitar & Bass
by Nicholas Boscorelli

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Appendices: Stomp Box Ingredients

PREFACE

Given the vast array of commercial stomp-box effects, the reader might wonder why anyone believes that we need more. The author sees this panoply as a hunger for new sound. Unfilled niches remain. And good as they are, canned effects live with limits imposed by retail viability. Building from scratch breaks those limits, and that's what this book is about.

The text is written at the level of those who understand basic electronics, and who have mastered the skills needed to build intermediate to advanced projects. Writing for an adept audience means not having to reinvent the wheel. Beginners would profit from building simpler projects than these. The References list books that brief readers from a standing start.

The text offers each project as a paradigm of the possible. The Appendices detail stomp-box 'ingredients' to help the builder create effects unavailable commercially, or unavailable in a specific form desired. Rolling your own saves money, and can take a journeyman builder to the next level, besides being loads of fun.

Good luck—and good stomping!

DISCLAIMER

The author and the publisher presume the reader to be proficient in skills needed to build intermediate to advanced electronic projects. These include, but are not limited to: wiring, soldering, fabricating printed circuit boards, and drilling and cutting common materials. All instructions are offered on the presumption that the builder can carry them out competently and safely. Anyone who is not absolutely certain that he or she can build projects competently and safely is warned not to attempt to build any project.

The author has vetted information presented herein, to the extent of his having built functional prototypes largely according to the instructions given. Nevertheless, the reader is warned that the text may contain errors, both typographic and otherwise. The author and the publisher disclaim any and all liability and responsibility to any party for damage or loss caused, or alleged to be caused, directly or consequentially, by the use, improper use, negligent use, or inability to use the information contained herein; or caused, or alleged to be caused, by an omission of information from this book. The author and the publisher make no assurances or guarantees as to any project's performance, safety, or suitability to a particular application, nor should the reader infer that such assurances or guarantees have been made. The author and the publisher present this book "as-is," and without expressed or implied warranty of any kind.

AUTHOR'S NOTE

The response to the First Edition of this book confirmed a hunch that many stomp-box builders were ready for afterburners. It also made clear that much remained unsaid. This revision fills out the card with techniques whose simplicity belies their power. A soft-knee compressor with auto-variable attack and decay goes together as easily as a fully manual box.

The text hasn't sought to duplicate specific pedals, classic or not, though keys to that avenue abound. The point of the exercise is to take control of your sound, and to create new sounds. Most analog effects lie within the builder's grasp, including many that have no counterpart among commercial products.

To all who sent comments and questions: Thanks. I appreciate your input. This volume answers many of your questions. I will try to answer letters, if brief and to the point, and accompanied by an addressed, stamped envelope.

For obvious reasons I cannot troubleshoot anybody's projects. As a concession to those who build these boxes sooner than they should, this volume outlines the procedure I follow when a board fails to work.

Nick Boscorelli

General Considerations

Project discussions cover only the essentials needed to build and use each box. The principles underlying design are discussed at length in the Appendices.

All resistors are ⅛W or ¼W 5% carbon film types unless otherwise specified. Capacitors' rated working voltages must exceed the greatest AC or DC voltage to which they will be subject. This holds particularly for power supply bypass capacitors.

Power feed indications given on schematics and layout diagrams show only connections to V+, V−, and ground. The necessary power switch connections are shown below.

Circuit descriptions do not list individual passive components that form a larger functional block; rather, they refer to, for example, "a precision fullwave rectifier made up of IC1-b & -c, and associated components." The *associated components* are the resistors, capacitors, or diodes connected to IC1-b and -c. Nor do circuit descriptions make note of power supply bypass capacitors, stray-capacitance compensation caps, or RF shunt caps.

Unless otherwise specified, function of each box is described with it running off a pair of freshly charged nicad '9V' batteries. Battery-powered projects incorporate polarity protection diodes into the power feeds.
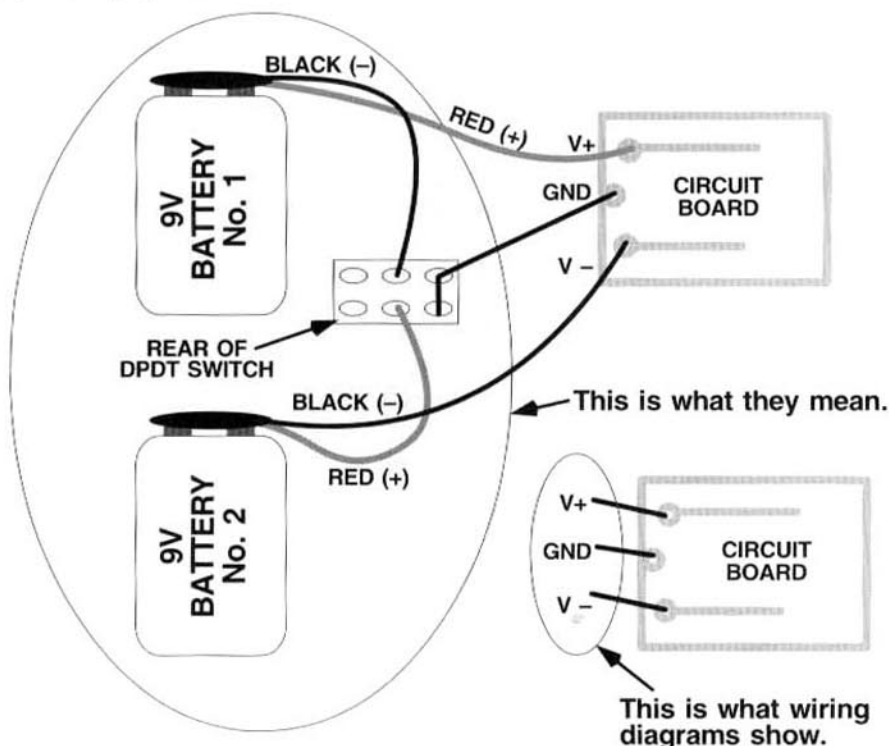
Brief wrong-way application of power, as might occur during battery replacement with the power switch ON, will destroy some semiconductors. The diodes prevent this, but will not protect parts that have been wired incorrectly.

Several projects specify reverse-audio–taper (RA) potentiometers because that taper gives best control in those circuits. Unfortunately, RA pots can hard to find in hobby channels. Substituting a linear-taper pot for the RA pot is generally satisfactory, with some loss of fine control at the clockwise extreme. To regain fine control, the builder can substitute an audio-taper pot for the RA pot, but reverse the connections to the end terminals of the pot. This option also reverses the direction of knob rotation from that which held for the RA pot.

No attempt has been made to implement noiseless switching. While some players expect this feature in commercial boxes, it adds at least one chip to the circuit. The projects use a hardwired bypass that removes the circuit completely from the signal path.

The level-dependent nature of certain effects means they respond differently to instruments having disparate output levels. The builder may have to alter the preamp gain to accommodate different levels.

To minimize clutter on wiring diagrams, power switch and battery connections are not shown. Use this wiring arrangement for dual-9V-battery-powered projects.

BLACK (−)
RED (+)
V+
GND
CIRCUIT BOARD
V −
9V BATTERY No. 1
REAR OF DPDT SWITCH
BLACK (−)
RED (+)
9V BATTERY No. 2

◄— This is what they mean.

V+
GND
CIRCUIT BOARD
V −

This is what wiring diagrams show.

This simple step is discussed in an Appendix.

The presence of patented elements in these projects, whether introduced intentionally or not, does not relieve anyone of obligations to patent holders.

Circuit Board Patterns

Printed circuit templates given in this book are shown full size and are laid out for easy duplication using rub-on pads and tape. Each board appears inside a box with labeled dimensions, to let the builder compensate for distortion that might have accompanied the printing process. The copyright holder grants individual builders permission to copy the printed circuit patterns to build boards for their own personal, non-commercial use. The builder can, of course, build projects on any platform, such as perfboard or universal printed circuit boards.

Board layout and wiring diagrams appear as X-ray views in which the foil side of the board is visible in gray, to aid orientation of parts and identification of wiring points.

Most boards are designed to fit a traditional stompbox–style housing, but the builder should not feel bound to this enclosure. The projects work equally well when mounted in small hobby boxes or rackmount cases.

Chips appearing in prototype photos are those that happened to be socketed at the time of the photo, not necessarily those given in the parts list.

Necessary Skills & Hardware

The presentation assumes that the reader has considerable experience building electronic projects, as well as access to the necessary construction tools and electronic test gear. The novice attempting these projects is courting frustration.

Assembly and testing require a digital multimeter. Several boxes are difficult to set up without a signal generator and oscilloscope; serious troubleshooting demands a scope.

Initial Checkout

Set volume low when connecting a box to an amp for the first time. This avoids cacophony in case a malfunction causes an unusually large output signal.

The player who does not find "the sound" straight off should feel neither frustrated nor surprised. Most projects demand familiarization to live up to their potential. Many feature interactive controls. Patient experimentation is the best way to find the useful control settings.

Project No. 1
Sustain-O-Matic

Downward sustain, with variable upward sustain, suitable for guitar and bass.

Circuit Function

Signal Path: Instrument feed couples through C2 to IC1-a, a preamp with gain of 3.7. IC1-a output couples through R4 to IC2-a, an op amp configured as an inverting amplifier whose voltage gain can vary from +22 dB to –35 dB, depending on the state of the light-dependent resistor (LDR) in the feedback loop, and the setting of pot R6.

IC2-a output couples through R7 to IC2-b, an inverting amp whose voltage gain is variable from 0 to 5 by R9. Signal couples through R11 & C8 to the output path.

Control Path: IC2-a output couples through R8 to input of IC1-b, an inverting amp with gain variable 0.20–20 by R16. IC1-b output couples through C11 to a precision fullwave rectifier made up of IC2-c & -d, and their associated components. The output polarity is positive and couples through R10 to the anode of the LED potted with the LDR in the CLM6000. The result of the control path is for a rise in signal level to generate a voltage that lights the LED acting on the LDR, lowering its resistance and reducing gain of IC2-a. When signal level drops, the LED dims, and LDR resistance rises, raising gain of IC2-a, that gain being limited by the setting of R6.

Use

Switches and pots have these functions:

S1   sustain/bypass
R6   upward compression limit, 0–22 dB
R9   output level
R16   sustain threshold (~70 mv$_{p-p}$ to ~4V$_{p-p}$)

Fig. 1–1. Sustain-O-Matic prototype board.
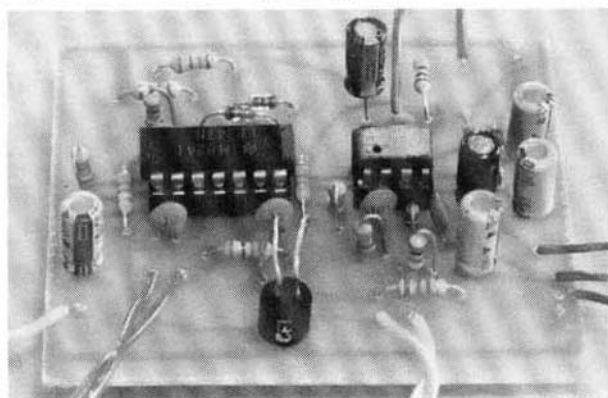
Initial settings: S1 sustain, R6 fully CCW, R9 straight up, R16 fully CW. In this state the box acts as a preamp with gain of about 3.7. Connect the unit to instrument and amp, establish desired volume.

Turning R16 CCW lowers the downward compression threshold. The unit's compression ratio is high enough to qualify as limiting; change R10 to 10K or 15K to reduce the ratio. Increase output level if necessary to bring compressed level back up to match that of the unprocessed feed. This mode sustains the note by reducing the amount of gain reduction as the note decays.

Strike and hold a note, let it decay past the point of downward compression. Slowly turn R6 CW and note the boost applied.

The boost limit can be raised to +27 or +33 dB by changing R6 to 500K or 1M, respectively, at the cost of runaway gain in the absence of an input signal.

Sustain-O-Matic takes practice to apply to full advantage. Avoid the temptation to turn everything up to max. The sweet spots are between 2 and 8.

SUSTAIN-O-MATIC PARTS LIST

Resistors
R1 150K
R2, 17 1K
R3 2.7K
R4, 5 22K
R6 250K audio-taper pot
R7, 12, 13, 14, 15 10K
R8, 10 4.7K
R9 50K audio-taper pot
R11 100
R16 50K reverse-audio pot

Capacitors
C1, 5, 8, 13 10μF electrolytic
C2 10μF nonpolar electrolytic
C3, 6, 7, 12 10pF
C4, 9, 10 100pF
C11 1μF electrolytic

Semiconductors
D1, 2 1N914 diode
D3, 4 1N4001 diode
IC1 TL072 dual op amp
IC2 TL074 quad op amp

Miscellaneous
CLM6000 optocoupler
S1 DPDT switch
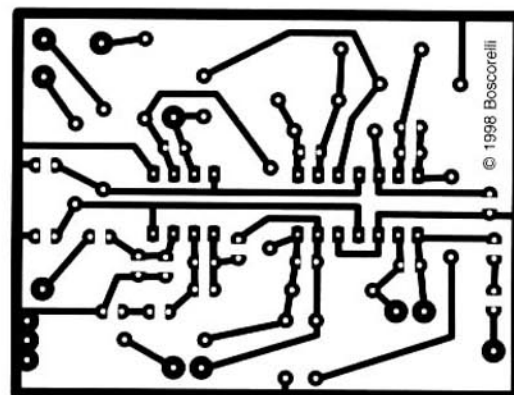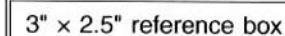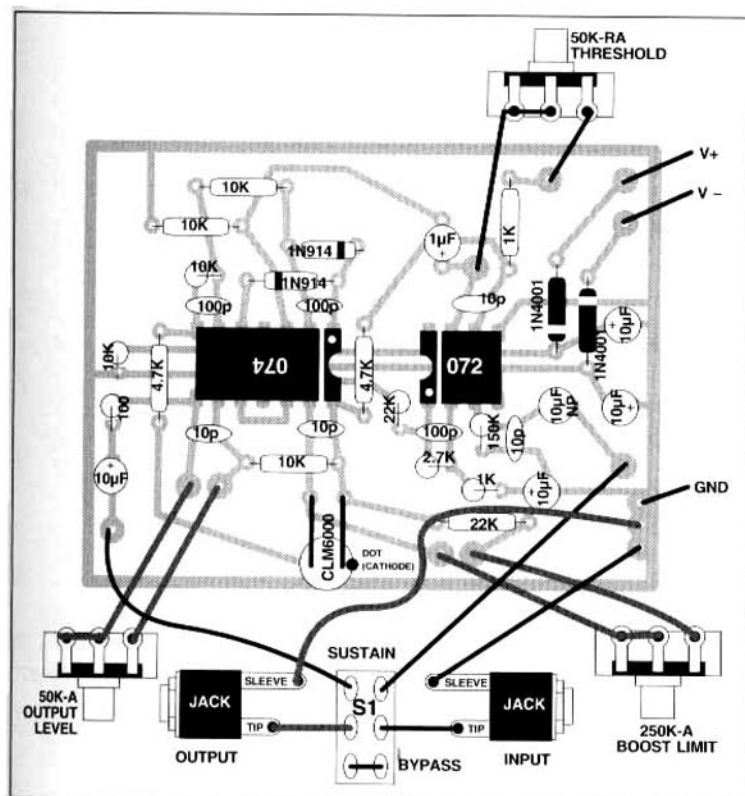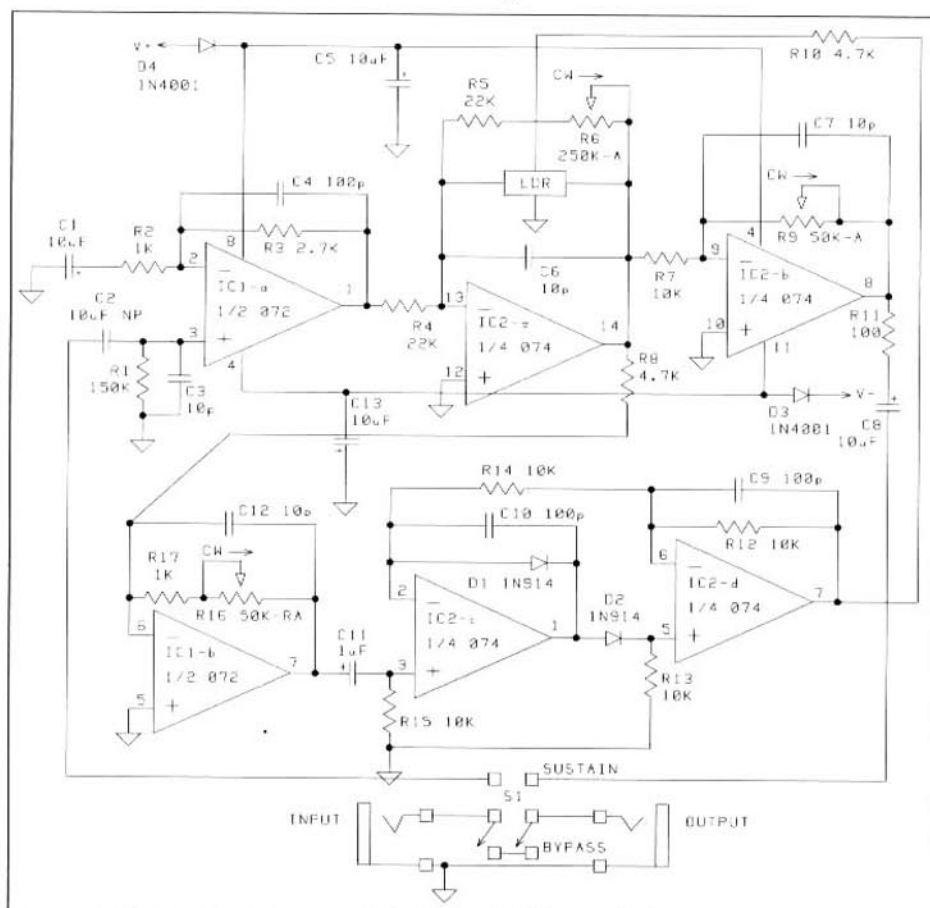enclosure, ¼" jacks, wire, knobs, mounting hardware, 9V batteries, etc.

Fig. 1–3. Sustain-O-Matic layout & wiring diagram.

Fig. 1–2. Sustain-O-Matic circuit board.

Fig. 1–4. Sustain-O-Matic schematic.

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Project No. 2
Distort-O-Matic I

Enough distortion modes exist to build dozens of boxes, each giving a singular sound. Distort-O-Matic I combines a full squarewave fuzz with a soft effect whose transfer function, at some settings, resembles tube-stage overload. Clean and distorted feeds can mix in any ratio.

Circuit Function

Instrument feed couples through C7 to input of IC1-a, an op amp configured as a noninverting amp whose gain is variable from 1–11 by R3.

IC1-a output couples through divider R7-8 to R9, to input of summing amp IC1-b. Signal couples through R12-C5 to the output path.

IC1-a output also couples through R4 & C4 to input of IC2-a, a CMOS hex inverting buffer. IC2-a input is biased through R5 and pot R6. IC1-a output feeds through two more buffers that square up the signal. S2 selects output from pin 2 (passage through a single buffer; 'soft') or pin 6 (passage through two additional buffers; 'harsh').

S2 center terminal couples through C6-R14 to IC3, an inverting amp whose gain is variable from 0–1.5 by pot R13. IC3 output couples through R11 to summing amplifier IC1-b.

Use

Switches and pots have these functions:

S1 distort/bypass
S2 select soft/harsh distortion
R3 preamp gain
R6 hex buffer bias (distortion threshold)
R8 clean level
R13 distortion level

Initial settings: S1 distort, S2 soft; R3, R8 fully CW, R13 fully CCW, R6 straight up. In this configuration the box acts as a clean preamp with gain of about 11. Connect unit to instrument and amp, establish desired volume.

Kill the clean feed by turning R8 fully CCW. Advance R13 a few degrees, just enough to hear the distorted feed when R6 hits the "sweet spot," which is only 10–15° wide. Once the distorted feed is obtained, raise and lower preamp gain and note the effect on tone. Lower the distorted output level before toggling S2 to the harsh setting, to avoid an abrupt jump in volume. Simple but versatile box.

Fig. 2–1. Distort-O-Matic I prototype board.

DISTORT-O-MATIC I PARTS LIST
Resistors
R1 1K
R2 150K
R3, 8 10K audio-taper pot
R4, 7, 10 10K
R5, 11 22K
R6 10K pot
R9 4.7K
R12 100
R13 50K audio-taper pot
R14 33K
Capacitors
C1, 5, 8, 9 10µF electrolytic
C2, 11 10pF
C3, 10 100pF
C4, 6 0.1µF
C7 10µF nonpolar electrolytic
Semiconductors
D1, 2 1N4001 diode
IC1 TL072 dual op amp
IC2 4069 hex inverting buffer
IC3 TL071 op amp
Miscellaneous
S1 DPDT switch
S2 SPDT switch
1/4" jacks, enclosure, wire, 9V battery snaps, knobs, etc.
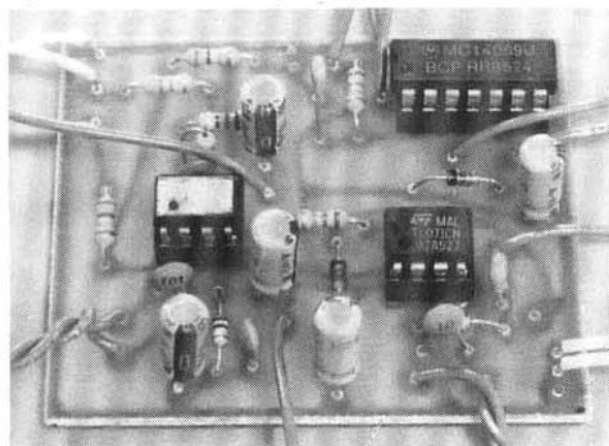
Fig. 2–2. Distort-O-Matic I schematic.
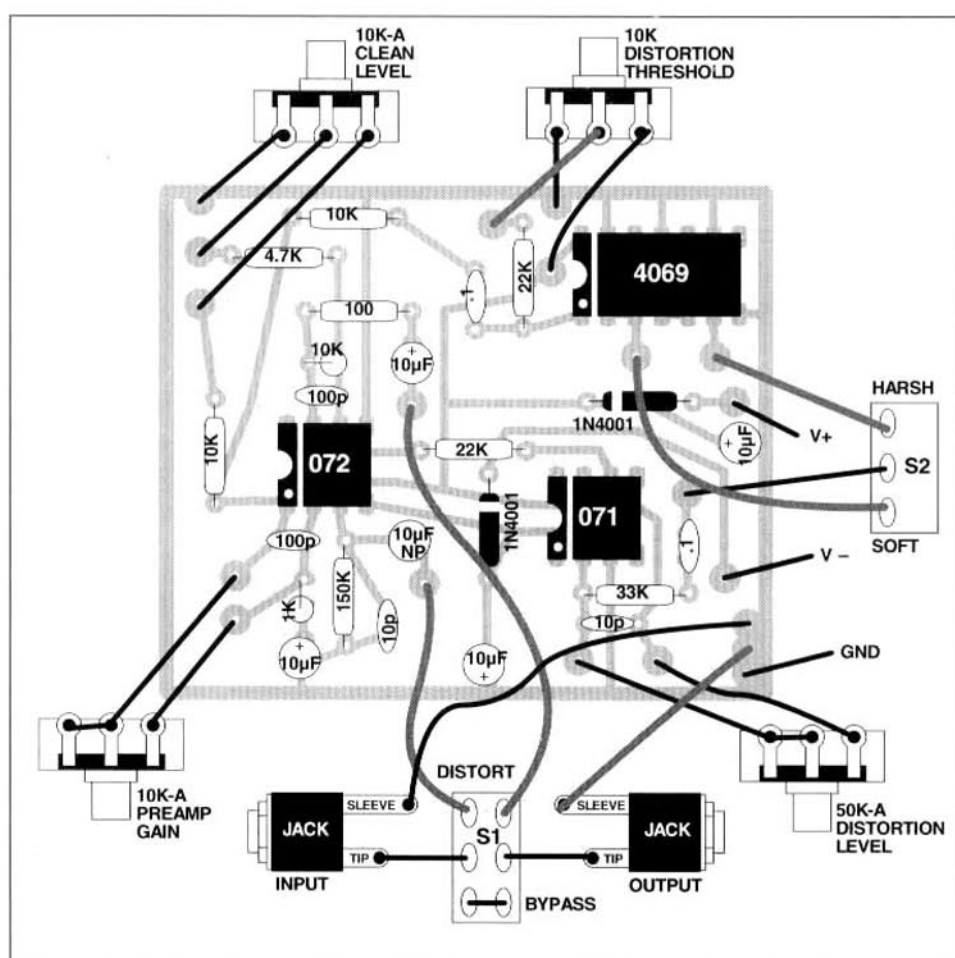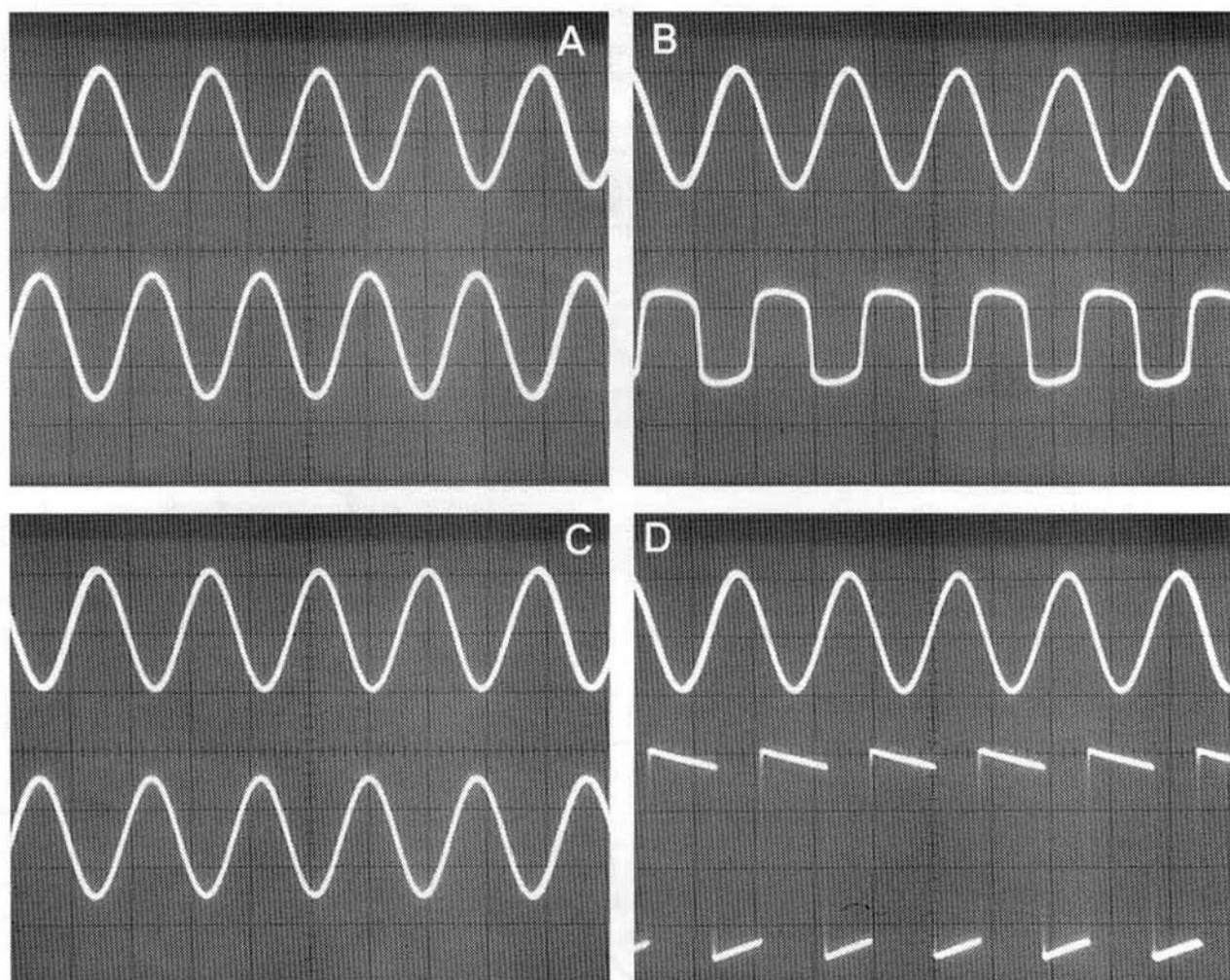
Fig. 2–3. Distort-O-Matic I layout & wiring diagram.

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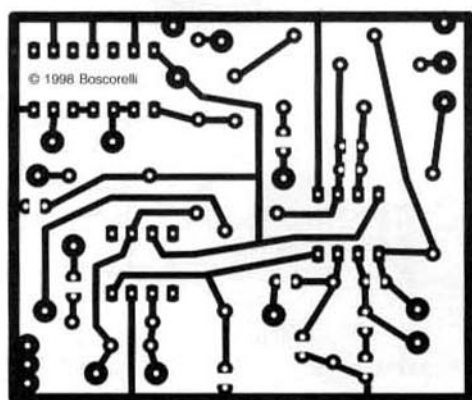
Fig. 2–4. Distort-O-Matic I circuit board.
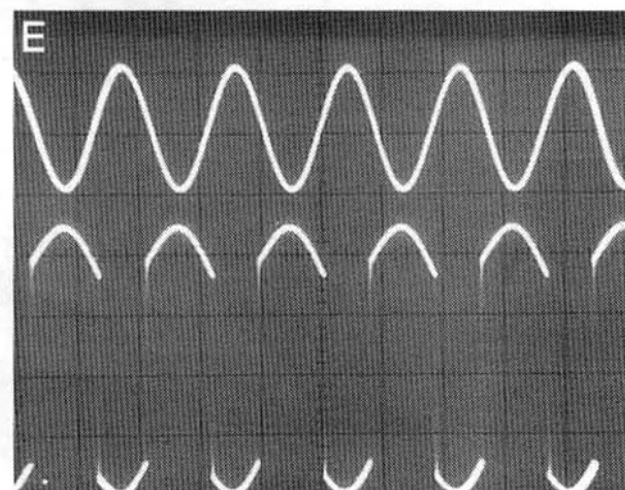
Fig. 2–5. DM1 I/O. All photos 1-KHz sinewave; top trace (input) 100 mv/div., bottom trace (output) 500 mv/div. A—Preamp gain max, clean output 50%; box acts as flat preamp. B—Preamp gain max, clean level 0, distortion threshold in middle of sweet spot, dirty level 50%, distortion switch in 'soft' position. C—Preamp gain minimum, dirty level 50%; note that low-level input generates sine-like output from dirty channel, which will gradually change to resemble photo B as input level rises or as preamp gain is raised. D—Preamp gain max, dirty level 50%, harsh output selected. E—Same as D, but clean level has been advanced to 50%, mixing clean and dirty.

Project No. 3
Parametro-Matic

By enabling continuous control over boost, frequency, and bandwidth, parametric EQ can apply a touch as fine as a scalpel or as broad as a scythe.

Circuit Function

Instrument feed couples through C8 to input of pre-amp IC1-a, whose gain is fixed at 3.2. IC1-a output couples through R16 to IC1-b, an inverting amp with nominal gain of 1. Boost/cut is achieved by divider action of R17 passing the signal through a state-variable bandpass filter comprised of IC1-d, IC2, and associated components; the signal emerges from IC1-d and couples through R11 to the input of IC1-b. The result is that, when R7 is turned fully CW, IC1-b sees an input impedance reduced by a factor of ~6.5 (1.5K, the value of R11), but only for the frequencies passed by the bandpass filter. Gain for those frequencies then becomes 10K÷1.5K = ~6.5. When R7 is turned fully CCW, the bandpass is placed in the feedback loop of IC1-b, lowering feedback resistance to 1.5K, again only for the frequencies passed by the bandpass filter. Gain for those frequencies becomes 1.5K÷10K = 0.15. With R7 centered, IC1-b exhibits no frequency emphasis and passes all signals at unity gain. R7 varies frequency over the range ~20–400 Hz; R3 varies bandwidth from <$\frac{1}{4}$ octave to >1 octave. Independent control over boost, bandwidth, and frequency defines a parametric equalizer.

IC1-b output couples through R14 to input of IC1-c, an inverting amp whose gain varies 0–5 by R12. Signal couples through R13 & C4 to the output path.

Use

Switches and pots have these functions:

S1	equalize/bypass
R3	bandwidth
R7	center frequency
R12	output level
R17	boost/cut

Initial settings: S1 equalize, R3 fully CW; R7, R12, R17 centered.

Connect unit to bass guitar and amp, establish desired volume. Turn R17 fully CCW (full cut), take R7 through full range, note effect on sound. Turn R3 fully CCW (minimum bandwidth), again take R7 though its full range and note the effect of narrowed bandwidth.

Now slowly rotate R17 CW past center and explore boost functions.

In stock configuration, Parametro-Matic provides versatile bass EQ. It can emphasize the fundamental on the open E without boosting the octave above. It can notch out a bothersome peak, or apply quasi-emphasis to the low bass notes by notching out their harmonics.

Boost/cut of the prototype measured ±13–15 dB, depending on frequency and bandwidth setting.

Modifying Parametro-Matic

The 0.1μF values of C1 & C2 give the device a range of ~20–400 Hz, suiting bass. To cover a range suitable for guitar, change C1 & C2 to 0.056μF or 0.033μF.

Finer control of center frequency (at the cost of a

PARAMETRO-MATIC PARTS LIST

Resistors
R1, 2, 4, 5, 6, 8 4.7K
R3 50K dual pot
R7 100K reverse-audio dual pot
R9, 10, 14, 15, 16 10K
R11 1.5K
R12 50K audio-taper pot
R13 100
R17 10K pot
R18 2.2K
R19 1K
R20 150K

Capacitors
C1, 2 0.1μF polypropylene
C3, 6 10pF
C4, 7, 9, 10 10μF electrolytic
C5 100pF
C8 10μF nonpolar electrolytic

Semiconductors
D1, 2 1N4001 diode
IC1, 2 TL074 quad op amp

Miscellaneous
S1 DPDT switch
¼" jacks, enclosure, wire, battery snaps, knobs, etc.
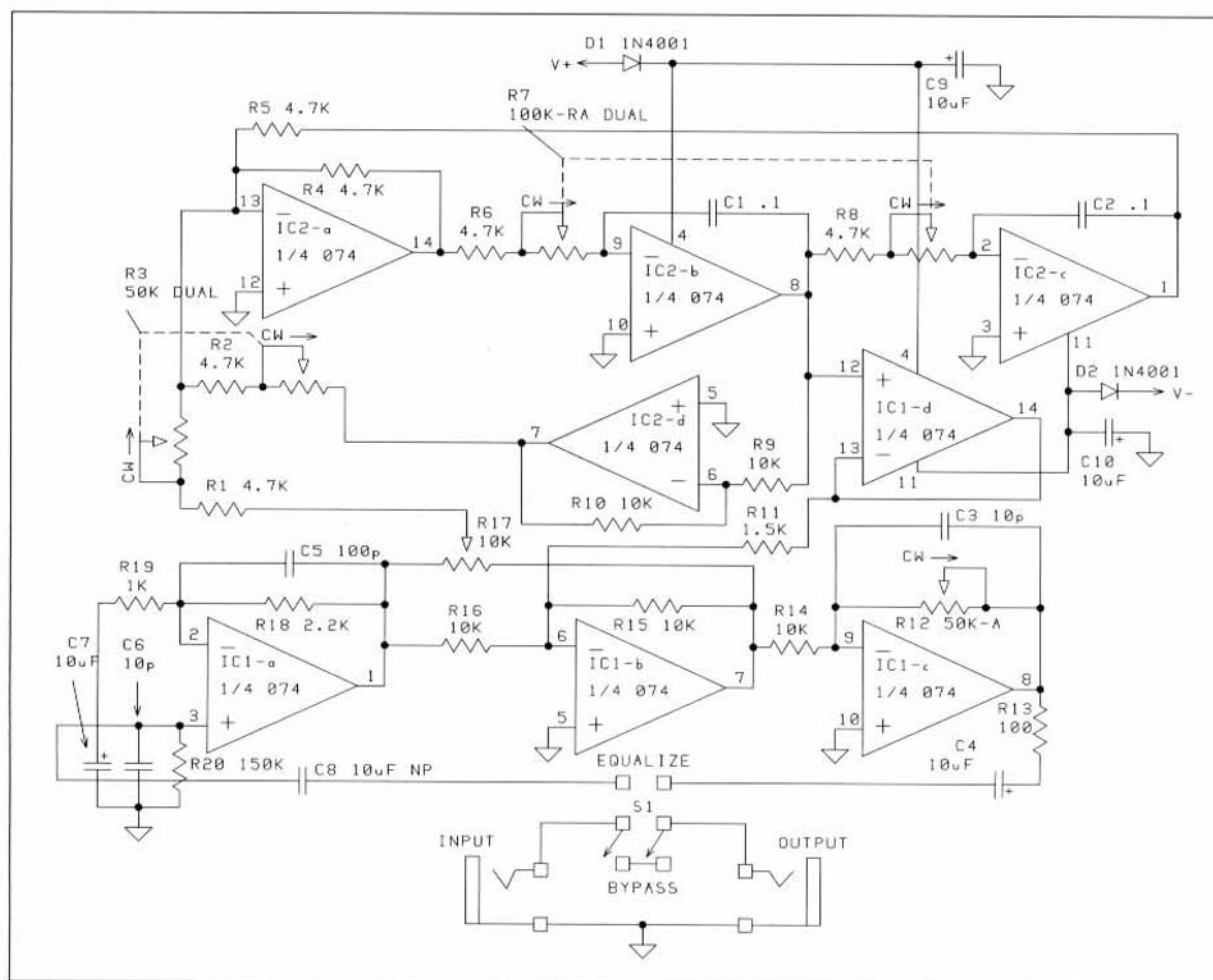
Fig. 3–1. Parametro-Matic schematic.
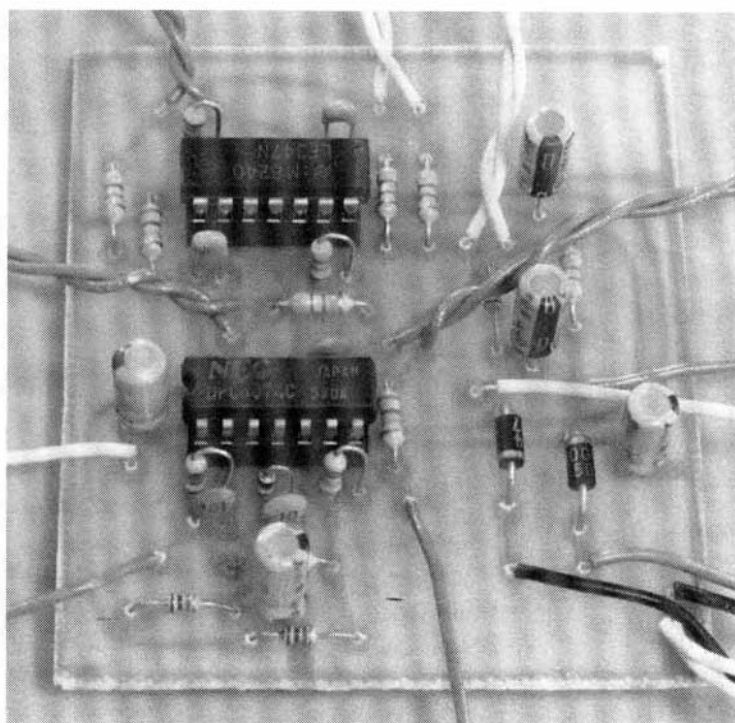
Fig. 3–2. Parametro-Matic
prototype board.

Fig. 3–3.
Parametro-Matic
layout & wiring
diagram.
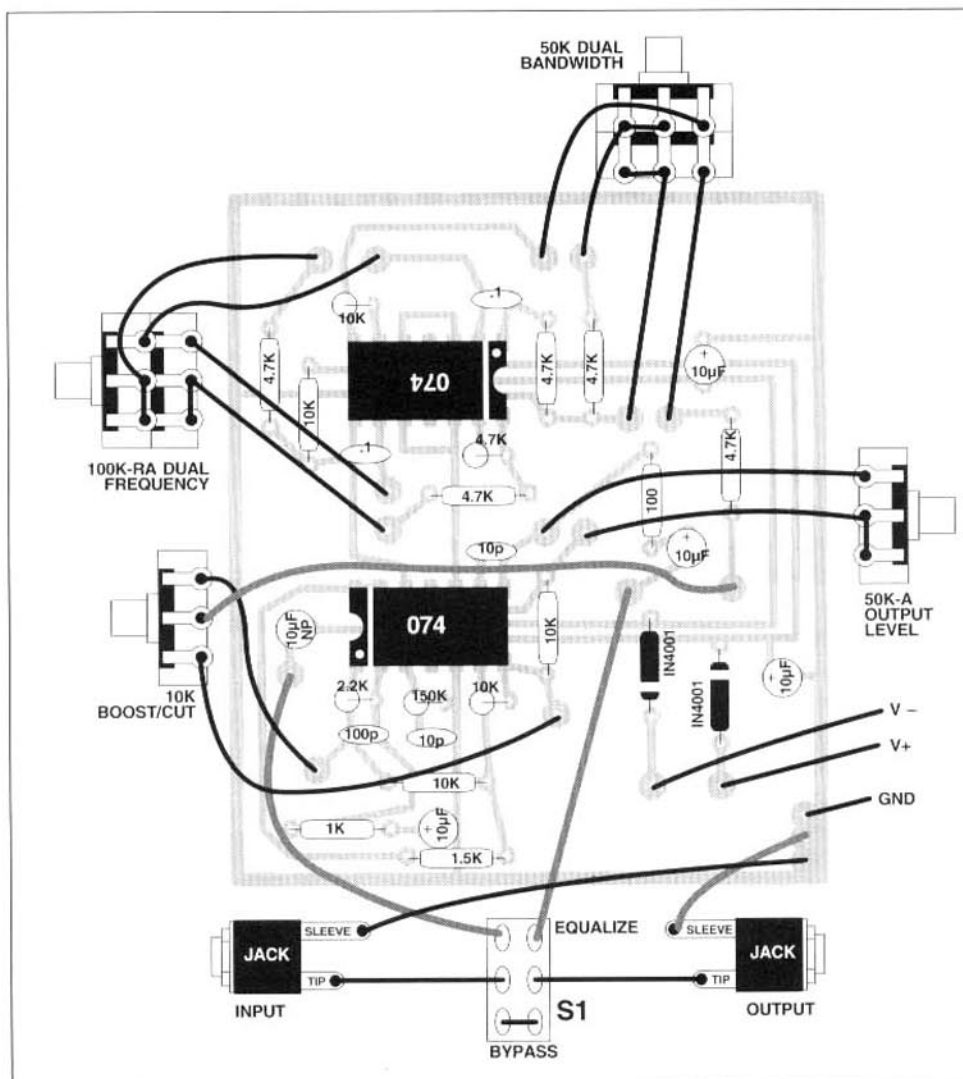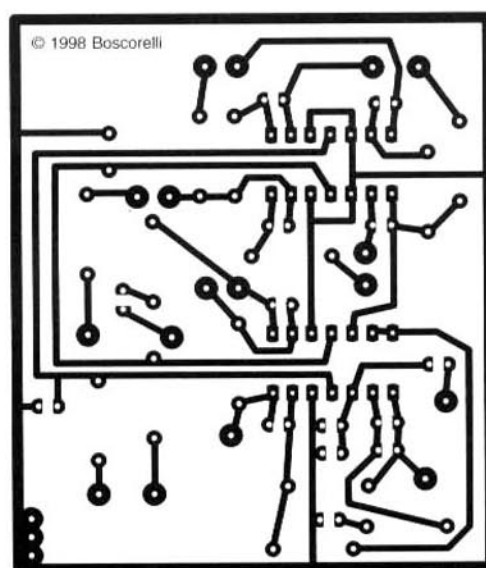
Fig. 3–4. Parametro-Matic
circuit board (below right).

3" × 3.5" reference box

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narrower tuning range) can be obtained by making R7 a 50K or 20K dual pot. When R7 is fully CW, center frequency = $1 \div (6.28 \times 4700 \times C1)$, where C1 is in farads. When R7 is fully CCW, center frequency = $1 \div [6.28 \times (4700 + \text{value of R7}) \times C1]$.

Ideally, R7 should have a reverse-audio taper, but dual RA pots are hard to find. One means to get finer control over frequency is to use a dual audio-taper pot and reverse the wiring to the pot end-terminals. In this configuration, highest frequency occurs with R7 fully CCW.

Parametro-Matic provides enough boost to demand careful level management. If raw instrument output measures peaks at 1V, it will peak at 3.2V coming off IC1-a. Those frequencies boosted by a factor of 6.5 will clip in a unit running off a pair of 9V batteries. Lower volume on the instrument if clipping occurs with boost. Parametro-Matic makes a great candidate for a ±18V supply to extend headroom beyond $30V_{p-p}$.

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Project No. 4
Distort-O-Matic II

The basic diode clipper can produce an astounding variety of sounds. Distort-O-Matic II is a clipper-based box that throws in a sound-fattening frequency doubler. Distorted and clean feeds mix in any ratio.

Circuit Function

Instrument feed couples through C8 to input of IC1-a, a preamp with gain of 3.7. IC1-a output couples through C7 to input of IC1-b which, with IC1-c and associated components, forms a precision fullwave rectifier. IC1-c output couples to one throw of S2, whose other throw ties to IC1-a output. S2's pole couples signal through C9-R15 to input of IC2-a, an op amp configured as a simple diode clipper, modified by variable resistance R14 in series with D5-6. This alters amplitude and shape of the clipped waveform, as described below. IC2-a output couples through R13 to IC2-b; R12 varies gain applied to the distorted feed from zero to one. IC1-a output also couples to pot R8 whose wiper ties through R9 to the input of summing amp IC1-d. Audio couples through R16-C11 to the output path.

Use

Switches and pots have these functions:

    S1   distort/bypass
    S2   distortion feed select, ×1 or ×2

fundamental frequency
    R8   clean level
    R12   distortion level
    R14   distortion contour

Initial settings: S1 distort, S1 ×1; R8 straight up; R12, R14 fully CCW. Connect unit to instrument and amp, establish suitable listening level.

Eliminate clean level by turning R8 fully CCW. Slowly advance R12 to sample the distorted feed. Note effect on tone and volume of distortion contour control R14. Toggle S2 between ×1 and ×2 frequency, note the fattening effect of doubling the fuzz frequency. Either R8 or R12 must be open for any output to exist. Box suits guitar and bass.

DISTORT-O-MATIC II PARTS LIST

Resistors
R1 1K
R2 150K
R3 2.7K
R4, 5, 6, 7, 10, 11, 13, 15 10K
R8, 12 10K audio-taper pot
R9 4.7K
R14 100K pot
R16 100

Capacitors
C1, 3, 11, 14 10µF electrolytic
C2, 12, 13 10pF
C4, 5, 6, 10 100pF
C7 0.1µF
C8 10µF nonpolar electrolytic
C9 1µF electrolytic

Semiconductors
D1, 2, 5, 6 1N914
D3, 4 1N4001
IC1 TL074 quad op amp
IC2 TL072 dual op amp

Miscellaneous
S1 DPDT switch
S2 SPDT switch
¼" jacks, wire, circuit board, knobs, 9V battery snaps, etc.

3.25" × 2.5" reference box

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Fig. 4–1. Distort-O-Matic II circuit board.
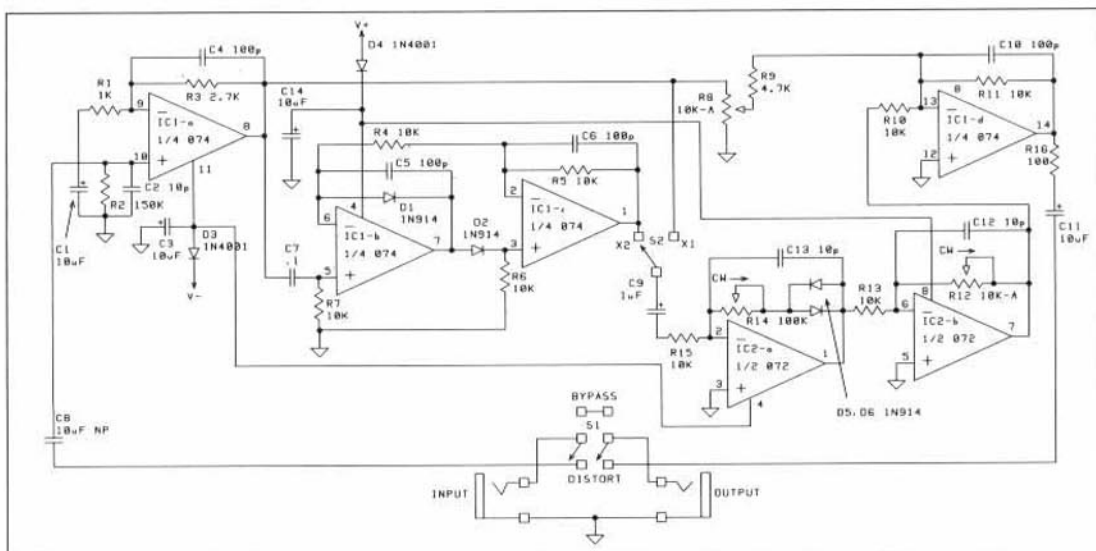
Fig. 4–2. Distort-O-Matic II schematic.

Fig. 4–3. Distort-O-Matic II layout & wiring diagram.
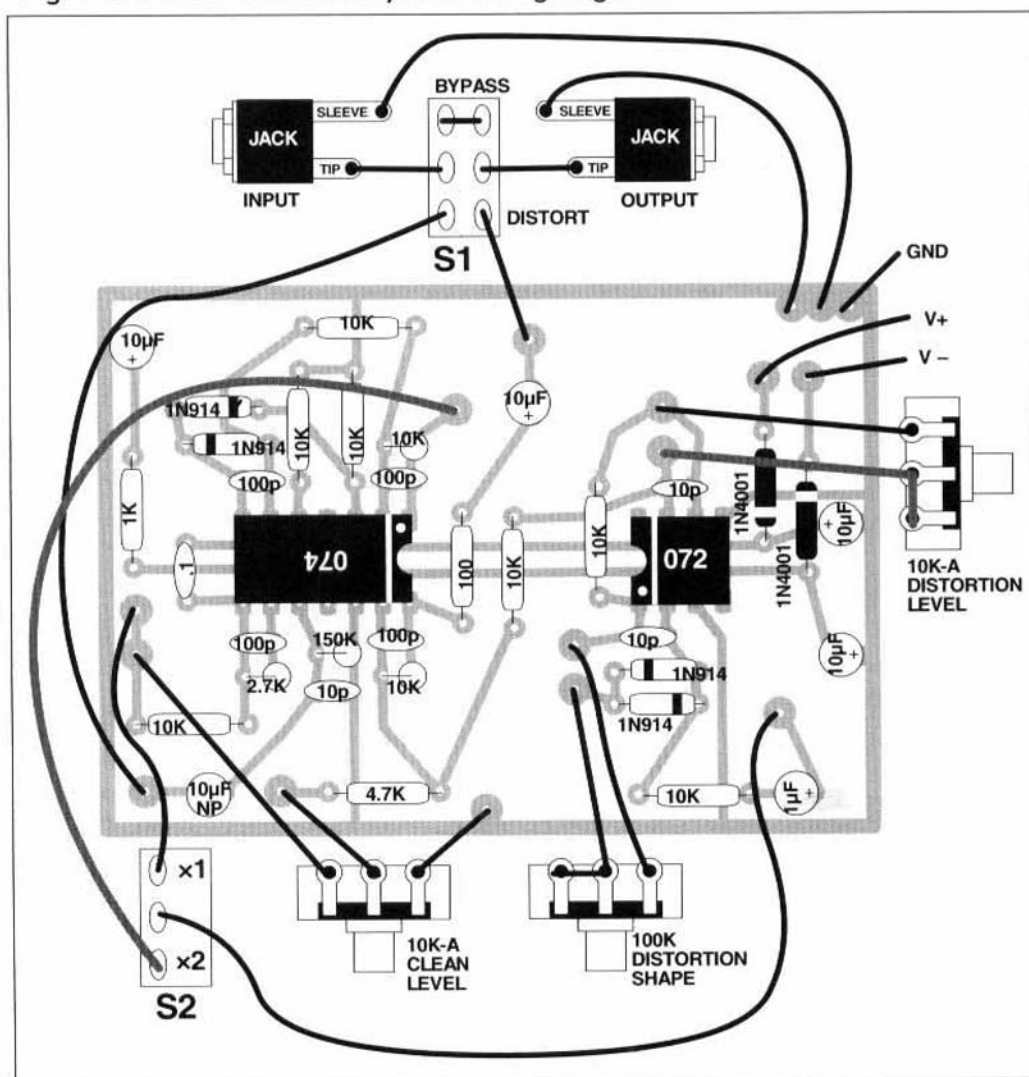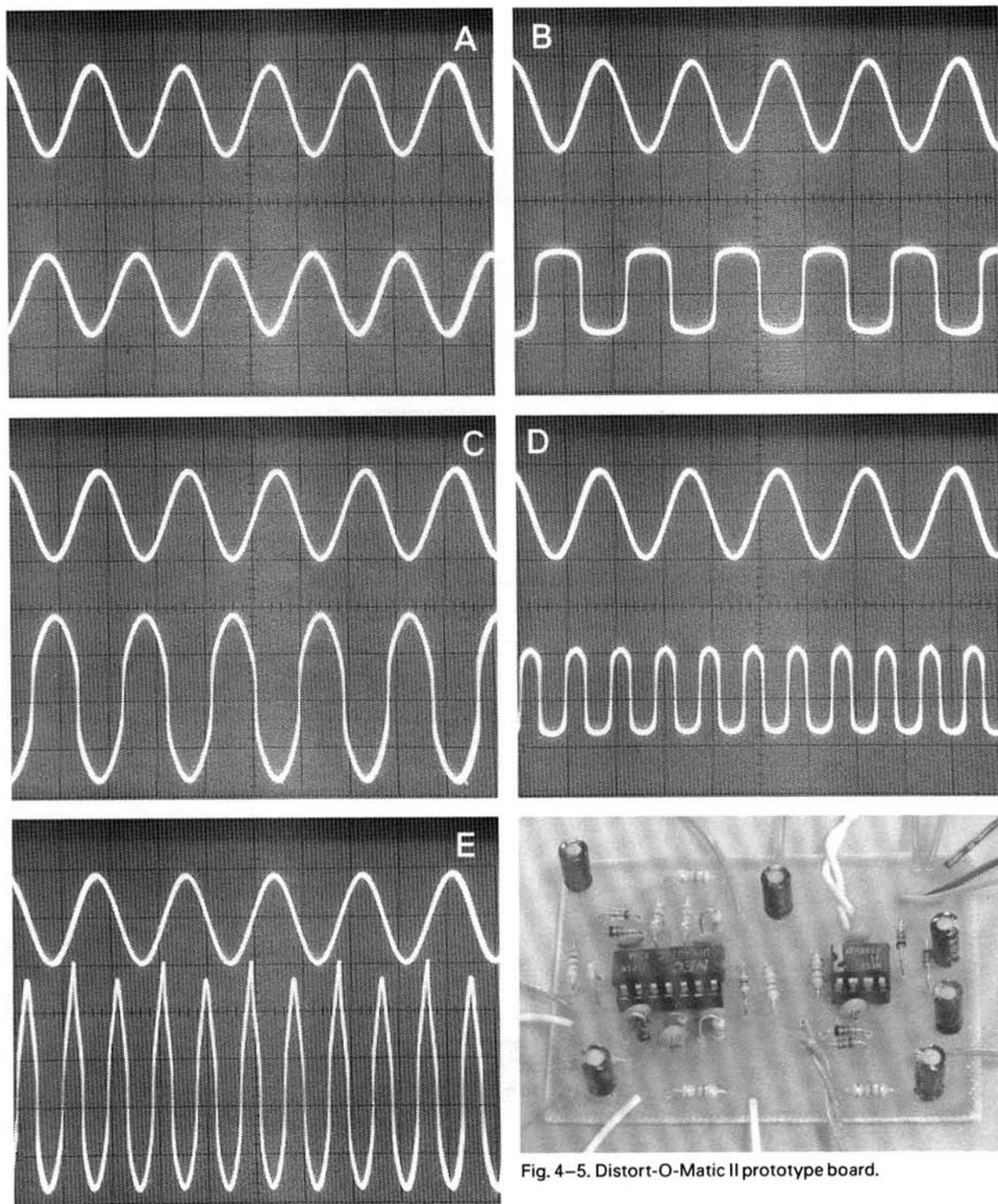
Fig. 4—4. DM2 I/O. All photos 1 KHz sinewave; top scale (input) 100 mv/div., bottom scale (output) 500 mv/div. A—Clean level 60%, distorted level 0. B—Clean level 0, distorted level 40%, R8 fully CCW. C—Same as B, but shape control R14 straight up. D—Same as B, but ×2 selected. E—Same as D, but shape control R14 straight up.

Fig. 4—5. Distort-O-Matic II prototype board.

Project No. 5
Tremolo-Matic

Tremolo contains greater nuance than rate and depth imply. Tremolo-Matic offers three distinct modes in one box.

Circuit Function

Signal Path: Instrument feed couples through C7-R3 to inverting preamp IC1, whose gain varies 0–5 by trimpot R4. IC1 output couples through C5-R7 to signal input of IC4, half an NE570 configured as a VCA whose gain varies 0–1, according to the control voltage supplied through R11. VCA output pin 10 couples through pot R9 to output path R10-C4. The signal path is noninverting.

Control Path: IC2-a/-b and associated components form a sinewave oscillator whose frequency varies ~1–10 Hz under control of R21. IC2-b output couples to depth control pot R16, through R15 to summing amp IC2-c. Pot R14 varies the DC offset present at IC2-c's output. Sinewave control voltage couples through R11 to VCA control port, pin 16 of IC4. IC3 and R5-R6 form a feedthrough trim network.

Use

Switches and pots have these functions:

R4	preamp gain
R5	VCA feedthrough trim
R9	output level
R14	static gain
R16	tremolo depth
R18	sine trim
R21	tremolo rate
S1	tremolo/bypass

First, trim the sinewave generator. Set R21 fully CW, connect oscilloscope probe to sinewave output pin 7 of IC2. Trim R18 to give $3V_{p-p}$.

Next, trim VCA feedthrough. Set R14 straight up, R16 fully CW. In this state the control voltage coming off IC2 pin 8 clips at both extremes. Apply scope probe to IC4 pin 10, trim R5 for least feedthrough.

If no scope is available, trim feedthrough by ear. Configure settings as above, short the signal input, turn output level pot R9 fully CW. Connect unit to amp whose volume is turned all the way down. Slowly increase amp volume until beating is heard. Exercise care, because the pulses coming off IC4 pin 10 could measure up to several $volts_{p-p}$. Trim R5 for least feedthrough.

Initial settings: S1 tremolo, R21 fully CW; R4 centered; R9, R14, R16 straight up. Connect unit to axe and amp, establish desired listening level. Clear tremolo should be heard with these settings. Trim R4 for desired preamp gain. Take R21 through its range and note change in rate.

Specific settings of depth and static gain let Tremolo-Matic provide three distinct types of tremolo:

▶ With static gain centered and depth at ~40%, the control feed causes instrument volume to intensify on positive control peaks; to soften on negative ones.

▶ With static gain at maximum and depth at ~50%, instrument volume softens only during negative control peaks. Positive control peaks clip in IC2-b and can-

TREMOLO-MATIC PARTS LIST

Resistors
R1, 2, 8, 15 36K
R3 22K
R4 250K trimpot
R5 50K multiturn trimpot
R6 220K
R7 4.7K
R9 10K audio-taper pot
R10 100
R11, 19 47K
R12, 13 100K
R14, 16 10K pot
R17, 22, 23, 24 2.2K
R18 200 ohm multiturn trimpot
R20 150K
R21 2M reverse-audio–taper pot

Capacitors
C1, 2, 5 1µF
C3 10pF
C4, 7, 8, 9 10µF aluminum electrolytic
C6 47pF

Semiconductors
D1 1N4001
D2, 3 1N914
IC1 MC33171 low-power op amp
IC2 TL064 quad low-power op amp
IC3 78L05 5V positive regulator (TO-92)
IC4 NE570/571 dual-channel VCA

Miscellaneous
S1 DPDT switch
¼" jacks, wire, solder, 9V battery, circuit board, etc.
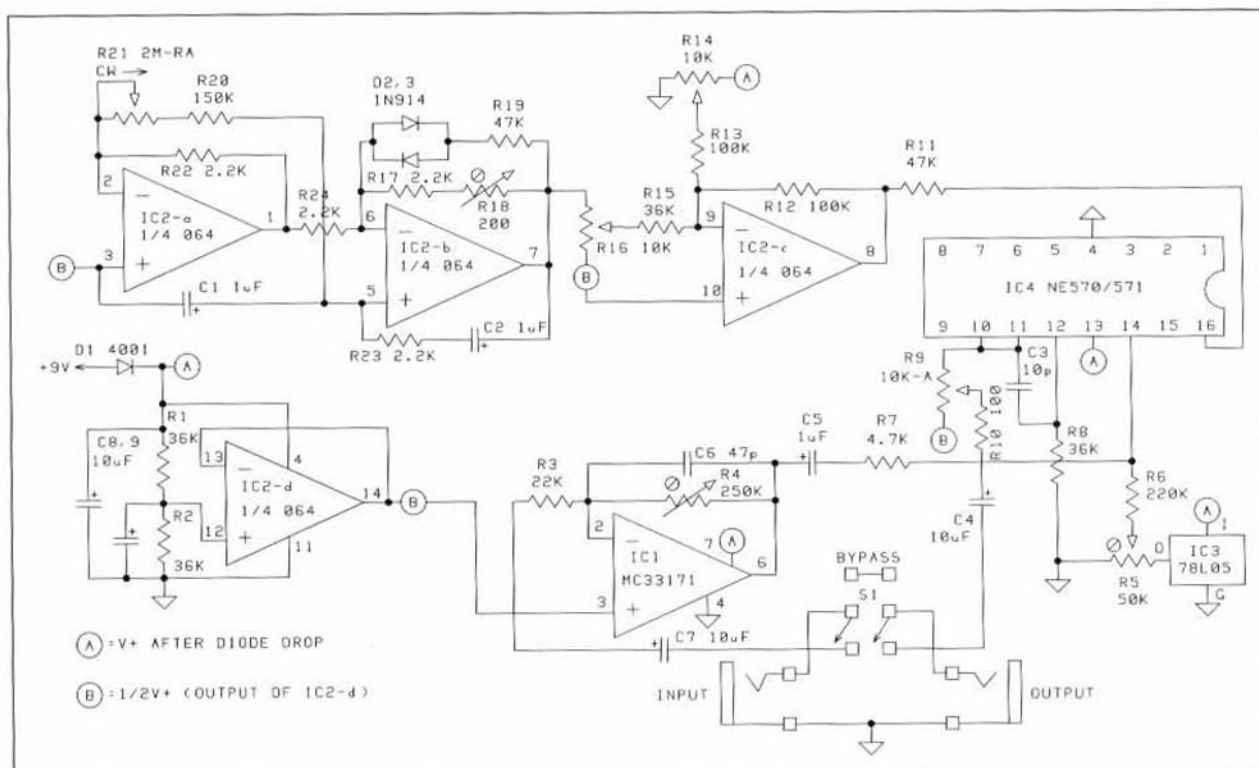
Fig. 5-1. Tremolo-Matic schematic.
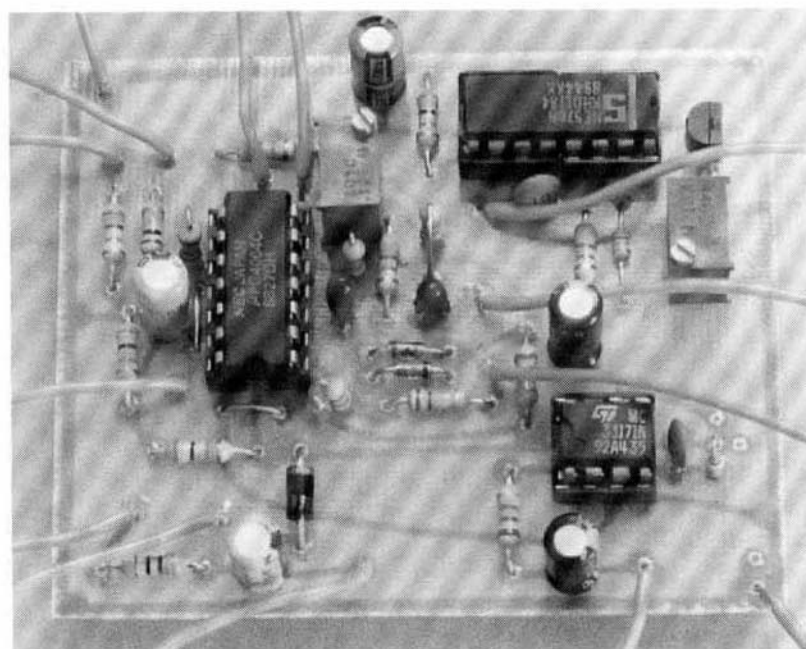
Fig. 5-2. Tremolo-Matic prototype board.

not further affect volume.

▶ With static gain at minimum and depth at ~60%, sound pulses from a background of silence, lending the effect a percussive air.

Notes

TM is a true 9V box, optimized for 7.5V on the positive power bus. Running off higher voltage requires increasing the value of R7 to accommodate greater signal amplitude, and reducing the value of R8 to keep IC4's output offset near $\frac{1}{2}$V+.

Preamp gain should be adjusted for an average output of 1V_{p-p}. This leaves plenty of room for peaks. The system has ~5V_{p-p} of headroom.

The use of an inverting preamp breaks one of the rules for low noise, but in this case keeps a noninverting signal path without having to add an inverting output buffer.

3.25" × 2.75" reference box

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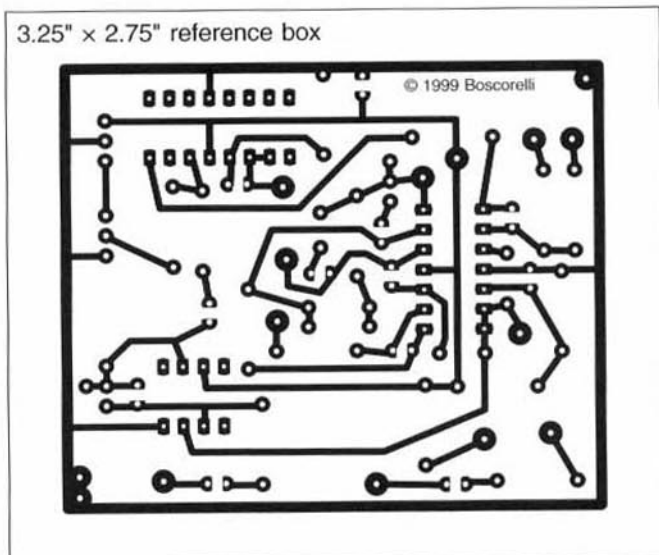
Fig. 5–3. Tremolo-Matic circuit board.
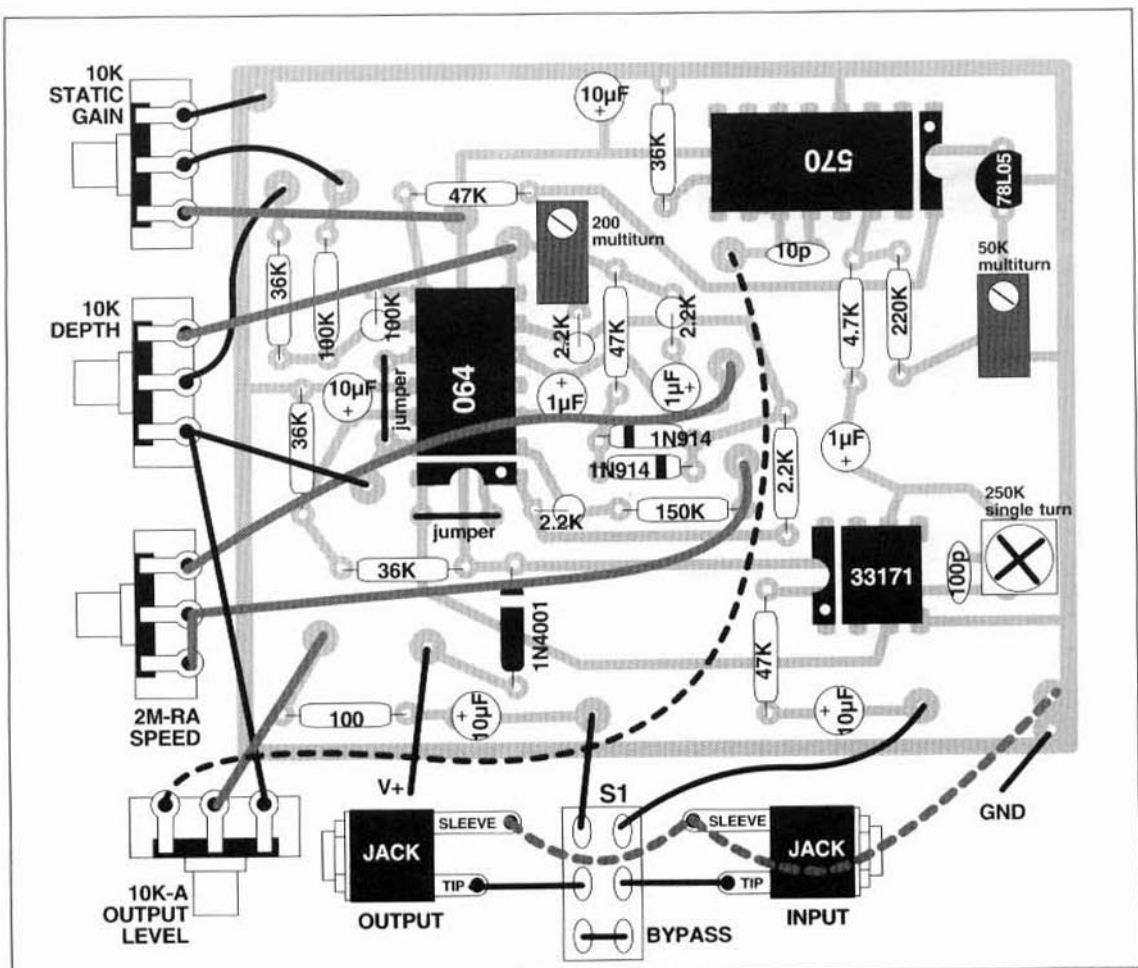
Fig. 5–4. Tremolo-Matic layout & wiring diagram.

21
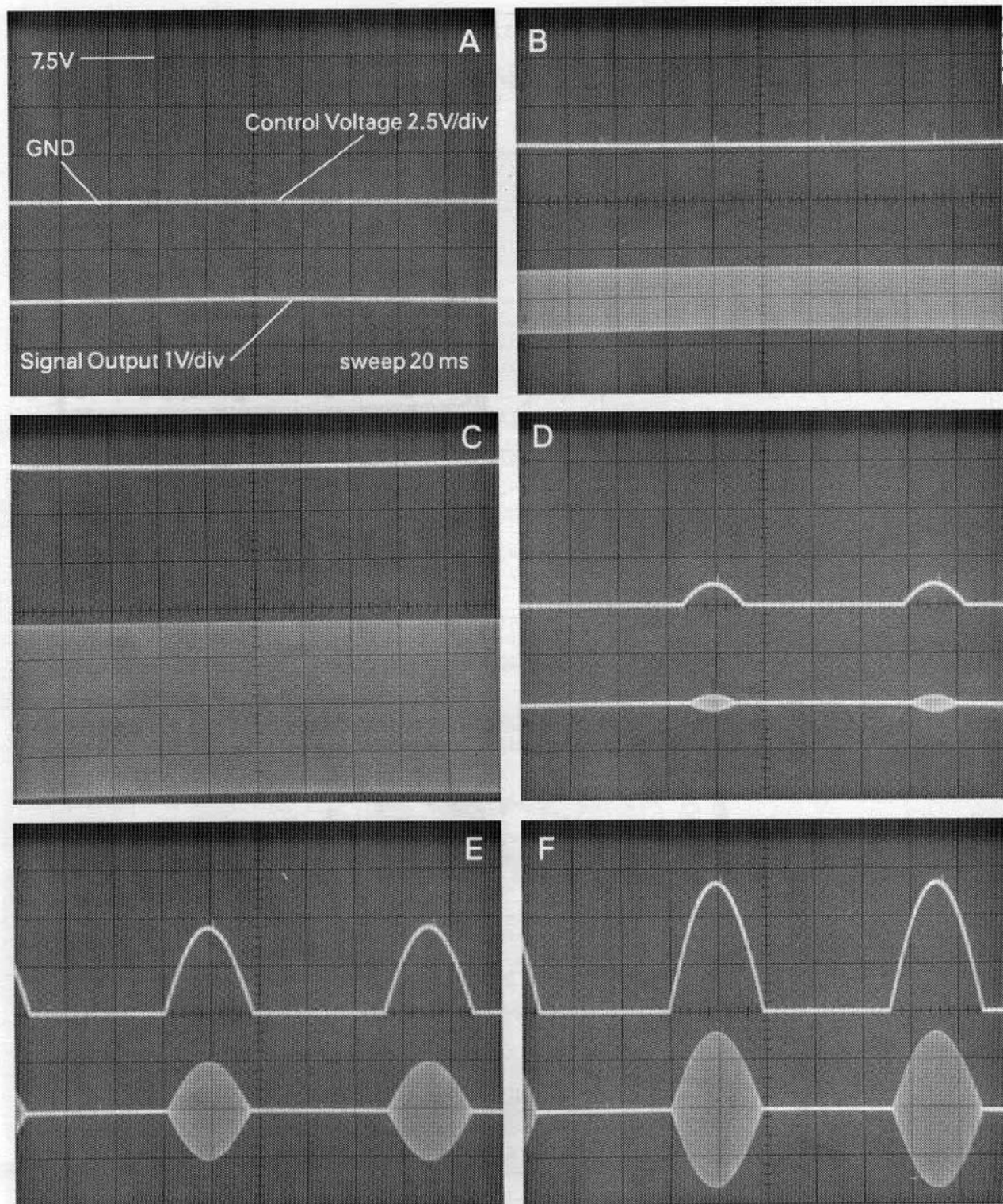
Fig. 5–5. Tremolo-Matic output displayed with VCA control voltage. Top trace shows control voltage, 2.5V/div.; bottom trace shows TM output, 1V/div. Input is a 5-KHz sinewave. A—Control voltage and static voltage at their lower limits; VCA output is 0. B—Static voltage (determined by setting of R14) rises to about 40%; signal output rises. C—Static voltage is at maximum, VCA gain is ~1. D—Static VCA gain back to 0; sinewave impressed on static resting gain modulates instrument volume. E, F—Amplitude of sine feed continues to rise, with resultant effect on signal volume. Pulse artifacts (most visible in photo B, top trace) result from an external sync generator, not the tremolo circuit.
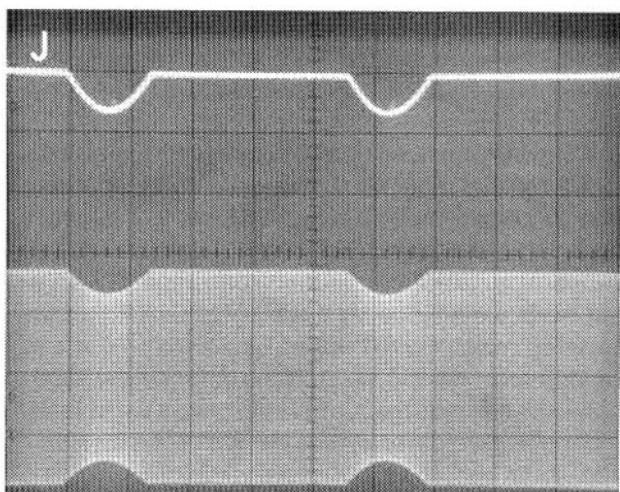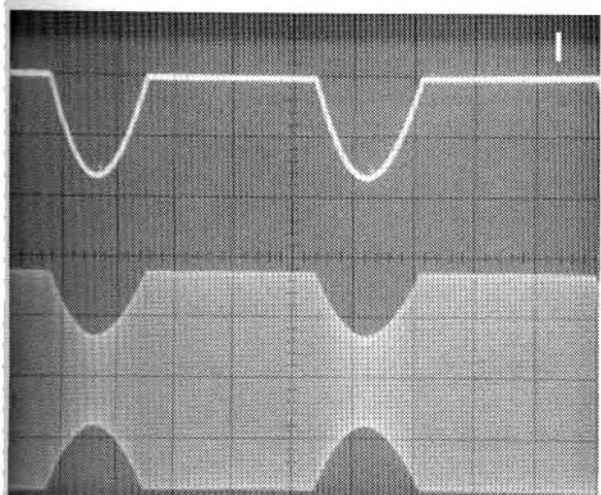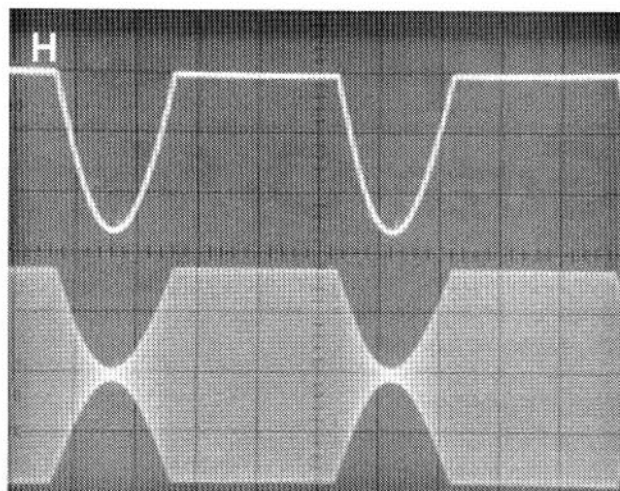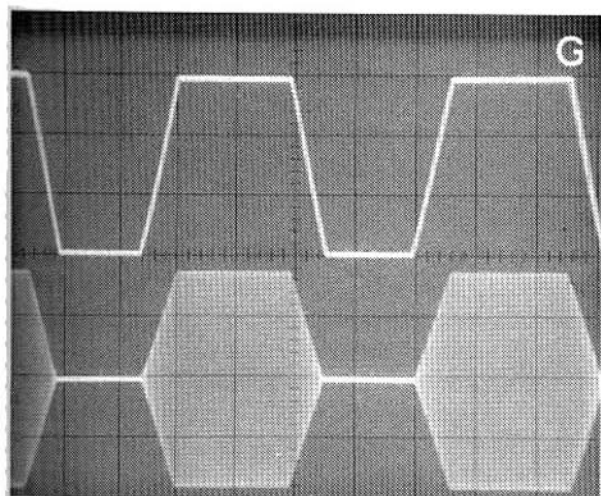
Fig. 5–6. G—Sinewave voltage has reached maximum, clipping at both extremes. VCA gain swings from zero to one with each cycle. H—Sinewave amplitude still at maximum, but static gain has been shifted upward. Now only the negative part of the sinewave modulates the signal. I & J—Sinewave amplitude is progressively reduced. K—A more typical tremolo control setting: static VCA gain is ~45%, sinewave amplitude ~60%. Figs. F, I, and K represent three audibly distinct manifestations of tremolo, attained by specific combinations of static gain and sinewave amplitude.
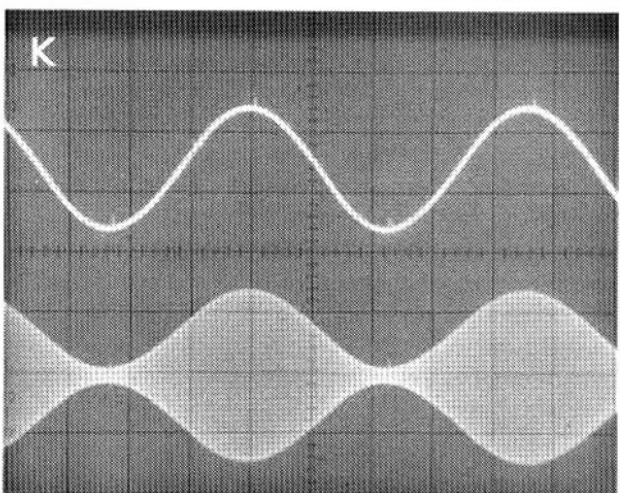
Project No. 6
Gate-O-Matic

Like sophisticated rackmount noise gates, Gate-O-Matic (GOM) offers independent control of attack, decay, threshold, and ratio.

Circuit Function
Signal Path: Line-level input couples through C1-R1 to unity-gain inverting buffer IC1-b, whose output couples through R4-C4 to IC3 signal input. R5-C5 acts as a snubber. IC3 signal output (pin 4) ties to input of current-to-voltage converter IC1-c. R22 varies the output level; audio couples through R23-C9 to the output path.

Control Path: IC1-b output couples through R3-C3 to IC3 rectifier input pin 9. Raw level detector output (pin 2) couples to noninverting amp IC2-a, which boosts the voltage by a factor of 40. A variable DC bias supplied by R20 varies the gating threshold. IC2-a output feeds variable-decay network D3/Q1/C6 and associated resistors. R11 varies decay from ~5 ms to several seconds. IC2-b buffers the output of the decay network and feeds the control voltage to a variable-attack network made up of D2/C7/R16. R16 varies attack from instantaneous to ~40 ms.

Output of the variable-attack network feeds buffer IC1-a, biased through R15 to allow its output to swing below ground. Control voltage couples through D1 to R13, which varies the percentage let through to buffer IC1-d, whose output feeds final divider R12-R7, which ties to IC3 pin 5, the VCA '+' control port.

The result of this control path is for a strongly negative voltage to exist at the output of IC1-a in the absence of an input signal. An input signal generates a positive voltage that keeps the potential at the anode of D1 at ground, preventing any control voltage from reaching the VCA. Only when signal falls below the level needed to overcome D1's forward drop does a negative voltage get to IC3 pin 5.

Use
Switches & pots have these functions:

S1	gate/bypass
R11	decay
R13	downward expansion ratio
R16	attack
R20	expansion threshold
R22	output level

Initial settings: S1 gate; R11, R16 fully CCW; R13, R20 fully CW; R22 straight up. In this condition the unit acts as a buffer with gain of about 1. Connect unit to line-level feed and line-level target device, such as an amp. Establish desired volume level.

Using program material suitable to detect gating, turn threshold control R20 CCW until obvious gating action is noted. Vary attack and release to suit taste. Note the effect of altering the downward expansion ratio.

Attack controls how quickly GOM opens after closing. R16 covers the range <1 ms to ~40 ms, longer transitions sounding less abrupt. Decay controls how

GATE-O-MATIC PARTS LIST
Resistors
R1, 2 22K
R3, 10 10K
R4 36K
R5 47
R6, 7 200
R8 1.5M
R9 10M
R11 250K pot
R12, 18 1K
R13 100K pot
R14, 15, 17, 19 39K
R16 250K audio-taper pot
R20 20K pot
R21 2.2K
R22 100K audio-taper pot
R23 100
R24 150K

Capacitors
C1 10μF nonpolar electrolytic
C2, 8 10pF
C3, 4, 9, 10, 11 10μF aluminum electrolytic
C5 0.0022μF
C6 0.01μF
C7 0.1μF

Semiconductors
D1, 3 1N914
D2 1N34A
D4, 5 1N4001
IC1 TL074 quad op amp
IC2 TL072 dual op amp
IC3 SSM2120 dual-channel dynamic
   controller

Miscellaneous
S1 DPDT switch
¼" jacks, circuit board, wire, knobs, etc.

Fig. 6–1. Gate-O-Matic schematic.

long GOM takes to close once input level drops below threshold. R11 varies decay from ~5 ms to several seconds. Ratio controls how much gain reduction the unit applies. Maximum ratio mutes the signal, minimum ratio is 1:1, which results in no change in gain.

The prototype was tested using a ±15V supply. Lowering the supply to the ±7.5V typical of '9V' batteries will require shorting R21, and possibly lowering the value of R19, to keep a useful threshold range.

Fig. 6–2. Gate-O-Matic layout & wiring diagram.

3" × 3" reference box

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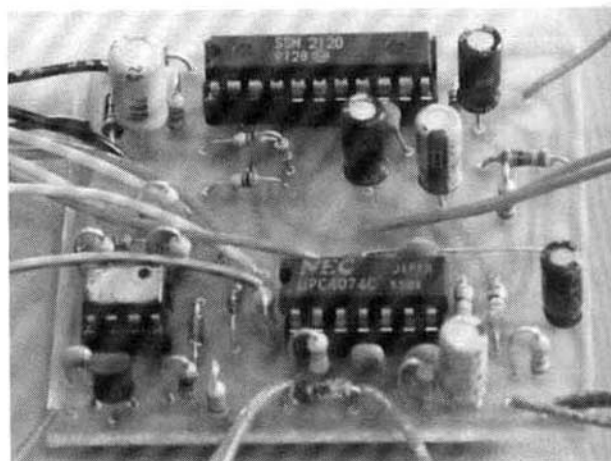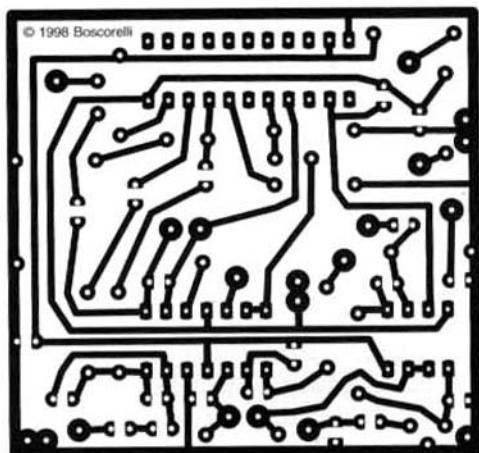
Fig. 6–3 Gate-O-Matic prototype board.

Fig. 6–4. Gate-O-Matic circuit board.

A   B

C   D
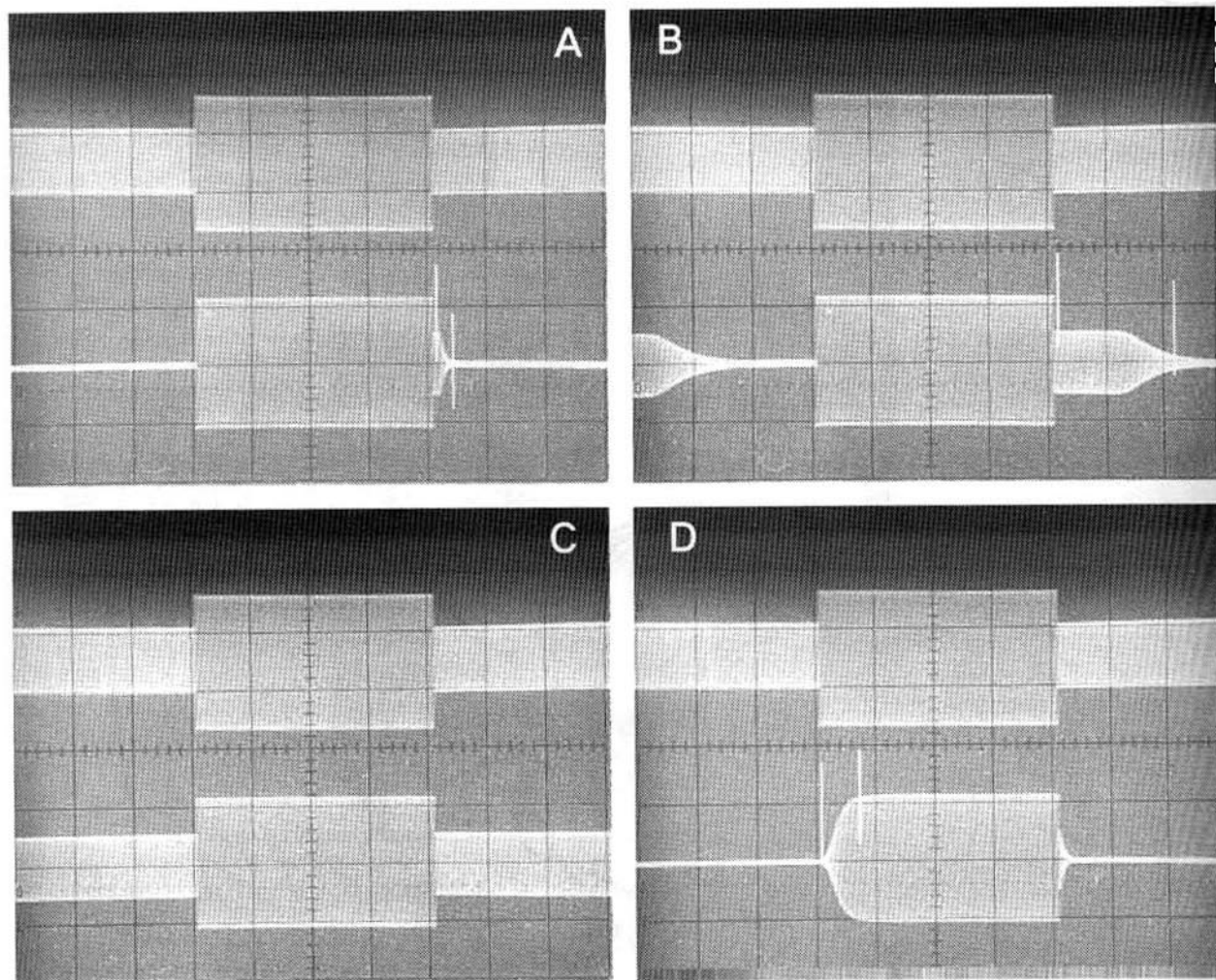
Fig. 6–5. Gate-O-Matic's effect on tone burst. Top trace input, bottom trace GOM output; sweep 20 ms/div., scale 1V/div., expansion ratio maximum. A—Gating threshold is below large pulse but above small pulse. Attack is at minimum, resulting in instant opening of gate. Decay is at its 5-ms minimum. B—Decay has been lengthened to about 40 ms. Note delay before gate closes. C—Decay has been extended to maximum. In this case, gate stays open between pulses. D—Decay back to minimum, attack extended to ~10 ms. Note delay for gate to open after being closed.

Project No. 7
Squeeze-O-Matic

The essential features (and specs) of a rackmount compressor, plus an extra seldom found in commercial products, all on a board that fits a stomp box with room to spare.

Circuit Function

Signal Path: Line-level feed couples through C1-R1 to unity-gain inverting buffer IC1-b. IC1-b output couples through R4-C3 to signal input of IC3. R23 & C5 form a snubber.

IC3 output (as a current) couples to input of IC1-c, an op amp configured as a current-to-voltage converter. Net gain of this path, with no voltage applied to control pin 7, is 1. IC1-c output couples through R18 to input of inverting buffer IC1-d, which varies output level according to setting of R19. Audio couples to the output path through R20-C12.

Control Path: The level detector input of IC3 (pin 9) connects through C4 to S2, which selects a control feed from the output of the input buffer through R3, or the output of the I–V converter through R21; these paths are parallel and feedback, respectively. Raw level detector output is taken off pin 2, biased through R26, feeding directly to noninverting amp IC2-a, whose gain is fixed at 40. R14 and surrounding resistors apply a variable DC offset to the output of IC2-a, which varies compression threshold.

IC2-a output feeds a positive peak detector made up of D3, Q1, R7-9, C6, and buffer IC2-b. R8 varies decay.

IC2-b output feeds a variable-attack network made up of D2, R6, and C7, buffered by IC1-a.

IC1-a output couples through D1 to R5, which varies the magnitude of the control voltage applied to buffer IC4, and thus varies compression ratio. IC4 output couples to the final control voltage divider, R22-24.

Use

SOM is a downward compressor with variable threshold, ratio, attack, and decay. Unlike most commercial units, this box lets the user select a feedback or a feedforward control path.

Switches and pots have these functions:

S1	compress/bypass
S2	control path select feedback/parallel
R5	compression ratio, 1:1–25:1 (feedback)
R6	attack

SQUEEZE-O-MATIC PARTS LIST

Resistors
R1, 2 22K
R3, 4, 16, 18 36K
R5 100K pot
R6, 19 100K audio-taper pot
R7, 21 10K
R8 250K pot
R9 10M
R10, 12 39K
R11 1K
R13, 15 6.8K
R14 10K pot
R17 150K
R20 100
R22 1.5K
R23 47
R24, 25 200
R26 1.5M

Capacitors
C1, 12 10μF nonpolar electrolytic
C2, 8, 9 10pF
C3, 4, 10, 11 10μF aluminum electrolytic
C5, 6 0.0022μF
C7 0.1μF

Semiconductors
D1, 3 1N914
D2 1N34A
D4, 5 1N4001
IC1 TL074 quad op amp
IC2 TL072 dual op amp
IC3 SSM2120 dual-channel dynamic controller
IC4 TL071 op amp
Q1 2N3904 NPN transistor

Miscellaneous
S1 DPDT switch
S2 SPDT switch
1/4" jacks, circuit board, enclosure, mounting hardware, batteries, wire, solder, etc.
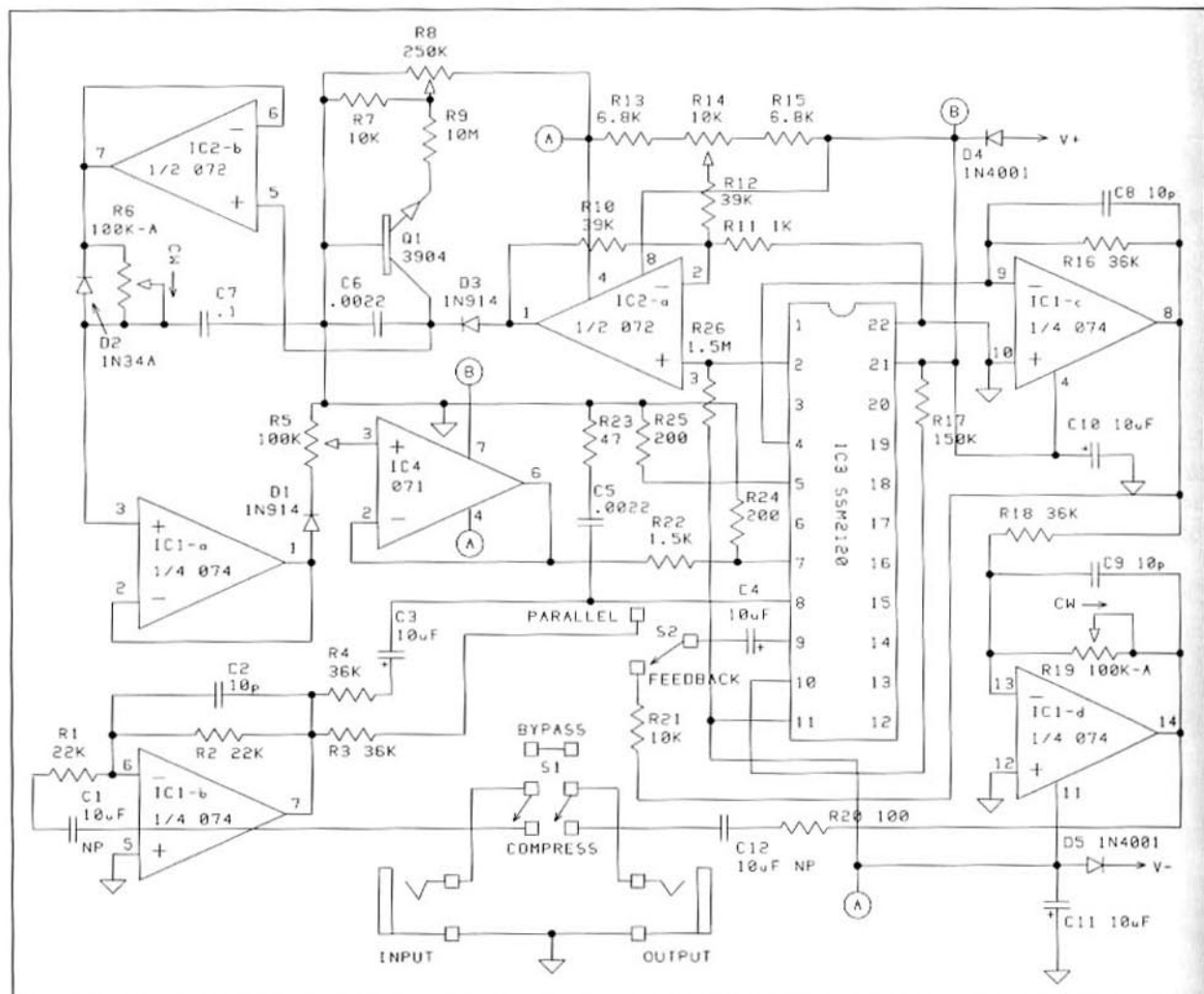
Fig. 7–1. Squeeze-O-Matic schematic.

R8   decay
R14   compression threshold
R19   output level

Initial settings: S1 compress, S2 feedback; R5, R14 fully CW, R6 fully CCW, R8 9 o'clock, R19 2 o'clock. In this state the box acts as a line-level buffer with gain of about 1. Connect unit to line-level feed and target output device; establish desired audio level.

Turn R14 CCW until obvious compression is noted. Vary the input level; vary ratio and note the effect on sound. Using appropriate program material, take attack and decay through their ranges and note the effects. Note audible distortion of low-frequency feeds when decay is at minimum.

Once familiar with feedback operation, return to initial settings and toggle S2 to parallel control path; repeat the checkout sequence. Parallel compression tends to be more obvious. In fact, with the ratio pot set past halfway, the device exhibits paradoxical dynamics: the louder the input, the softer the output. The parallel control path requires a much lower ratio for natural-sounding compression. At the proper setting, par-

allel compression can emulate 'sag' of some tube amps.

Notes

The prototype board was hardwired to a discrete version of the Append-A-Board power supply (Project No. 27), and for that reason does not include the rectifier diodes. If you elect to run SOM off batteries, install the polarity protection diodes shown on the wiring diagram.

Function described above was ascertained with the unit running off ±15V. At ±7.5V, the major change noted is a drop in maximum threshold to about $2V_{p-p}$, but only with S2 in feedback position. The threshold range can be restored by reducing the values of R13 & R15.

The user can replace R2 with a suitable pot if widely variant input levels are expected, but this introduces something of a see-saw effect, because altering the input level effectively alters the threshold.

Unless extremely rapid decay is needed, keep the decay control at 10 o'clock or higher, to avoid distortion of low frequencies.
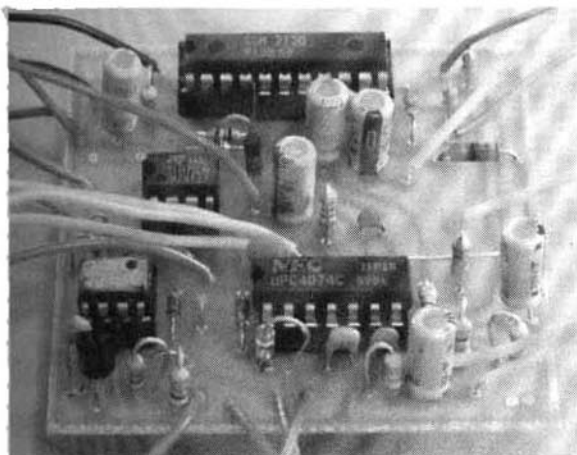
Fig. 7–2. Squeeze-O-Matic prototype board.

3" × 3" reference box

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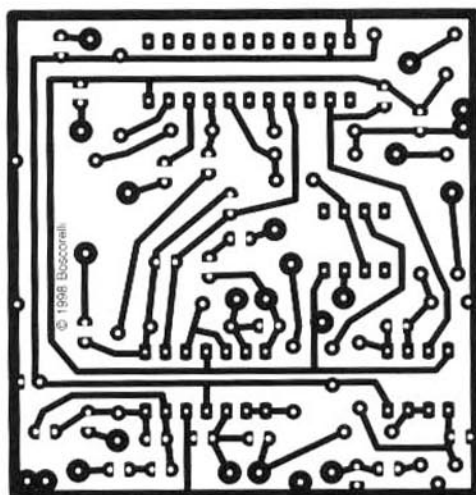
Fig. 7–3. Squeeze-O-Matic circuit board.
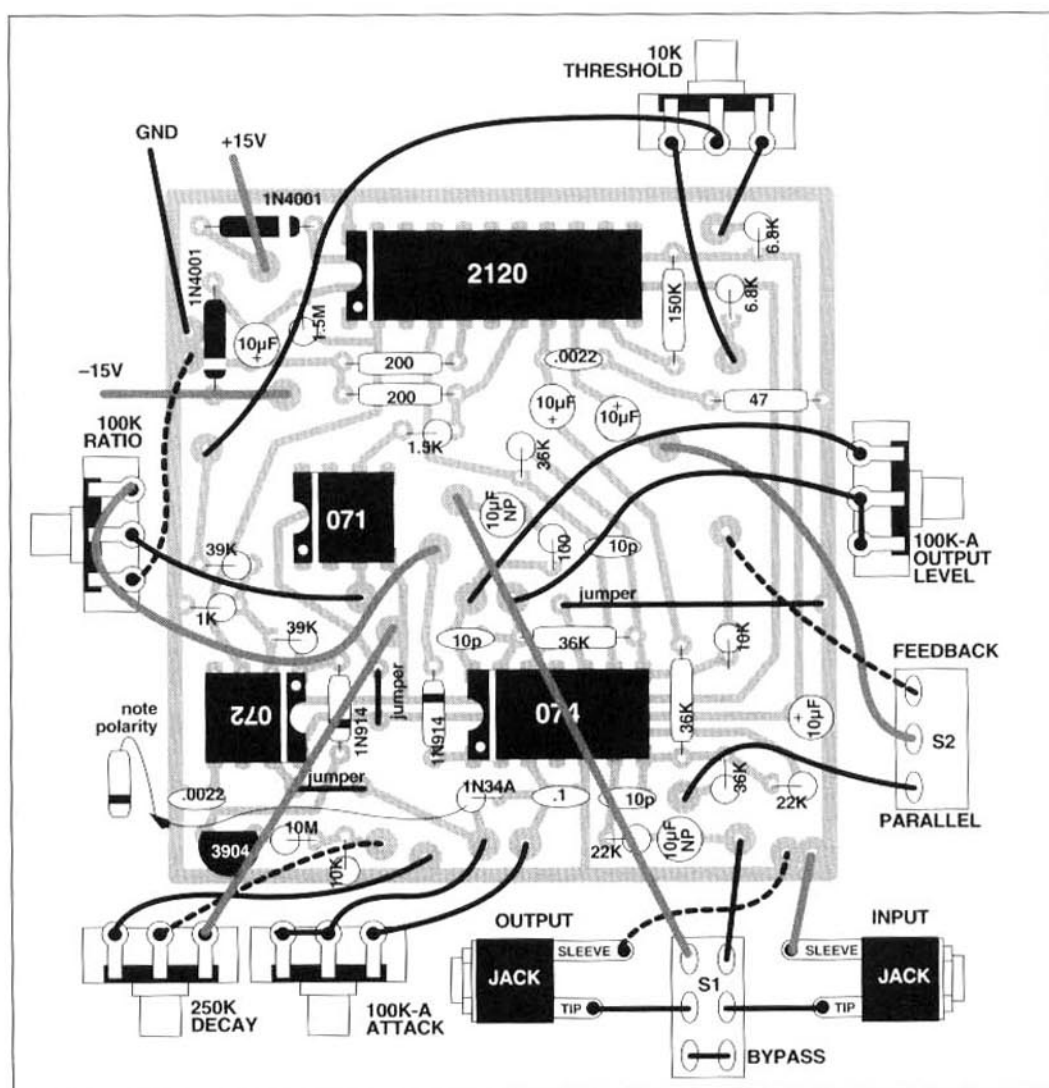
Fig. 7–4. Squeeze-O-Matic layout & wiring diagram.

29
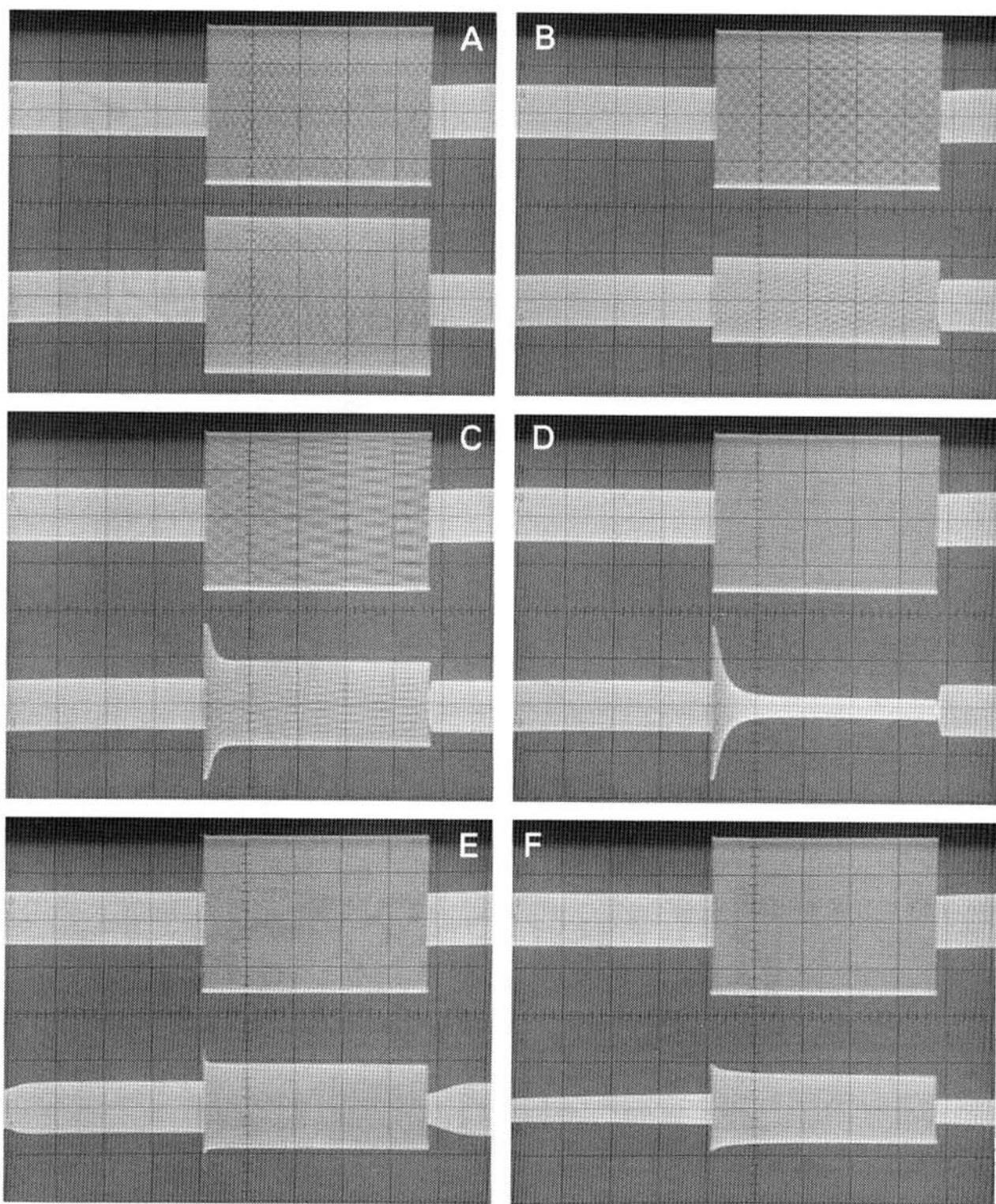
Fig. 7–5. Input/output of bi-level tone burst for selected control settings of SOM. All figures: top trace input, bottom trace SOM output; scale 2V/div., speed 20 ms/div., compression ratio maximum. A—Bi-level tone burst input, SOM output trimmed to match input. Threshold is at maximum, feedback mode; box acts as unity-gain buffer. B—Threshold has been lowered to ~3.8V$_{p-p}$; attack & decay at minimum; no detectable attack lag; decay takes about 3 ms. C—Attack has been lengthened to about 8 ms. D—Mode switched to parallel; note paradoxical dynamics: the part of the burst above threshold is attenuated so that its amplitude actually falls below that of the low-level burst. E—Back to feedback mode; decay has been increased to ~20 ms. F—Decay has been increased to ~100 ms; does not have time to decay fully between bursts.

Project No. 8

Distort-O-Matic III

Serial fullwave rectification generates progressively distorted and progressively higher-frequency products. The process makes a distortion device that sounds distinct from squarewave devices, and that retains dynamic tracking.

Circuit Function

Instrument feed couples through C14 to input of preamp IC3-a, whose gain is fixed at 3.2. IC3-a output couples through C10 to a precision fullwave rectifier made up of IC1-a & -b, and their associated components. IC1-b output couples to a switchable polarity-inversion block made up of IC2-a, S2, & associated components. IC2-a output couples to a variable gain block made up of IC2-b and associated components.

IC1-b output also couples through C1 to a second fullwave rectifier made up of IC1-c, -d, and associated components. IC1-d output couples to a switchable polarity-inversion block made up of IC2-c, S3, & associated components. IC2-c output feeds to a variable gain block made up of IC2-d and associated components.

IC3-a output (the clean feed) ties to summing node

(IC3-b, pin 6) through R19-20; the ×2 feed ties to summing node through R10; and the ×4 feed ties to summing node through R12. R22 varies the output level. Signal couples through R21-C15 to the output path.

Use

Switches & pots have these functions:

S1	effect/bypass
S2	×2 invert
S3	×4 invert
R8	×4 level
R9	×2 level
R19	clean (×1) level
R22	master output level

Initial settings: S1 effect in; S2 & 3 either position; R8,

Fig. 8–1. Distort-O-Matic III prototype board.
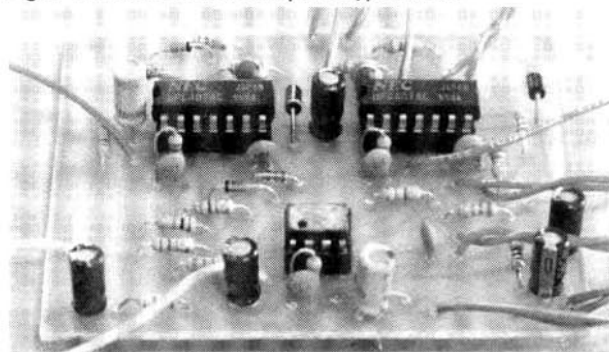
Fig. 8–2. Distort-O-Matic III circuit board.

3.5" × 3" reference box

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DISTORT-O-MATIC III PARTS LIST

Resistors
R1-4, 15-18 10K
R5, 6, 7, 10, 12-14, 20 22K
R8, 9, 22 100K audio-taper pot
R11 47K
R19 10K audio-taper pot
R21 100
R23 2.2K
R24 1K
R25 150K

Capacitors
C1, 10, 12, 15, 17, 18 10μF electrolytic
C2, 3, 8, 9, 11 100pF
C14 10μF nonpolar electrolytic
C4-7, 13, 16 10pF

Semiconductors
D1-4 1N914
D5, 6 1N4001
IC1, 2 TL074 quad op amp
IC3 TL072 dual op amp

Miscellaneous
¼" jacks
S1 DPDT switch
S2, 3 SPDT switches
solder, wire, 9V batteries, enclosure, etc.
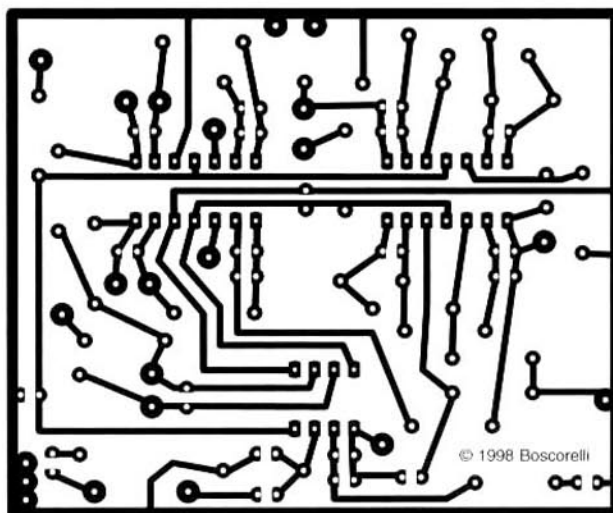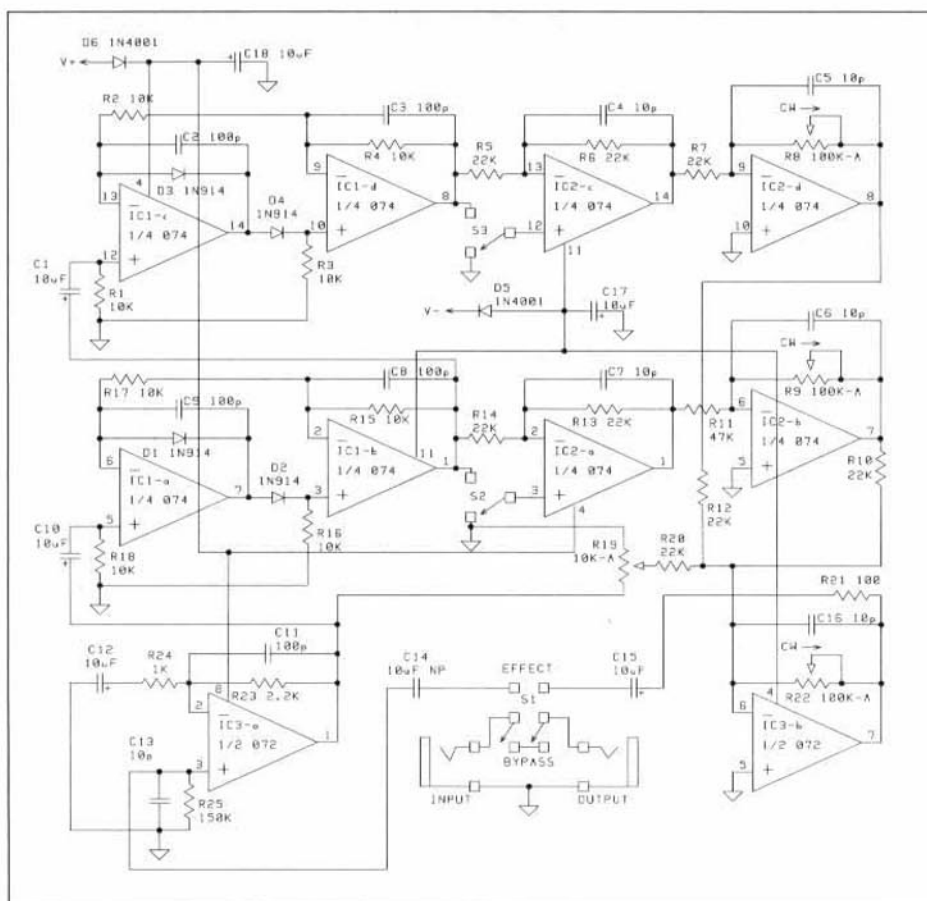
Fig. 8–3. Distort-O-Matic III schematic.

R9 fully CCW; R19, R22 straight up.

Connect unit to axe & amp, establish desired volume. Turn R19 fully CCW, slowly turn ×2 control R9 CW. Note the sound. Return R9 to fully CCW, slowly rotate ×4 level control R8 CW, note the sound.

Experiment with mixtures of clean and distorted feeds. Invert polarity of processed feeds and note the effect on the output mix.
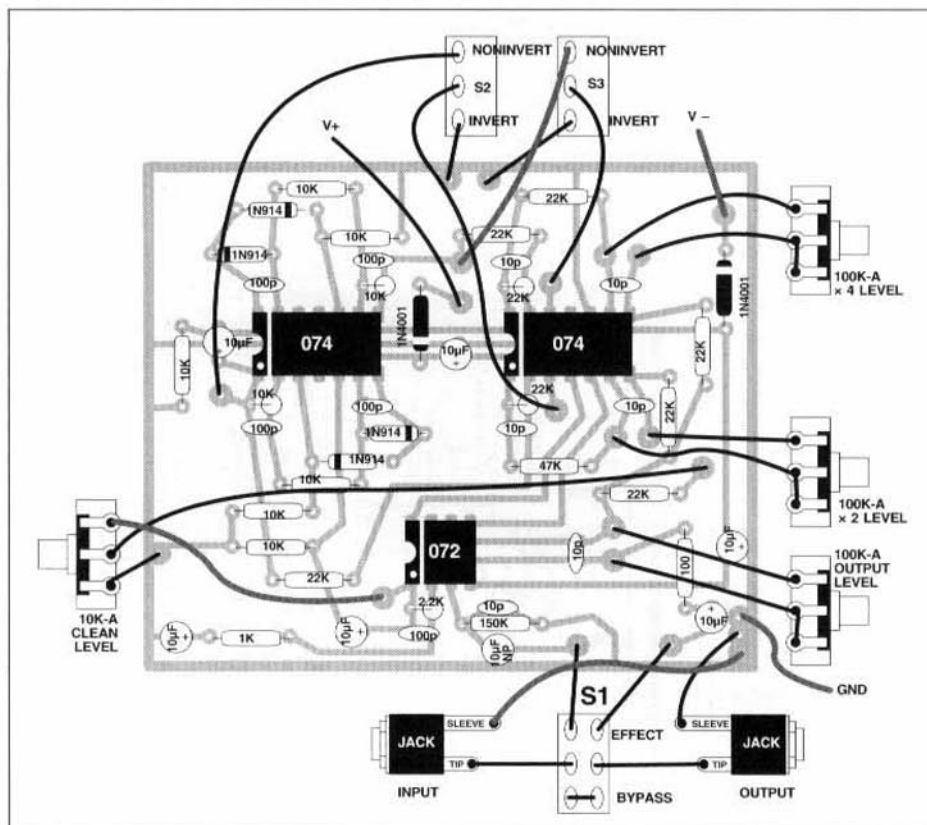
Fig. 8–4. Distort-O-Matic III layout & wiring diagram.
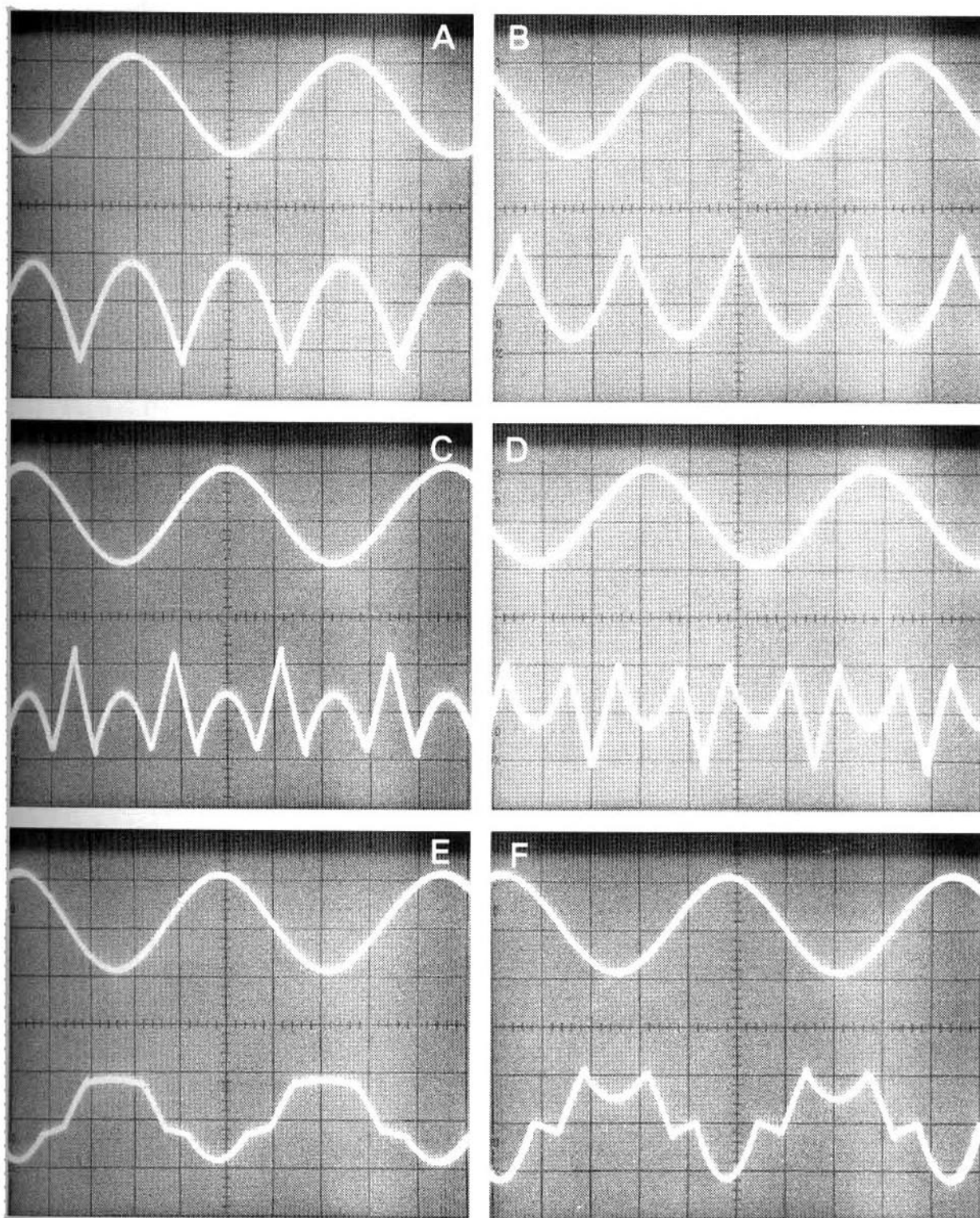
Fig. 8–5. DM3 I/O. All photos: top trace 1 KHz sinewave (input), bottom trace output, scale 1V/div. A—Clean & ×4 levels 0, ×2 level 100%; appearance of output is characteristic of fullwave rectification. B—×2 phase inverted. C—×2 & ×1 levels 0%, ×4 level 100%. D—×4 phase inverted. E & F—Variable mixtures of ×1, ×2, & ×4.

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Project No. 9
Direct-O-Matic

Going direct demands a clean drive stage and a balanced output. Direct-O-Matic delivers two channels of clean gain for guitar or bass, with balanced output and low-impedance drive capability.

Circuit Function

Both channels are identical; only channel 1 is described. Instrument feed couples through C1 to input of IC1-d, a noninverting amp whose gain varies 1–11 under control of R2. IC1-d output couples through R4-C4 to one balanced output terminal. IC1-d output also couples through R5 to unity-gain inverting buffer IC1-c, whose output couples through R7-C5 to the other balanced output. R8 & R9 act as bleeder resistors to prevent charge build-up in the output caps, which could cause a loud 'pop' in gear plugged into the charged caps. Both caps couple to polarity-inversion switch S1. S2 provides ground-lift.

Use

Pots & switches have these functions:

R2	channel 1 gain
R11	channel 2 gain
S1	channel 1 polarity invert
S2	channel 1 ground lift
S3	channel 2 ground lift
S4	channel 2 polarity invert

Connect unit to axe and target input device, establish desired level. Use is self-evident. If needed, preamp gain can be increased by changing R2 & R11 to any practical value.

To ensure ground-lift capability, use plastic XLR jacks with a metal case, or a plastic insert/case with metal XLR jacks.

While most quad op amps will work fine in this circuit, the 837 has studio-quality specs and can drive 600 ohms.

DIRECT-O-MATIC PARTS LIST

Resistors
R1, 10 1K
R2, 11 10K audio-taper pot
R3, 12 150K
R4, 7, 14, 15 51
R5, 6, 13, 18 22K
R8, 9, 16, 17 100K

Capacitors
C1, 4, 5, 10, 12, 13 10μF nonpolar electrolytic
C2, 11 10pF
C3, 9, 15, 16 10μF aluminum electrolytic
C6, 7, 8, 14 100pF

Semiconductors
D1, 2 1N4001
IC1 LM837 quad op amp

Miscellaneous
XLR jacks
S1, 4 DPDT switch
S2, 3 SPST switch
1/4" shorting jacks, wire, solder, circuit board, etc.

2.75" × 2.75" reference box

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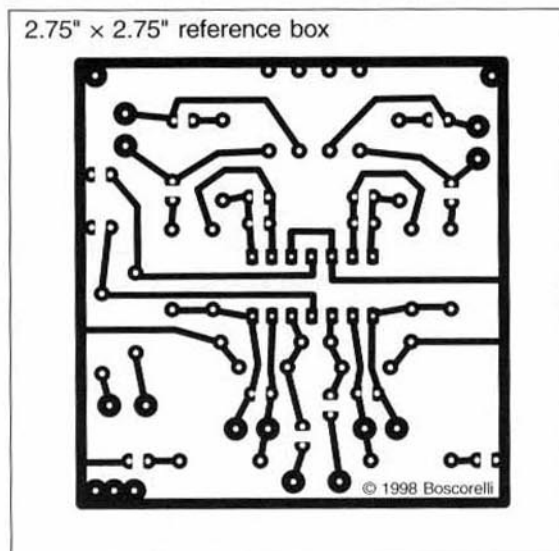
Fig. 9–1. Direct-O-Matic circuit board.
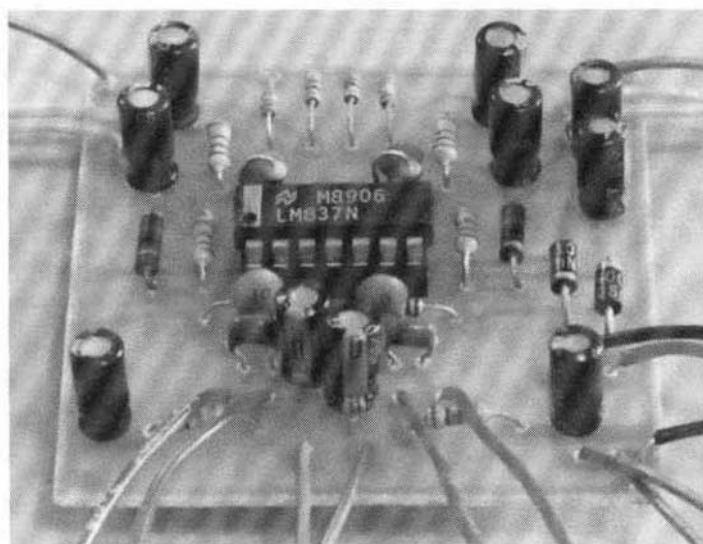
Fig. 9–2. Direct-O-Matic prototype board.

Fig. 9–3. Direct-O-Matic schematic.
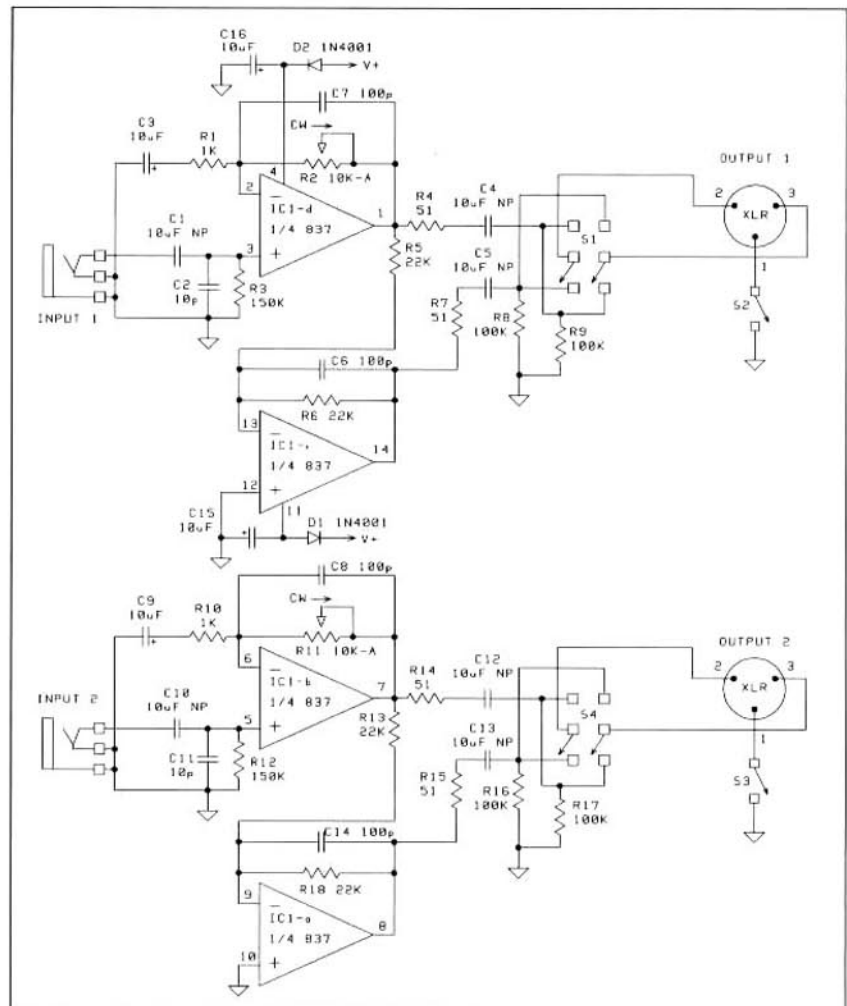
Fig. 9–4. Direct-O-Matic layout & wiring diagram.
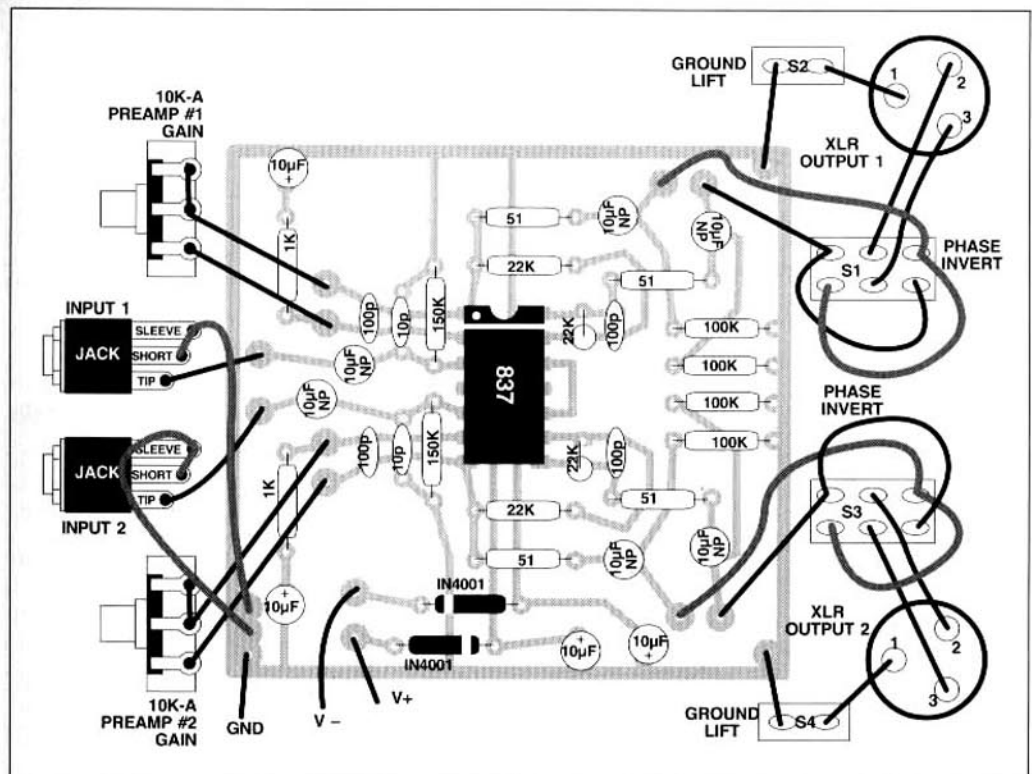
Project No. 10
Distort-O-Matic IV

DM4 lurks at the subtle end of the spectrum, definitely not a box to build if you hanker for flagrant fuzz. It might be a good idea to test the unit on the breadboard before etching a circuit board. This box suits only guitar, as its effects are too subtle for bass.

Circuit Function

Instrument feed couples through C8 to input of IC1-a, configured as a noninverting amp with gain of 3.7.

IC1-a output couples through R4 to inverting amp IC1-b, whose gain varies from 1–6 by dual pot R6 which, simultaneously, varies gain of inverting amp IC1-c from 1–$\frac{1}{6}$. Between the output of IC1-b and the input of IC1-c are interposed four unity-gain inverting amplifiers in series (IC2). The function of this block is detailed below.

IC1-a output also couples through R24-25 to IC1-d, configured by associated components R20-23 and D1-4 as a variable distortion device whose function is detailed below.

S2 selects output of either IC1-d (effect A) or IC1-c (effect B) and couples to input of inverting amp IC3 through C5-R19. IC3 gain is variable from 0–5 by pot R17. Signal couples through R18-C7 to the output path.

Use

Switches & pots have these functions:

S1	effect in/out
S2	effect select A/B
R6	effect B extent
R20	effect A shape
R24	effect A depth

Effect A generates a deadband with variable shape (R20) and variable extent (R24), resembling the crossover distortion seen in certain vintage tube amps. The effect is absent with both pots fully CCW.

Effect B takes advantage of distortion introduced by the LM324 on audio signals greater than $3V_{p-p}$. One type appears related to the 324's low slew rate; the other to crossover distortion in the output stage of the op amp. To accentuate these subtle effects, the signal passes through four 324-type op amps in series. Because both types of distortion are amplitude-dependent, dual pot R6 boosts the amplitude of the signal entering the 324 chain by a factor of up to 6, and simultaneously reduces gain by a factor of 6 at the end of the chain. This stage demands level management

DISTORT-O-MATIC IV PARTS LIST

Resistors
R1 1K
R2 2.7K
R3 150K
R4, 5, 7-16, 19 10K
R6 50K dual pot
R17 50K audio-taper pot
R18 100
R20 100K pot
R21, 22 1M
R23, 25 4.7K
R24 100K dual pot

Capacitors
C1, 6, 7, 9 10μF aluminum electrolytic
C2, 4 10pF
C3 100pF
C5 1 μF aluminum electrolytic
C8 10μF nonpolar electrolytic

Semiconductors
D1-4 1N914
D5, 6 1N4001
IC1 TL074 quad op amp
IC2 LM324 quad op amp
IC3 TL071 op amp

Miscellaneous
S1 DPDT switch
S2 SPDT switch
wire, printed circuit board, 9V batteries, solder, 1/4" jacks, etc.

Fig. 10–1. Distort-O-Matic IV prototype board.
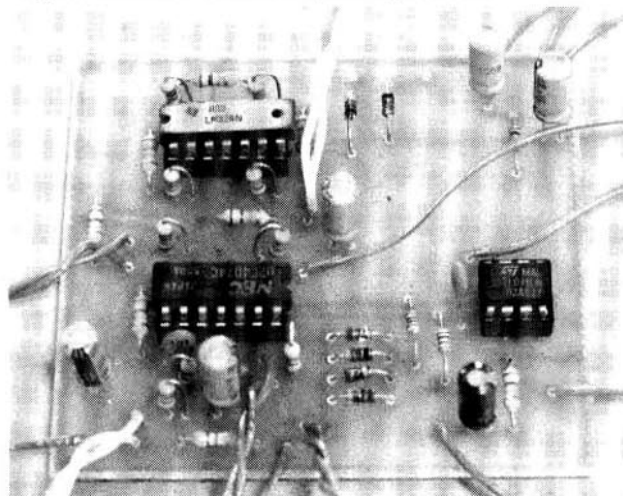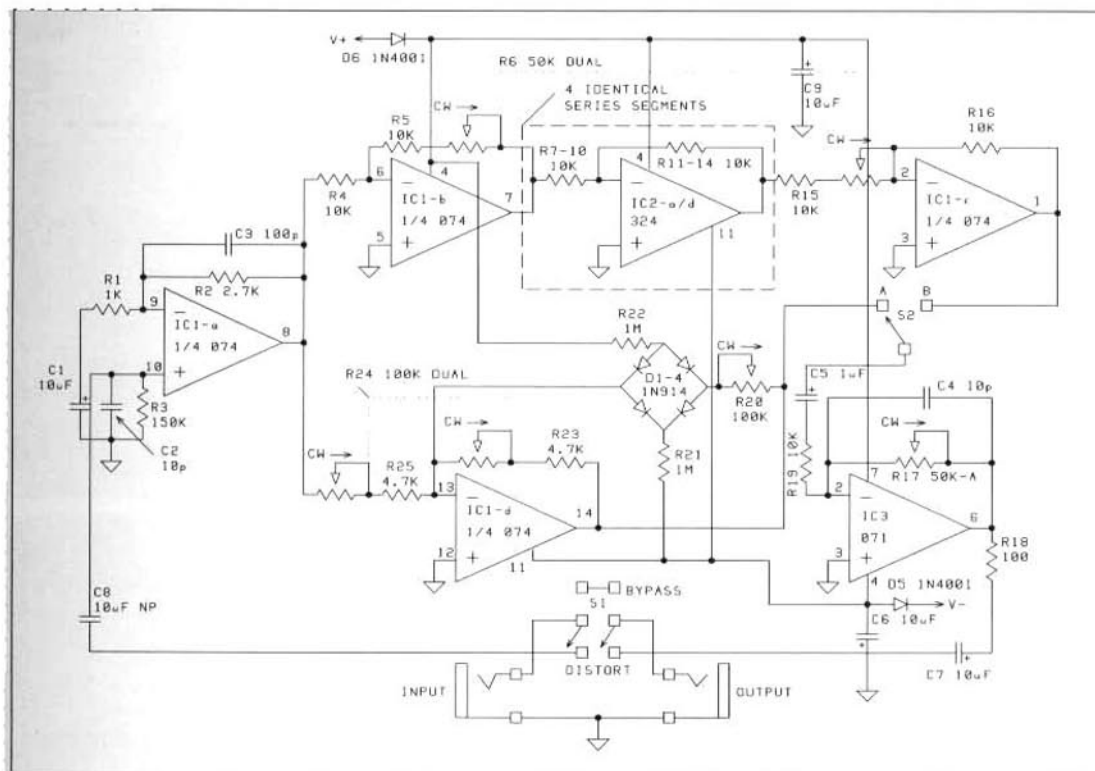
Fig. 10–2. Distort-O-Matic IV schematic.

to avoid clipping.

Effect B is subtle. Though dramatic in appearance on the scope, these changes affect only harmonics on some feeds, and may require treble boost from the amp to perceive. Only the LM324 generates this distortion. Most op amps act as unity-gain buffers that do not introduce audible distortion.
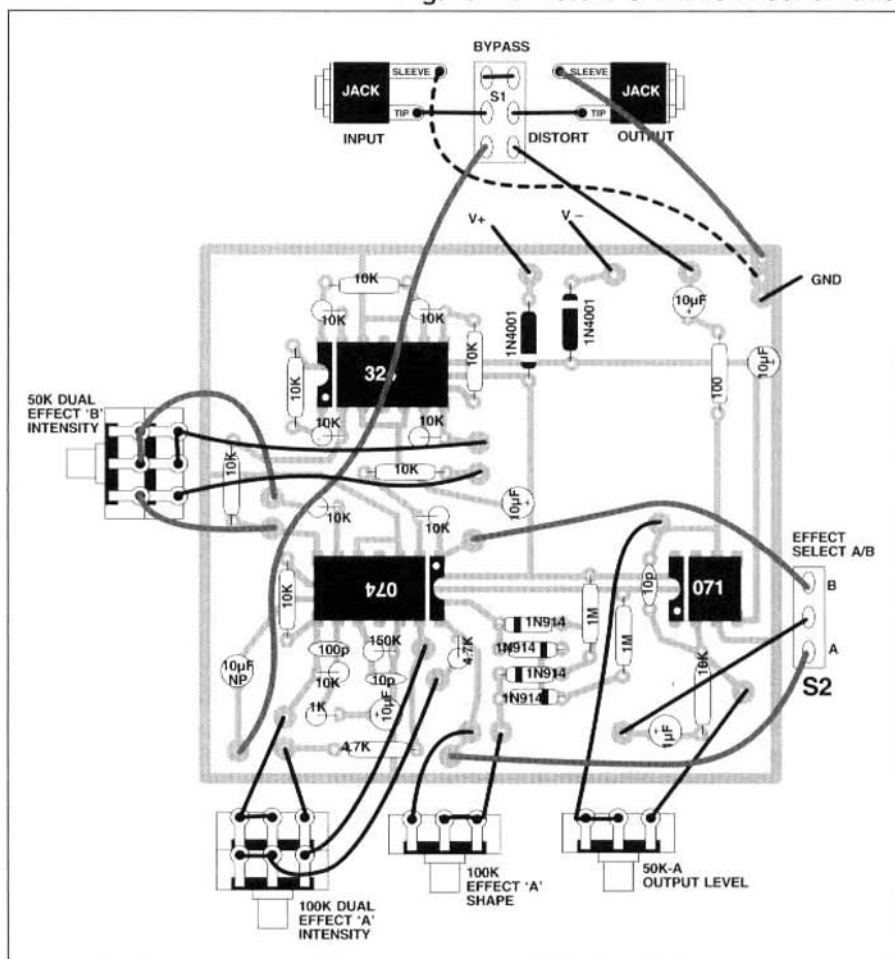
Fig. 10–3. Distort-O-Matic IV layout & wiring diagram.

37
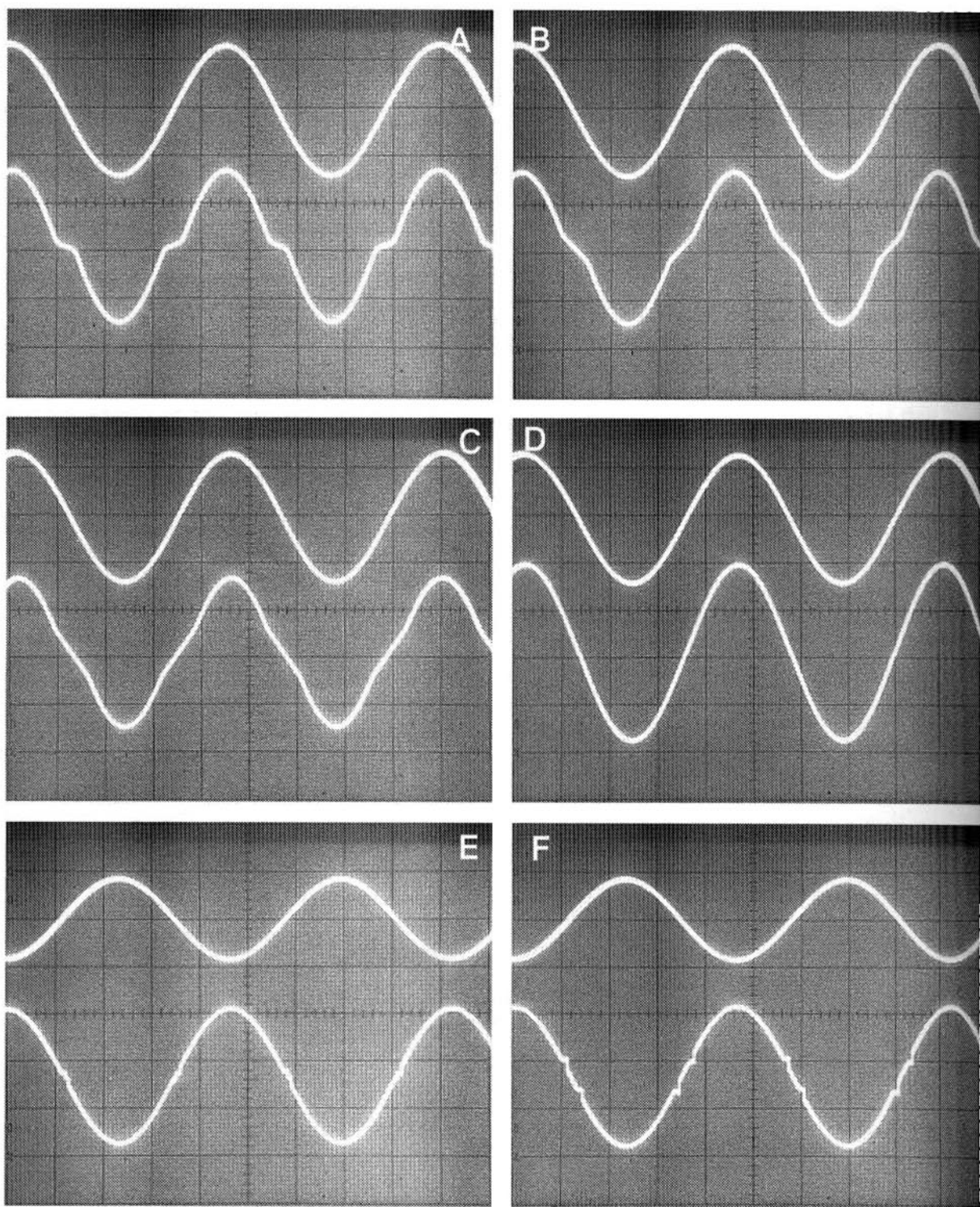
Fig. 10–4 (this and facing page) DM4 I/O. 1-KHz sinewave; scale top trace (input) 200 mv/div., bottom trace (output) 1V/div. All photos taken with preamp gain @ 11. High level needed to put effect A in best mode; low input level needed to avoid clipping when using effect B. A—Effect A, all controls fully CW. B—Shape control @ 50%. C—Shape control @ minimum. D—Shape back to max, extent control fully CCW; effect is absent. Facing page: DM4 I/O, effect B. Top trace 100 mv/div. (input), bottom trace 1V/div. (output). E—1 KHz, R6 fully CW. F—1 KHz, R6 fully CCW. G—3 KHz, R6 fully CW. H—3 KHz, R6 fully CCW. I—10 KHz, R6 fully CW. J—10 KHz, R6 fully CCW.
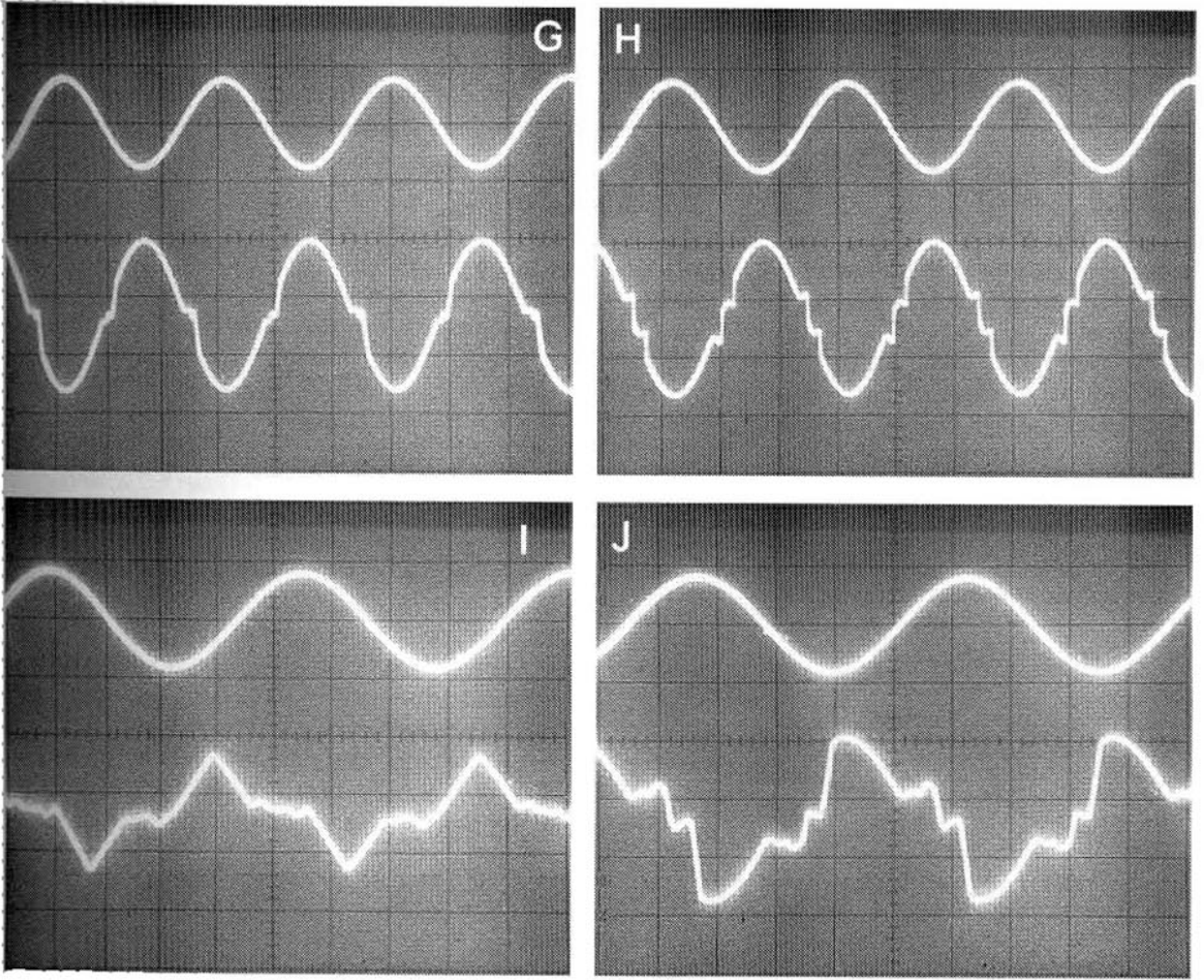
Fig. 10–5. Distort-O-Matic IV circuit board.

3.5" × 3.25" reference box

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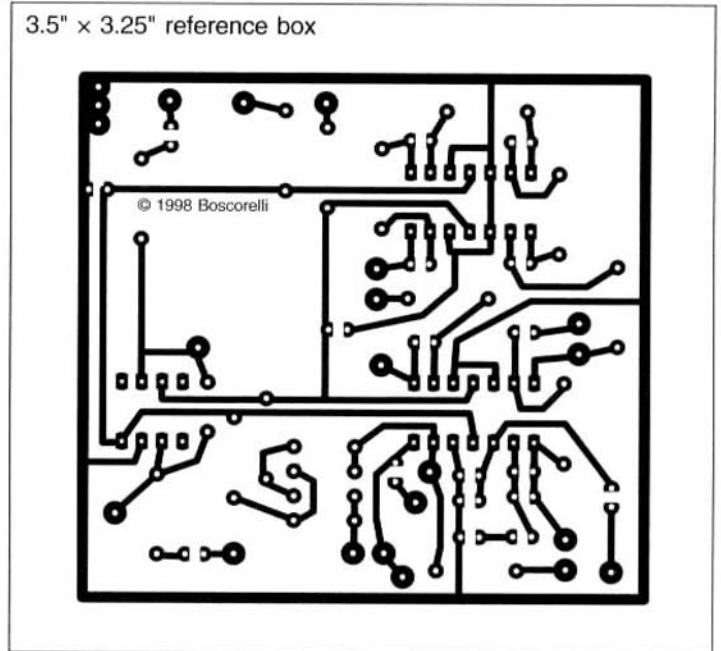
Project No. 11
Super Play-Along

A play-along is a mixer that blends an instrument feed with a stereo feed from a CD player, tape deck, or other source, enabling one to "play along" with the feed; turns any stereo system into a practice amp. Incorporates switchable highpass filter for bass, and limiter for bass and guitar. Not exactly a stomp box, but something many players find handy.

Circuit Function

Signal Path: Instrument feed couples through C1 to

Fig. 11–1. Super Play-Along prototype board.
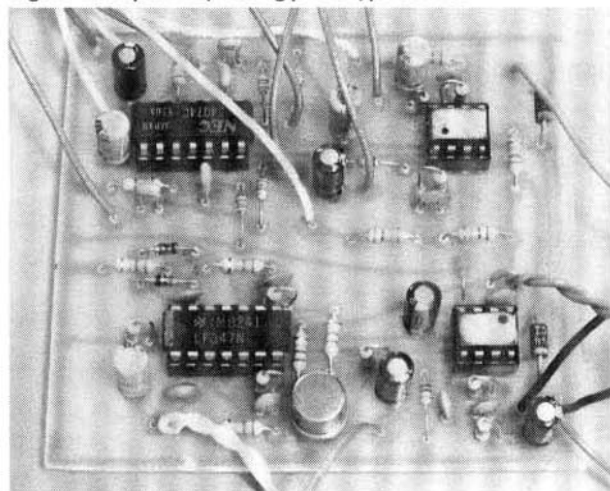
Fig. 11–2. Super Play-Along circuit board.

3.75" × 3.5" reference box

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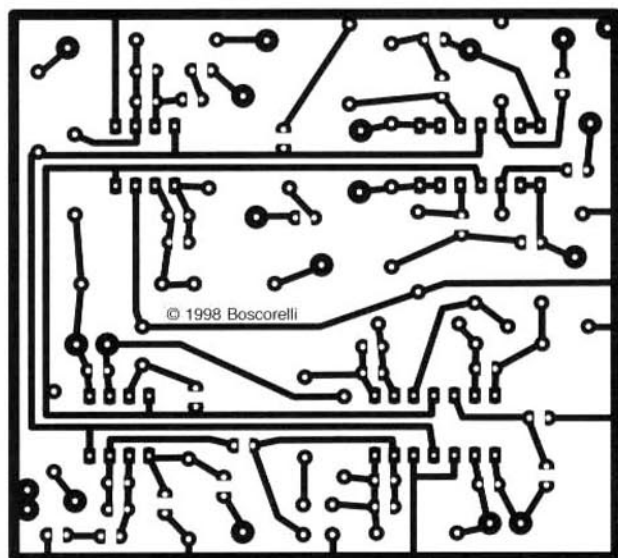
IC1-a, a preamp with gain of 5.7. IC1-a output couples through R3 to IC4-a, an inverting amp whose gain varies with the state of the LDR in the feedback loop. IC4-a output couples through R6 to variable gain stage IC1-b, whose output couples to inputs of IC3-a & -b.

Each line-level signal path is identical, so only the left channel will be described. Line-level input couples through C19 to AC voltage follower IC2-a, whose output couples to one throw of S1, and to a highpass filter made up of IC2-b and surrounding components. S1 selects between straight and highpass feed to couple to the inputs of output buffers IC3-a & -b. Audio couples through R16-C11 to the output path.

Control Path: IC4-a output couples through R8 to a level detector made up of variable gain stage IC4-b and a precision fullwave rectifier made up of IC4-c & -d & surrounding components. IC4-d output couples through R14 to the LED in the LDR. The result of this arrangement is a downward compressor whose

SUPER PLAY-ALONG PARTS LIST
Resistors
R1 1K
R2, 14 4.7K
R3, 4 22K
R5 150K
R6, 8, 10, 11-13, 15, 17-20, 22 10K
R7 50K audio-taper pot
R9 50K reverse-audio pot
R16, 21 100
R23-26 6.8k
R27, 28 47K
Capacitors
C1, 19, 20 10μF nonpolar
C2, 8, 11, 12, 21, 22 10μF aluminum electrolytic
C3, 5, 6, 7 10pF
C4, 9, 10, 13, 14 100pF
C15-18 0.1μF
Semiconductors
D1, 2 1N914
D3, 4 1N4001
IC1, 2 TL074 quad op amp
IC3, 4 TL072 dual op amp
Miscellaneous
S1 DPDT switch
Vac-Tec VTL2C2 optocoupler
RCA jacks, ¼" phone jack, wire, solder, circuit board, etc.
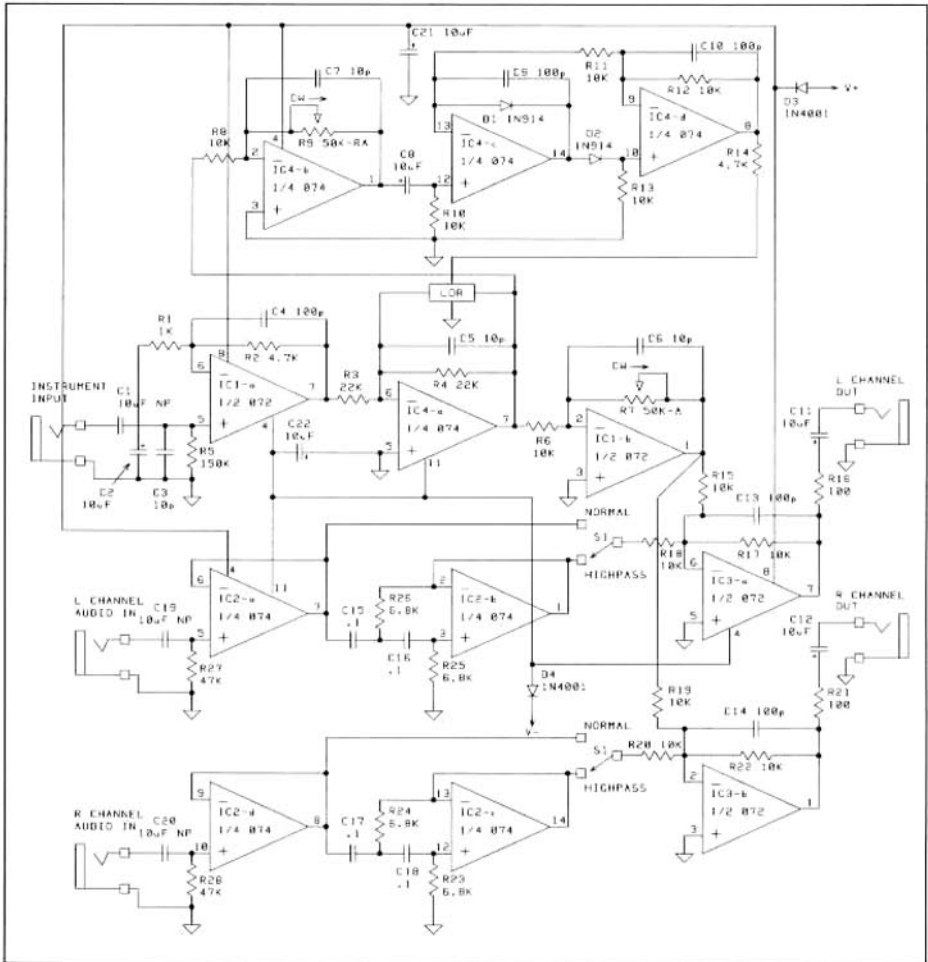
Fig. 11–3. Super Play-
Along schematic.

Fig. 11–3. Super Play-Along schematic.

threshold is set by R9.

Use

Switch & pots have these functions:
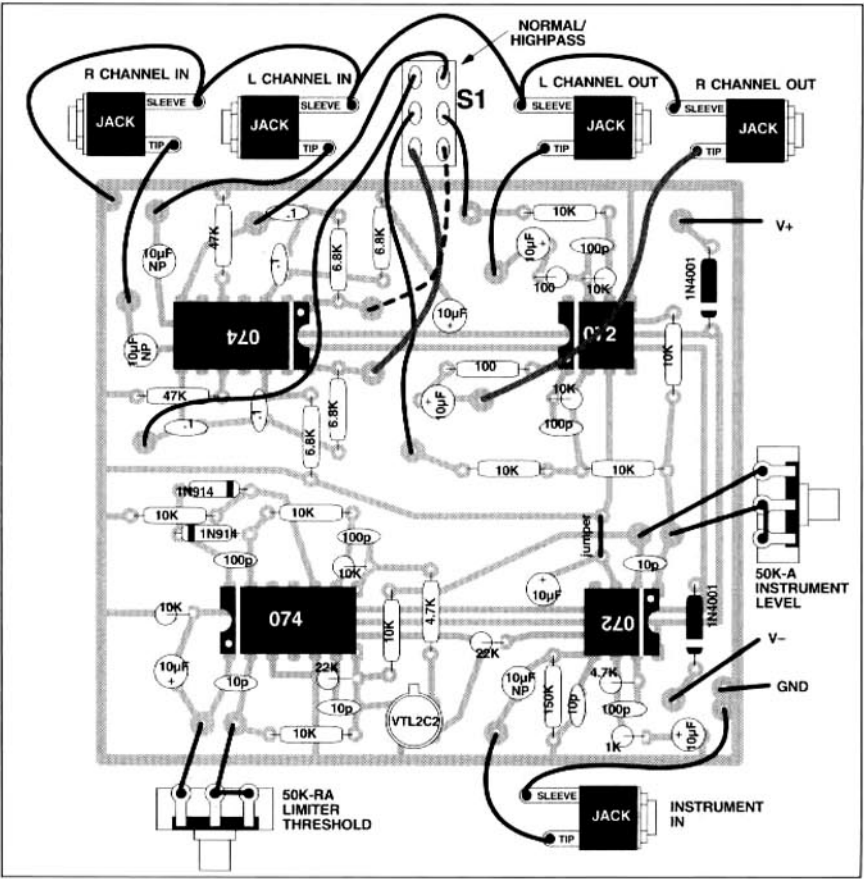
S1 highpass filter
R7 instrument volume
R9 limiter threshold

Initial settings: S1 highpass out, R7 straight up, R9 fully CW. Connect unit to axe and stereo (e.g., the tape loop of an FM receiver). Power up, establish desired program volume and instrument volume.

Turn R9 CCW until limiting action is perceived. If compression is too severe, change R14 to 10K or 15K.

Switch highpass filter in/out to note effect on stereo feed. This feature lets bass players kill bass tones in the stereo feed.

Fig. 11–4. Super Play-Along layout & wiring diagram.

Project No. 12
Pan Tremolo-Matic

Same basic specs as Tremolo-Matic, but generates stereo output that crossfades between channels. When one channel intensifies, the other softens. Offers option to invert polarity of one output, lending the sound a phasey air. Special effect that suits many instruments, including vocals.

Circuit Function

Signal Path: Instrument feed couples through C16 to input of IC1, a preamp whose gain varies 1–21 depending on setting of R36. IC1 output couples through C4-R34 and C3-R22 to different channels of IC4, an NE570 configured as two VCAs, each of whose gain varies 0–1, depending on the control voltage input. IC4's two outputs couple to inverting buffers IC2-a & -b that yield a net noninverting signal path. S2 allows polarity inversion of Output 2. Dual pot R24 controls the output level. Signals couple through R30-C10 and R29-C11 to their respective outputs. R20 and R32 trim feedthrough for their respective channels.

Control Path: IC5 and its associated components form a sinewave oscillator whose frequency is variable by R1 from ~1–10 Hz. IC5-b output couples to pot R9 which varies the sinewave level feeding two paths, one of which inverts the control voltage. Pot R14 applies an identical static DC voltage to the inverting inputs of IC6-a & -c. The result of the control path is that R14 sets an identical resting DC voltage applied to both VCA control ports of IC4 (R19/pin 16, R35/pin 1). The sinewave feed is inverted between the channels. When one VCA sees a peak and increases gain, the other VCA sees a trough and reduces gain. This crossfades the signal between channels.

Use

Pots and switches have these functions:

S1	effect/bypass
S2	output 2 polarity invert
R1	tremolo rate
R9	tremolo depth
R14	static gain of both VCAs
R24	output level
R36	preamp gain

Set R36 at minimum; short the input. Set R1 for maximum rate; trim the sine generator (R7) to give $3V_{p-p}$ at pin 1 of IC5-b. Center R14, turn R9 fully CW. In this state the control voltage clips at both extremes. Attach scope probe to pin 10 of IC4, trim R20 for minimum feedthrough; move scope probe to pin 7 of IC4,

trim R32 for minimum feedthrough.

If no scope is available, trim feedthrough by ear. Connect Output 1 to an amp whose volume is turned all the way down; set R24 fully CW. Slowly advance

PAN TREMOLO-MATIC PARTS LIST
Resistors
R1 2M reverse-audio pot
R2, 38 150K
R3, 4, 8 2.2K
R5, 19, 35 47K
R6 1.5K
R7 1K multiturn trimpot
R9 10K pot
R10, 11, 12, 18 33K
R13, 15, 16, 17 100K
R14 100K pot
R20, 32 100K multiturn trimpot
R21, 33 220K
R22, 34 4.7K
R23, 31, 38, 39 36K
R24 10K dual audio-taper pot
R25, 26, 27, 28 10K
R29, 30 100
R36 10K trimpot
R37 470

Capacitors
C1, 2 1μF tantalum, 10%
C3, 4, 10, 11, 15, 18, 19 10μF aluminum electrolytic
C5, 6 22pF
C7, 8 1μF nonpolar electrolytic
C9, 12, 14 100pF
C13 0.1μF
C16 10μF nonpolar electrolytic
C17 10pF

Semiconductors
D1, 2 1N914
D3 1N4001
IC1 OP-27 low-noise op amp
IC2 TL072 dual op amp
IC3 78L05 5V positive regulator (TO-92)
IC4 NE570/571 dual-channel compander
IC5 TL062 dual low-power op amp
IC6 TL064 quad low-power op amp

Miscellaneous
S1 DPDT switch
S2 SPDT switch
solder, wire, circuit board, 9V battery snap mounting hardware, ¼" jacks, etc.
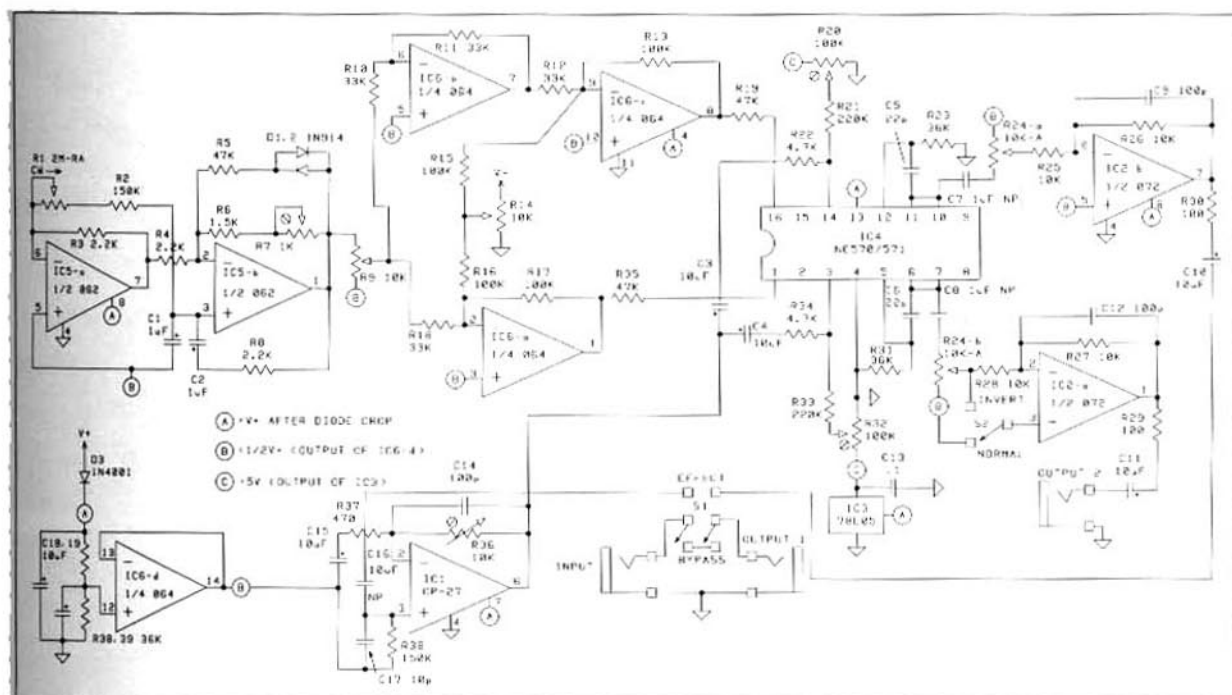
Fig. 12–1. Pan Tremolo-Matic schematic.

Fig. 12–2. Pan Tremolo-Matic circuit board.
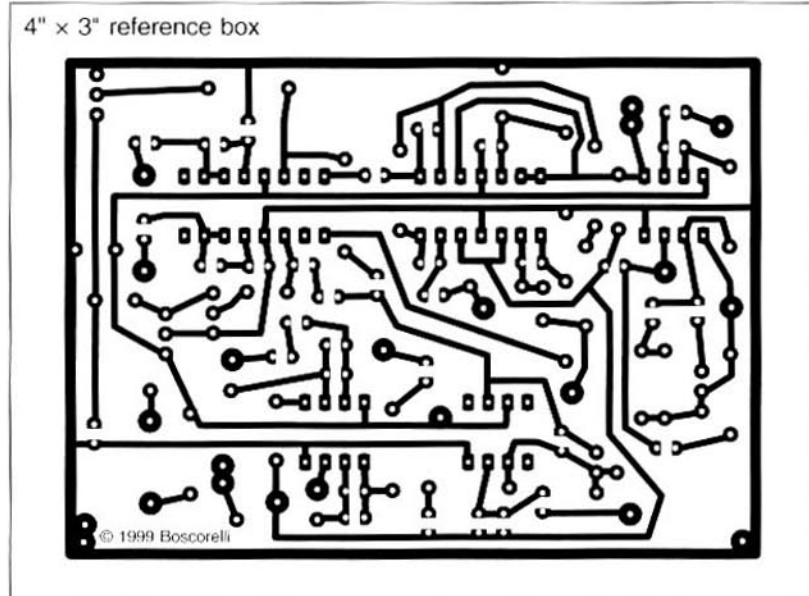
amp volume control until feedthrough artifacts are heard. Depending on initial setting of feedthrough trimpots, this could reach several volts$_{p-p}$. Trim R20 for minimum feedthrough. Turn amp volume all the way down, connect Output 2 to amp input, slowly advanced amp volume until feedthrough artifacts are heard, trim R32 for minimum feedthrough.

Initial settings: S1 effect in, S2 normal; R1 fully CW; R14 straight up; R9, R24, R36 fully CCW. In this state static gain of both VCAs is about 0.5. Connect unit to axe and stereo amp, turn R24 fully CW, trim R36 for desired preamp gain.

Slowly advance R9 and confirm that signal cross-fades between channels. Toggle S2 to note the effect of inverting phase of one channel. Test the three types of tremolo effects described under Tremolo-Matic; note the way stereo dramatizes them.

Notes

PTM is a 9V box designed to run on 7.5V at the positive supply (after drop across D3). It can run at higher voltages, but R23 & R31 will have to be changed to keep DC output bias of both VCAs near $1/2$V+.

The sinewave oscillator may take a few seconds to start up. Set R36 to give an average preamp output of ~1V$_{p-p}$. This leaves reasonable headroom.

43

2M-RA RATE

150K
2.2K
2.2K

002

jumper
jumper
36K

10µF +

100K
35K

220K
36K
22p

100K multiturn

47K
33K
4.7K

064
570

100K multiturn

10µF NP

10K DEPTH

1.8K
47K

1µF
100K
100K
47K
33K
33K
33K

10µF +

100K
10µF +

4.7K
36K
220K

22p

10µF NP
10µF +

jumper

78L05

10K-A DUAL OUTPUT LEVEL

2.2K

1N914
1N914

10K multiturn

1µF

100p

10K

OP-27

100p
10µF +

072

jumper

100

10K

1N4001

V+

10K STATIC GAIN

1K multiturn

10µF +

150K
10p

470

100

10K

10µF +

100p

10K

GND

JACK
INPUT

SLEEVE
TIP

S1

BYPASS

SLEEVE
TIP

JACK
OUTPUT 1

SLEEVE
TIP

JACK
OUTPUT 2

INVERT

S2

NORMAL

Fig. 12–3. Pan Tremolo-Matic layout & wiring diagram.
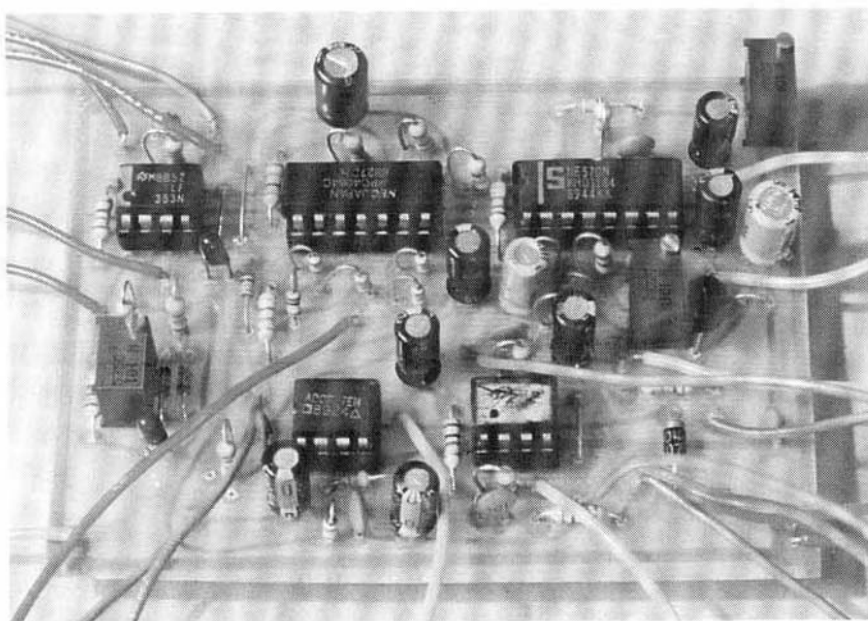
Fig. 12–4. Pan Tremolo-Matic prototype board.
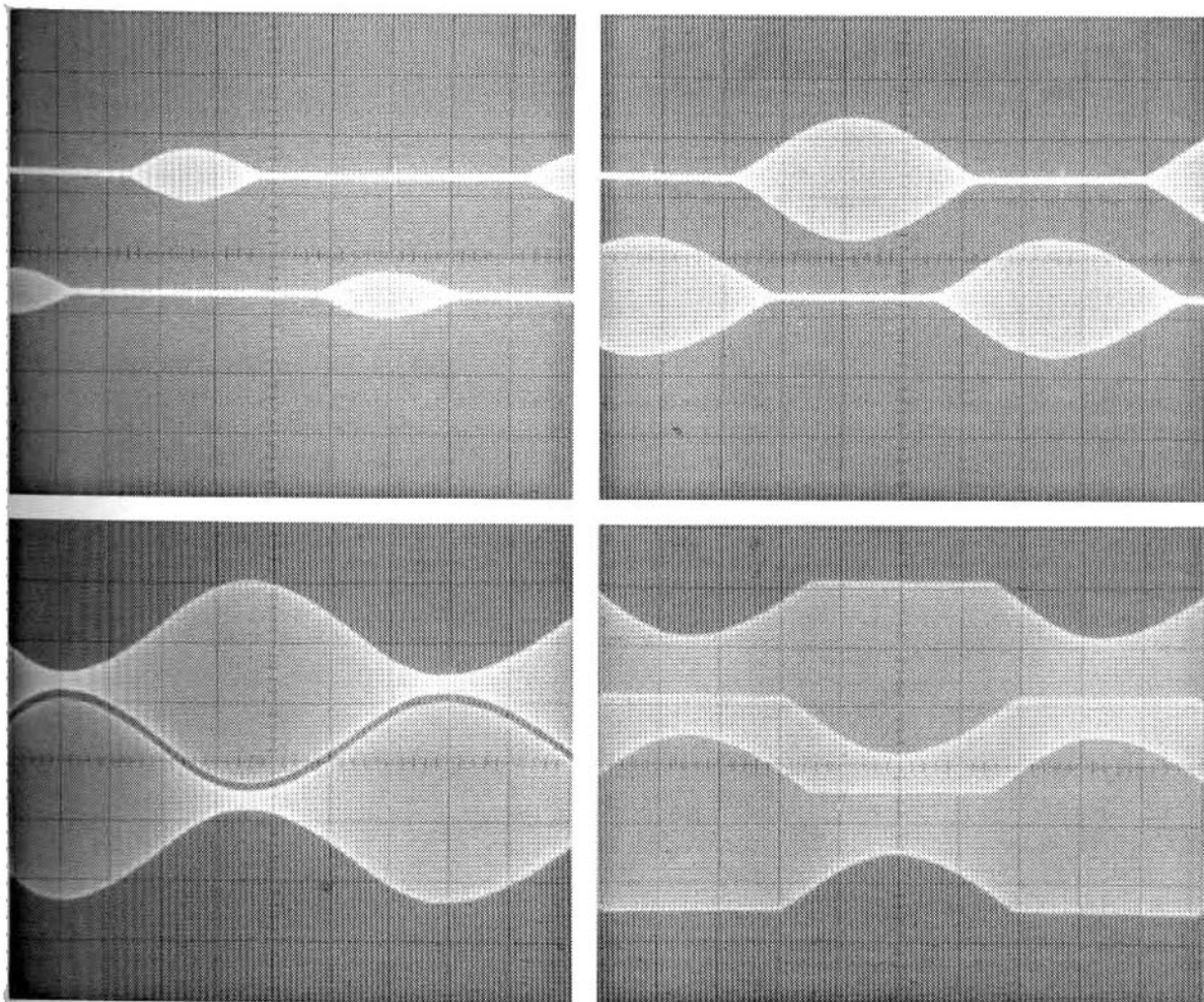
Fig. 12–5. Pan Tremolo-Matic simultaneous output photos. Scale 1V, sweep 20 ms. Unit generates the same spectrum of modulation modes as Tremolo-Matic. Pulse artifact is due to external sync generator.

Project No. 13
Tone-O-Matic

Tone-O-Matic duplicates the passive and active tone control functions of the original Ampeg® SVT.

Circuit Function

Instrument feed couples through C2 to input of noninverting preamp IC1-a whose gain is, nominally, 3.7. IC1-a output couples through R4-C6 to a switched passive tone-shaping network made up of R6-9, C7-9, and S2. S3 and C10 form the ultra-high boost circuit, pot R10 acts as master volume control, whose output is buffered by IC1-b.

IC1-b output couples through C11 to passive bass/treble control network comprised of R11-15 & C12-16. The output is taken at the juncture of R14 and the wiper of R15. Signal couples through C14 to buffer IC1-c, biased by R16.

IC1-c output couples directly to a network made up of IC1-d; R17, 18, 20, 21; C17-18, T1, & S4, that mimics the function of the original SVT 'midrange' control.

Signal couples through R19-C1 to the output path.

Use

Pots and switches have these functions:

S1	effect/bypass
S2	bass cut/bypass/ultra-low
S3	ultra-hi
S4	midrange select 3000 Hz, 800 Hz; or shelf below 220 Hz
R10	master volume
R12	bass
R15	treble
R20	midrange boost/cut

Initial settings: S1 effect in, S2 centered, S3 open, S4 centered; R10, 12, 15, & 20 centered.

Connect unit to axe and amp, establish desired volume. If possible, bypass or neutralize the amp's tone circuits.

S1 acts as a straight wire feed when centered. In "bass cut" position, it attenuates frequencies below 100 Hz. The "ultra-low" setting engages a 20-dB notch centered near 600 Hz. Test the effect of the "ultra-hi" switch with volume pot R10 centered, as it will have no effect with R10 fully CW. Test effect of bass & treble controls.

Return to initial settings. Take R20 through its full range, with S4 successively in the three positions. The control applies roughly 12 dB of boost when fully CW, 12 dB of cut when fully CCW. The midrange frequencies are centered near 800 and 3000 Hz; the bass shelf cuts in below 220 Hz.

The 47K value of R4 preserves some of the loading found in the original circuit, at the cost of noticeable signal loss. To compensate for this loss, either in-

TONE-O-MATIC PARTS LIST

Resistors
R1 1K
R2 2.7K (see text)
R3, 6, 7 150K
R4 47K
R5 100K
R8 820K
R9 68K
R10, 12, 15 1M audio-taper pot
R11 220K
R13 22K
R14 120K
R16 1M
R17, 18 10K
R19 100
R20 50K pot
R21 2.2K

Capacitors
C1, 4, 19, 20 10μF aluminum electrolytic
C2 10μF nonpolar electrolytic
C3 10pF
C5 100pF
C6, 11 0.1μF
C7, 8, 9 0.002μF
C10 500pF
C12 470pF
C13 0.0047μF
C14, 16 0.01μF
C15 0.001μF
C17 0.033μF
C18 0.2μF

Semiconductors
D1, 2 1N4001
IC1 MC33174 quad op amp

Miscellaneous
T1 600:600 ohm transformer, Mouser p/n 42TL016 or equivalent
S1 DPDT switch
S2, 4 DP3T switch
S3 SPST switch
wire, jacks, knobs, solder, battery connectors, circuit board, etc.

crease R2 to raise preamp gain, or change R4 to 100 ohms, or do both. The 2.7K value of R2 is nominal. The builder should feel free to change it to suit the output of a given axe. When the prototype was tested with a low-output axe, R2 was increased to 22K, boosting gain to 23.

The treble control and the bright switch are quite aggressive. The bass control has little effect on the sound of guitar, kicking in mainly below 60 Hz, apropos of its bass amp origins. The "ultra-low" setting of S2 sucks out a lot of signal. The box needs extra preamp gain if you like the ultra-low sound. The three mid-range settings sound distinctive; boost & cut are not subtle. This box likes more headroom than a couple of 9V batteries can supply, making ±15V or even ±18V an attractive option.
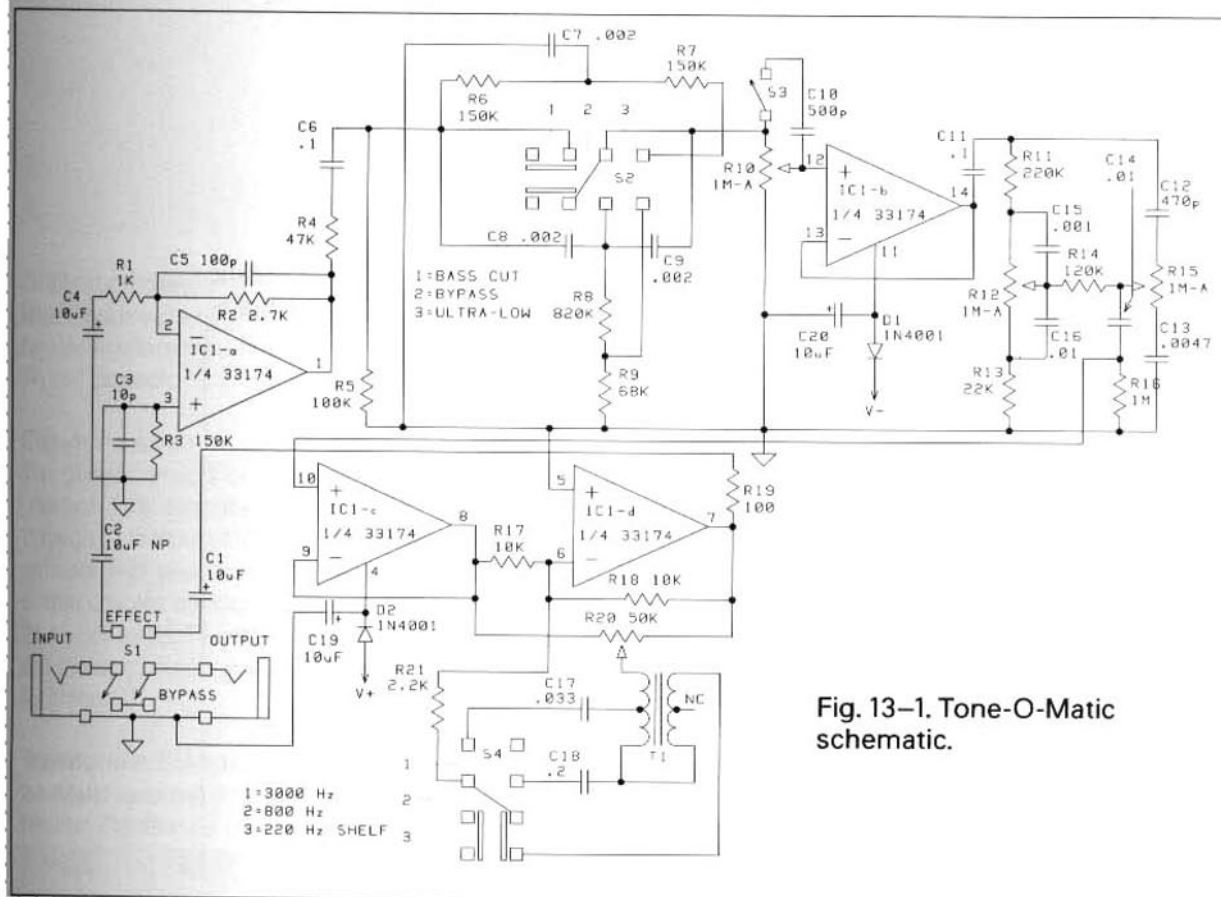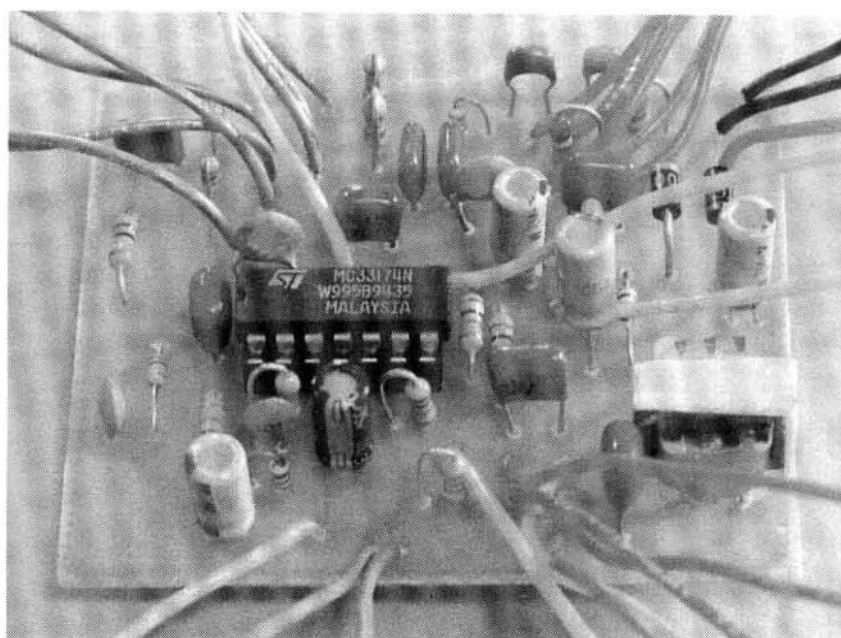
Fig. 13–1. Tone-O-Matic schematic.

Fig. 13–2. Tone-O-Matic prototype board.

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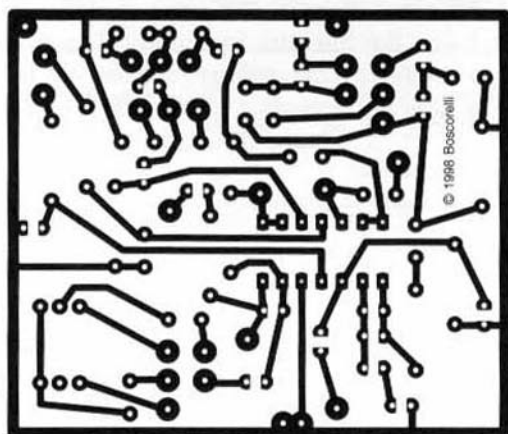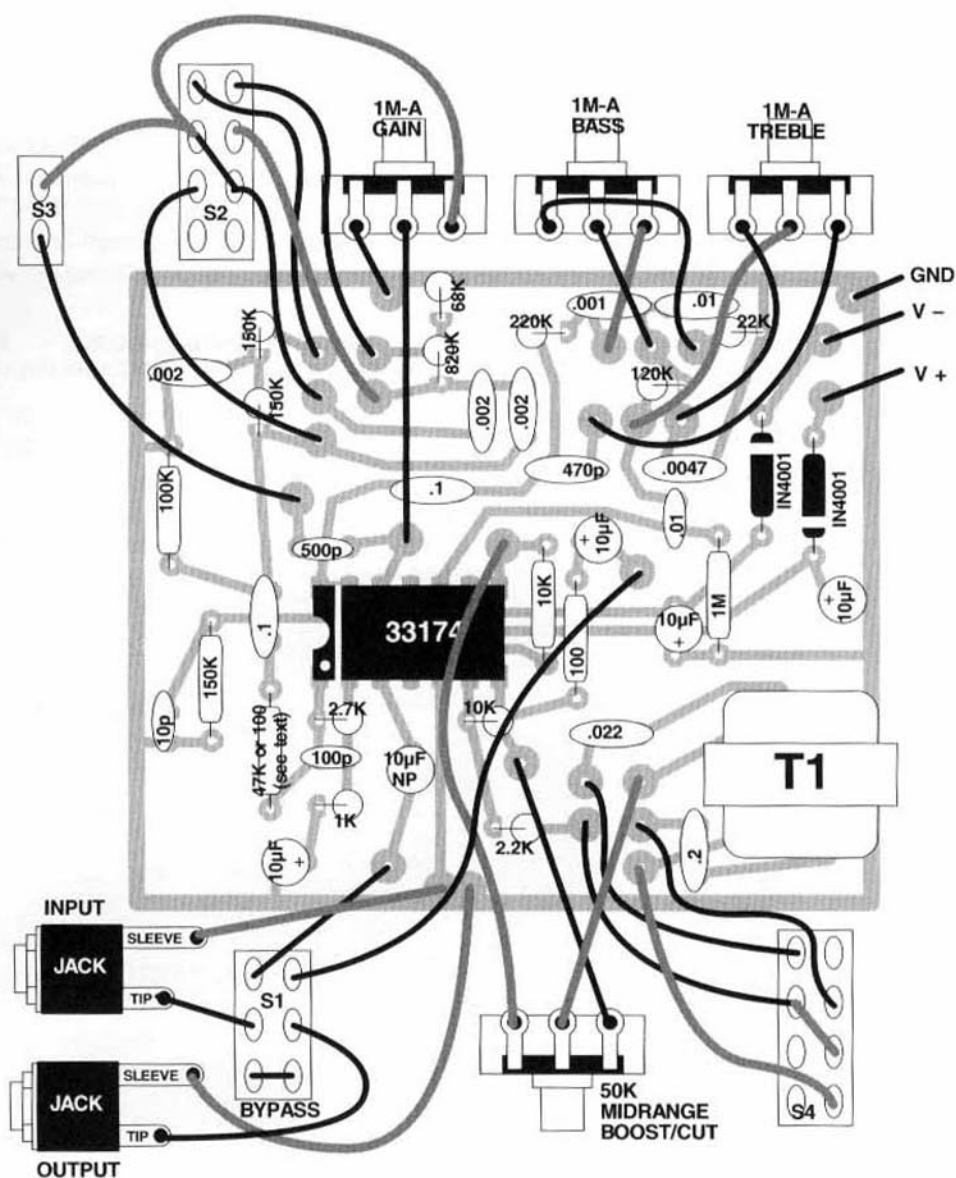
Fig. 13–3. Tone-O-Matic circuit board.

Fig. 13–4. Tone-O-Matic layout & wiring diagram.

Project No. 14
Iso-Matic

Of all direct boxes, only a transformer provides full isolation that breaks otherwise intractable ground loops. Iso-Matic offers two channels with gain enough for a sixties' Gretsch.

Circuit Function

The circuit consists of two identical segments; only channel 1 is described. Instrument feed couples through C1 to input of IC1-a, a preamp whose gain varies from 1–21 depending on the setting of R2. IC1-a output couples through R4-C5 to primary winding of T1, loaded by R5. T1 secondary couples to an XLR output through polarity-reversal switch S1. S2 provides for ground lift.

Transformer Selection

Iso-Matic uses the small audio transformers sold by Mouser Electronics (Fig. 14–2). While these pieces lack the specs of studio standards, they weigh in quite a bit lighter, in every sense, $2–$3 a pop. Choice depends on anticipated operating level and frequency range. At 20 Hz, the 42TL016 begins to saturate at about 800 mv_{p-p}; the 42TM016 at around 2V_{p-p}; the 42TU016 just below 4.5V_{p-p}. None of the three transformers exhibited obvious distortion at 25V_{p-p}, over the range 1 KHz – 20 KHz. The circuit board provides pads for each transformer's footprint; drill only the pads that fit the transformers you select.

Mouser 019-series transformers match 10K to 600 ohms. At 20 Hz, they saturate at significantly higher voltage than do members of the 016-series, but incur ~12 dB of voltage loss, easily recovered elsewhere in the gain chain. Choice of -016 or -019 series is a matter of personal preference.

These transformers lack shielding, making a ferrous enclosure helpful. Be aware that the box may be sensitive to placement and orientation.

ISO-MATIC PARTS LIST

Resistors
R1, 9 1K
R2, 8 20K audio-taper pot
R3, 10 150K
R4, 7 100
R5, 6 4.7K

Capacitors
C1, 10 10μF nonpolar electrolytic
C2, 6, 9, 12 10μF electrolytic
C3, 11 10pF
C4, 8 100pF
C5, 7 22μF electrolytic

Semiconductors
D1, 2 1N4001
IC1 NE5532 dual op amp

Miscellaneous
S1, 3 DPDT switch
S2, 4 SPST switch
T1, 2 transformer (see text)
¼" shorting jacks, XLR jacks, wire, solder, circuit board, etc.

3" × 3" reference box

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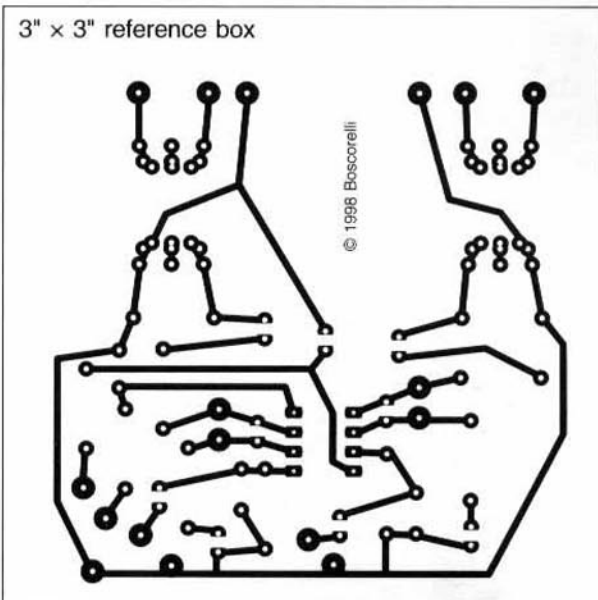
Fig. 14–1. Iso-Matic circuit board.

While both transformer windings offer nominally identical impedance, greater consistency results from using the higher-resistance winding for the primary. This is marked 'P' on the body of the transformer. The prototype board used a 42TU016 for channel 1, a 42TM016 for channel 2.

To retain ground-lift capability, use plastic XLR jacks with a metal case, or a plastic case/mounting panel with metal XLR jacks.

Use

Switches & pots have these functions:

R2	channel 1 gain
R8	channel 2 gain
S1	channel 1 polarity
S2	channel 1 ground lift
S3	channel 2 polarity
S4	channel 2 ground lift

Operation is straightforward and self-evident. The output need not use XLR jacks; $\frac{1}{4}$" TSR jacks work fine. You can also wire an unbalanced output in parallel with the balanced output.

Fig. 14–2. Top photo—Diagonally L–R, 42TL016, 42TM016, and 42TU016. The -019 series is outwardly similar. Bottom photo—Iso-Matic prototype board.
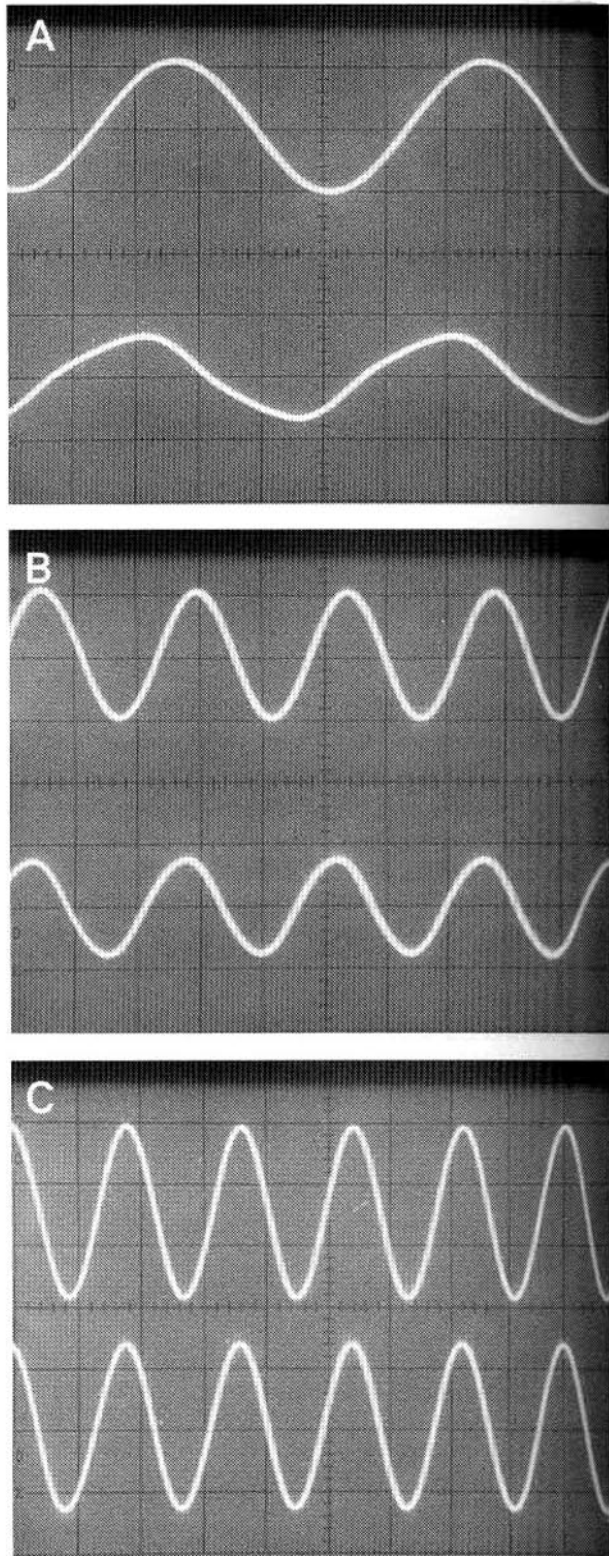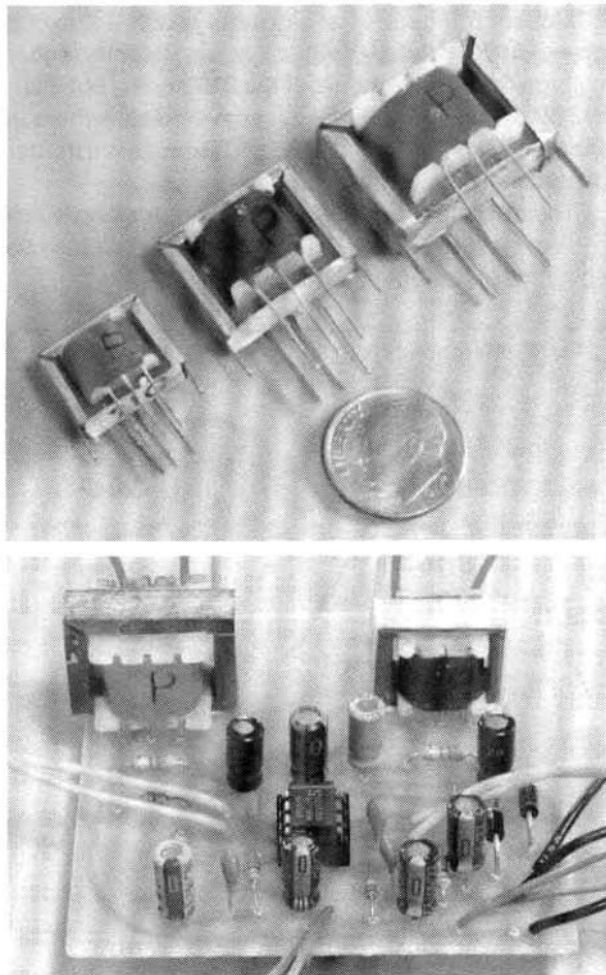
Fig. 14–3. I/O photos of 42TL016 transformer. Scale A & B: 500 mv/div. A—20-Hz sinewave input (top trace) shows obvious distortion (bottom trace output). B—50-Hz sinewave at same level shows much less distortion. Slight amplitude loss is due mainly to loading effect with 100-ohm series resistor, and is easily made up by gain of op amp. C—Scale changed to 10V/div.; transformer passes $27V_{p-p}$ 1-KHz sinewave virtually undistorted. The 42TM016 and 42TU016 tolerate higher amplitudes at low frequencies due to the higher saturation threshold of their larger cores.
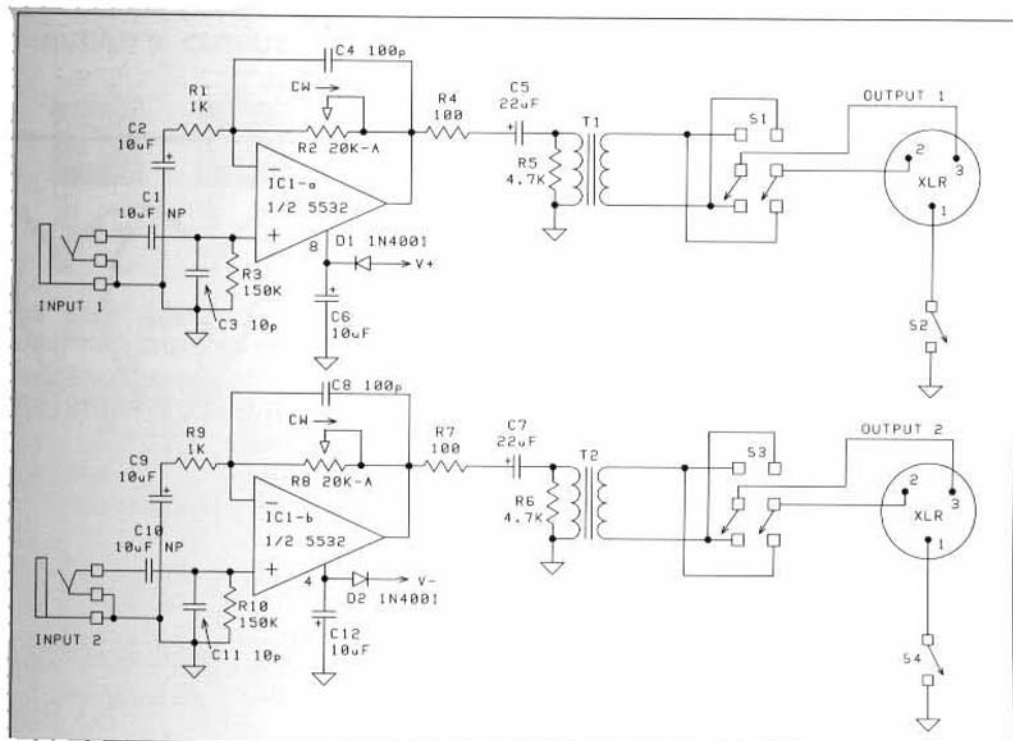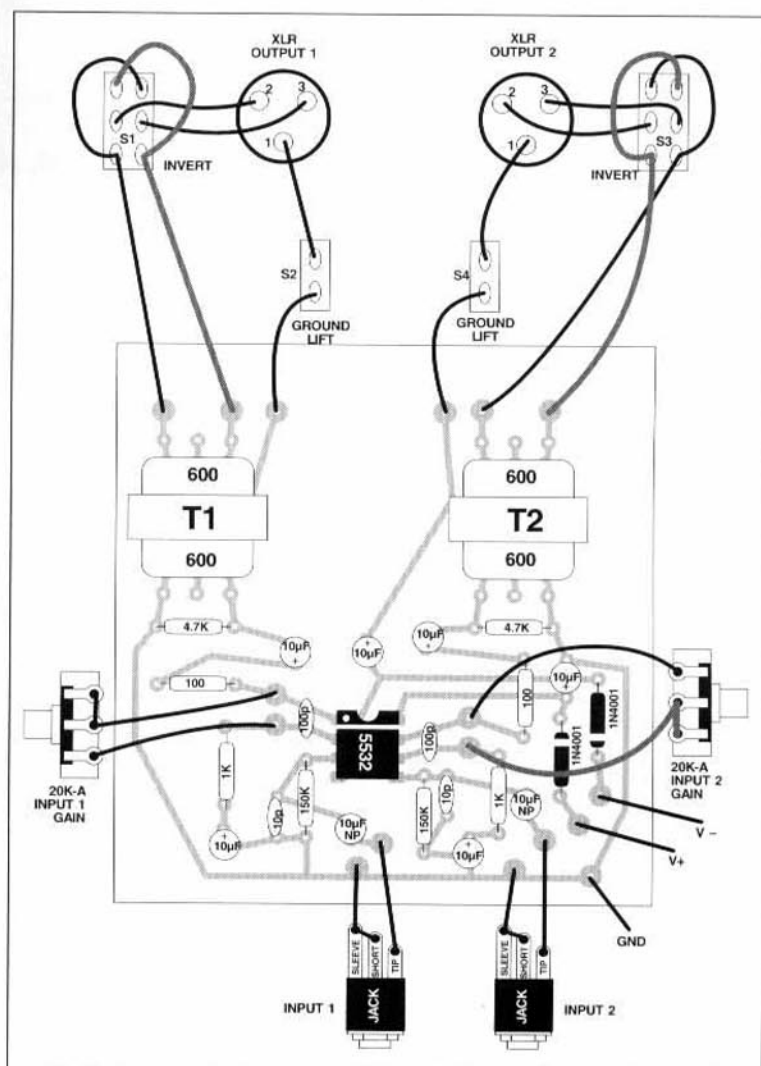
Fig. 14–4. Iso-Matic schematic.

Fig. 14–5. Iso-Matic layout & wiring diagram.

Project No. 15
Split-O-Matic

Parallel processing demands parallel feeds. Split-O-Matic generates four identical feeds off a variable-gain preamp.

Circuit Description

Instrument feed couples through C2 to preamp IC1-a, whose gain varies 1–11 by pot R2. IC1-a output couples to R4-C6 to one output. IC1-a output also couples directly to three DC voltage followers, IC1-b/-c/-d. The nature of the op-amp noninverting input stage means virtually no loading on the output of IC1-a. The outputs of the remaining buffers couple through RC networks to their respective output ports. R5/7/9/11 act as bleeder resistors.

Use

The only control is R2, the preamp gain control. Function of the box is self-evident. The quad op amp specified draws <2 ma, and will run off ±1.5 to ±22V, but suffers relatively high input noise. Get quieter results with a TL074 or an LM837, at the cost of greater current drain and higher minimum supply voltage.
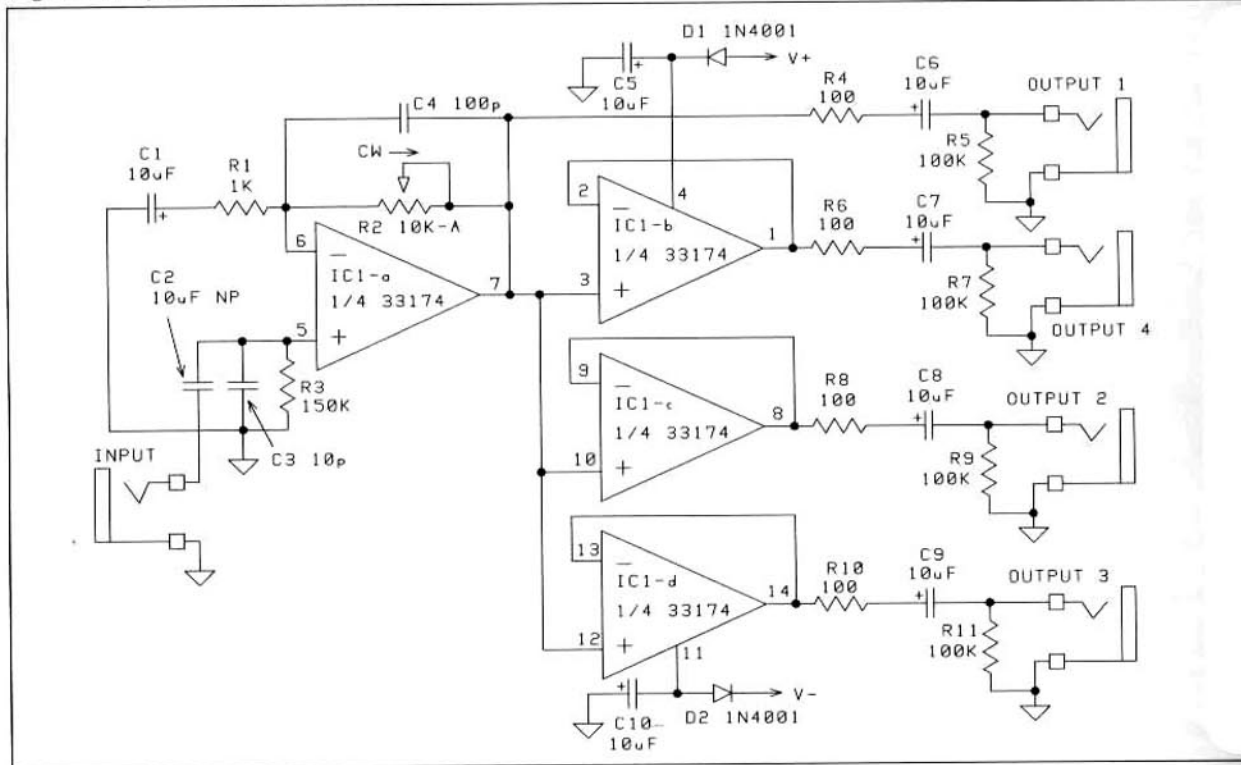
Fig. 15–1. Split-O-Matic schematic.

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OUTPUT 4   OUTPUT 3   OUTPUT 2   OUTPUT 1

JACK   JACK   JACK   JACK

SLEEVE TIP   SLEEVE TIP   SLEEVE TIP   SLEEVE TIP

100K   100K   100K   100K

10µF +   10µF +   10µF +   10µF +

100   100   100

10µF +

100

33174

100

1N4001

10µF +

V −

V +

1N4001

100p

1K

10µF NP

150K   10p

+ 10µF

+ 10µF

GND

JACK

SLEEVE

TIP

INPUT

10K-A
PREAMP GAIN

2" × 2.25" reference box

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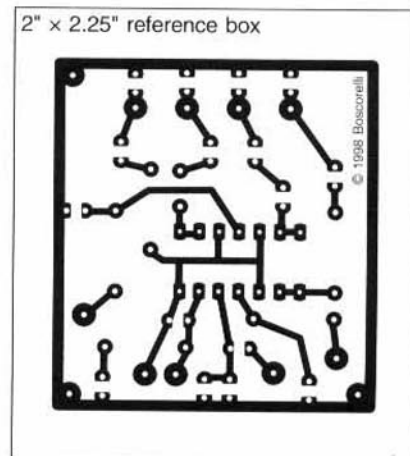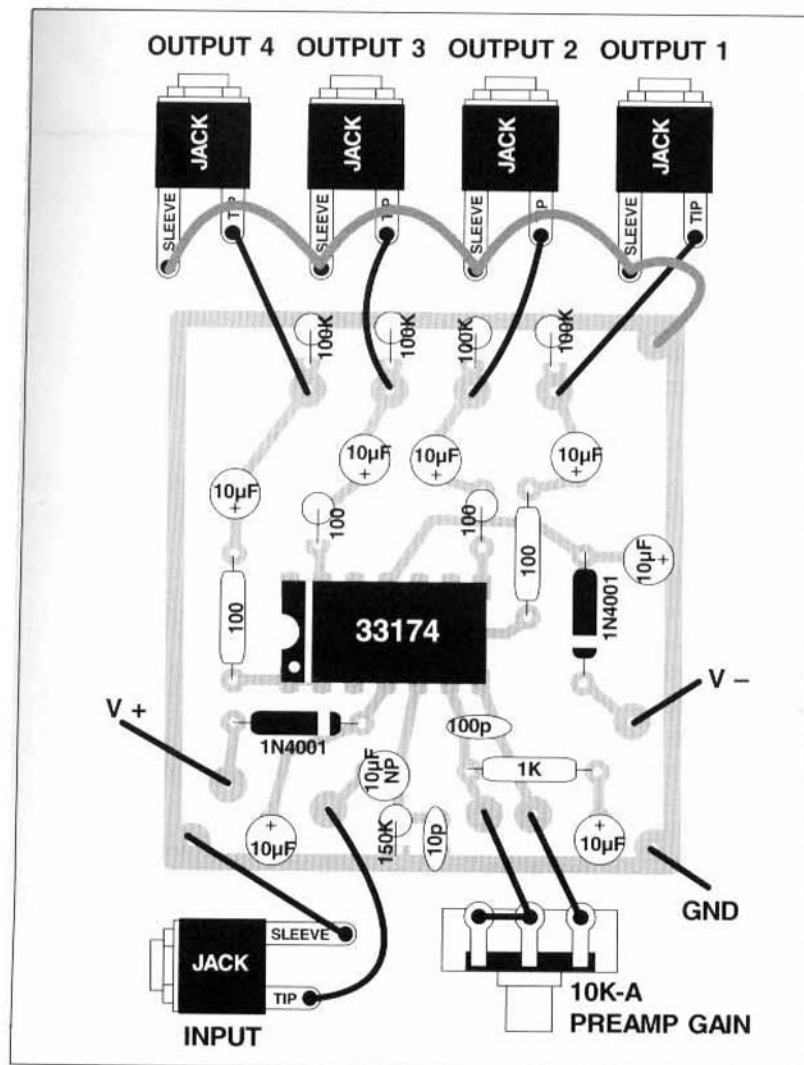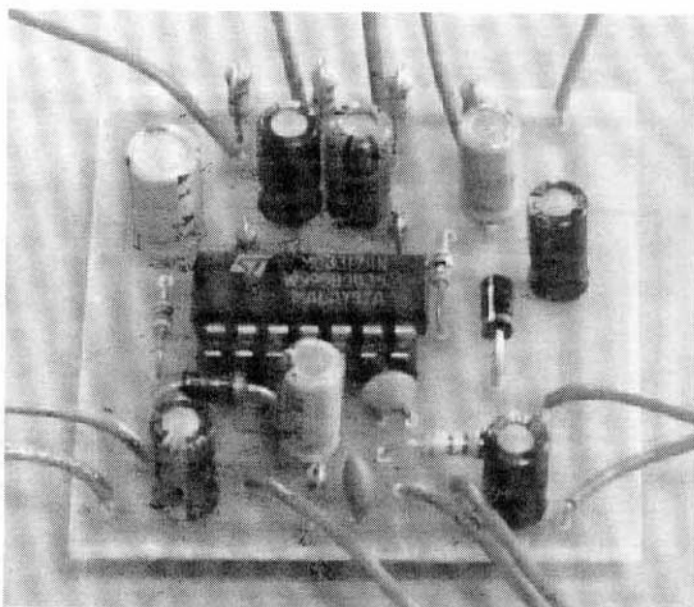
Fig. 15−2. Split-O-Matic
circuit board.

Fig. 15−3. Split-O-Matic
layout & wiring diagram.

Fig. 15−4. Split-O-Matic
prototype board.

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Project No. 16
Mix-O-Matic

Mix-O-Matic is a four-channel, general-purpose, line-level mixer, perfect for recombining parallel-processed feeds. The unit provides an inverting option for each channel, and a master level control.

Circuit Description

The four buffer channels are identical; only channel 1 is described.

Instrument feed couples through C1 to unity-gain AC voltage follower IC1-a, whose output couples to level-control pot R2, whose wiper ties to a polarity-inversion block consisting of S1, IC1-c, and associated components. IC1-c output couples through R5 to summing amp IC3, whose gain varies from 0–2 by action of R6. Audio couples through R7-C11 to the output path.

Use

Switches & pots have these functions:

S1	channel 1 polarity invert
S2	channel 2 polarity invert
S3	channel 3 polarity invert
S4	channel 4 polarity invert
R2	channel 1 level
R6	master output level

R12	channel 2 level
R15	channel 3 level
R22	channel 4 level

Use is self-evident.

MIX-O-MATIC PARTS LIST

Resistors
R1, 13, 14, 23 150K
R2, 12, 15, 22 10K audio-taper pot
R3, 4, 5, 9, 10, 11, 16, 17, 18, 19, 20, 21 22K
R6 100K audio-taper pot
R7 100
R8 100K

Capacitors
C1, 7, 8, 13 10μF nonpolar electrolytic
C2, 3, 4, 5, 6, 9, 10, 12, 14 10pF
C11, 15, 16 10μF aluminum electrolytic

Semiconductors
D1, 2 1N4001
IC1, 2 TL074 quad op amp
IC3 TL071 op amp

Miscellaneous
S1-4 SPDT switch
1/4" shorting jacks, 1/4" non-shorting jack
wire, pots, knobs, circuit board, enclosure, etc.

Fig. 16–1. Mix-O-Matic circuit board.

3.5" × 3" reference box

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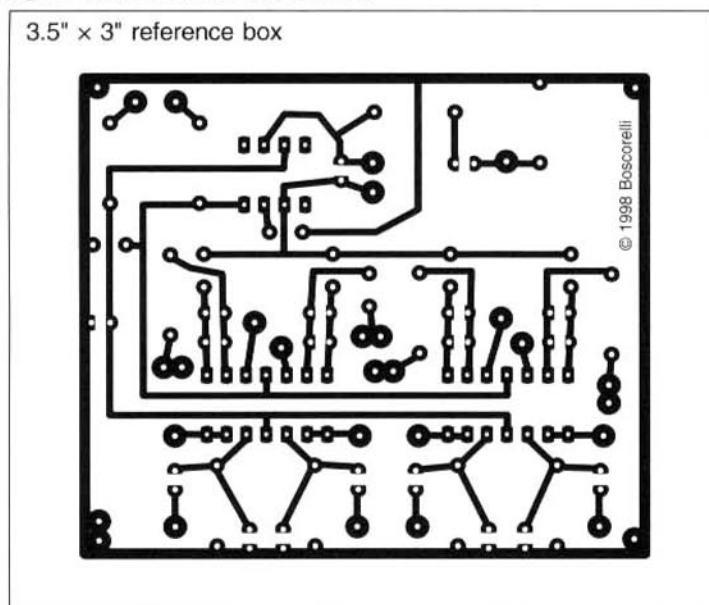
Fig. 16–2. Mix-O-Matic prototype board.
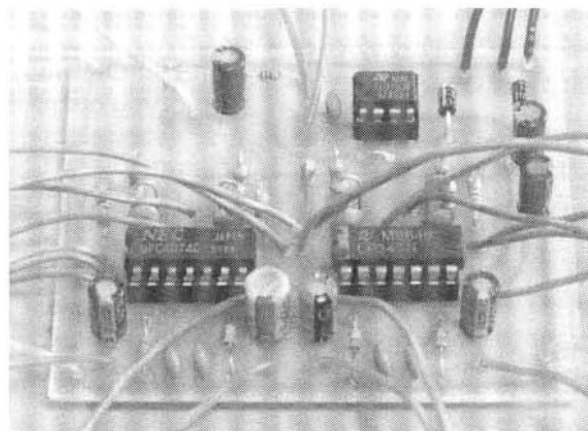
Fig. 16–3. Mix-O-Matic schematic.
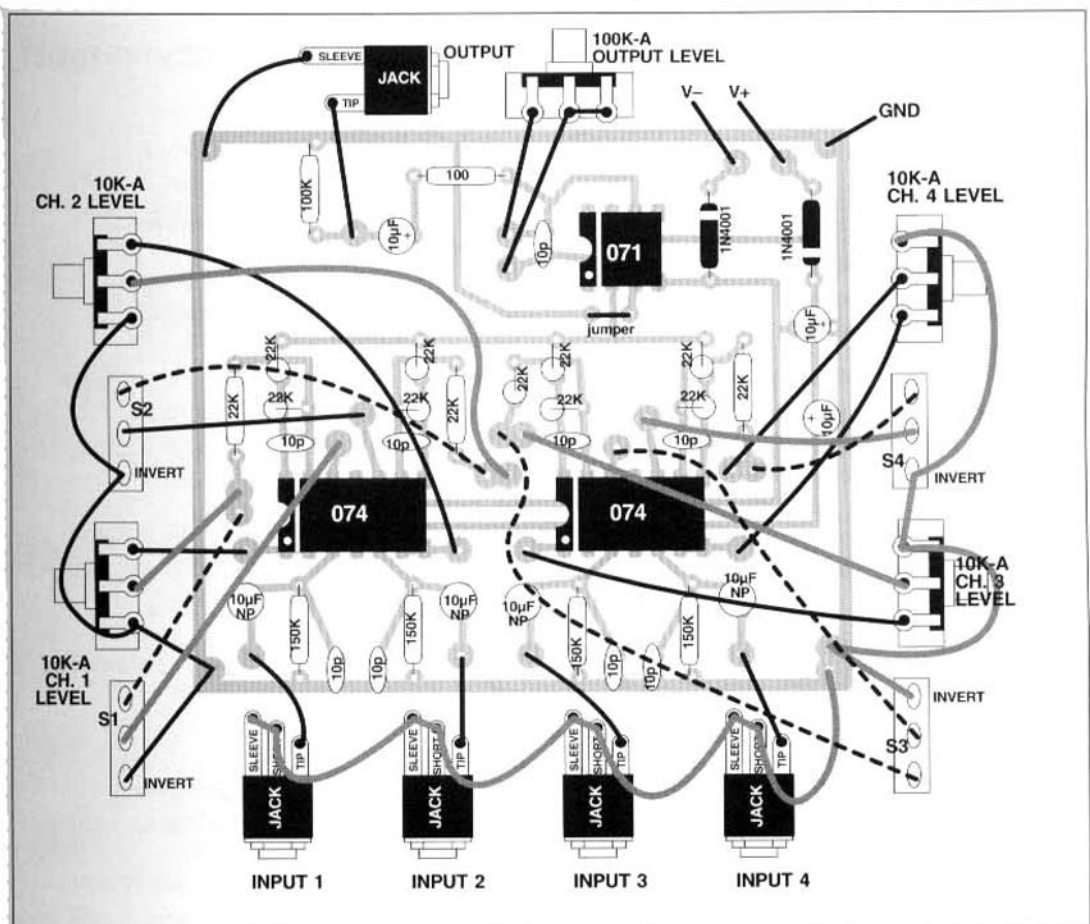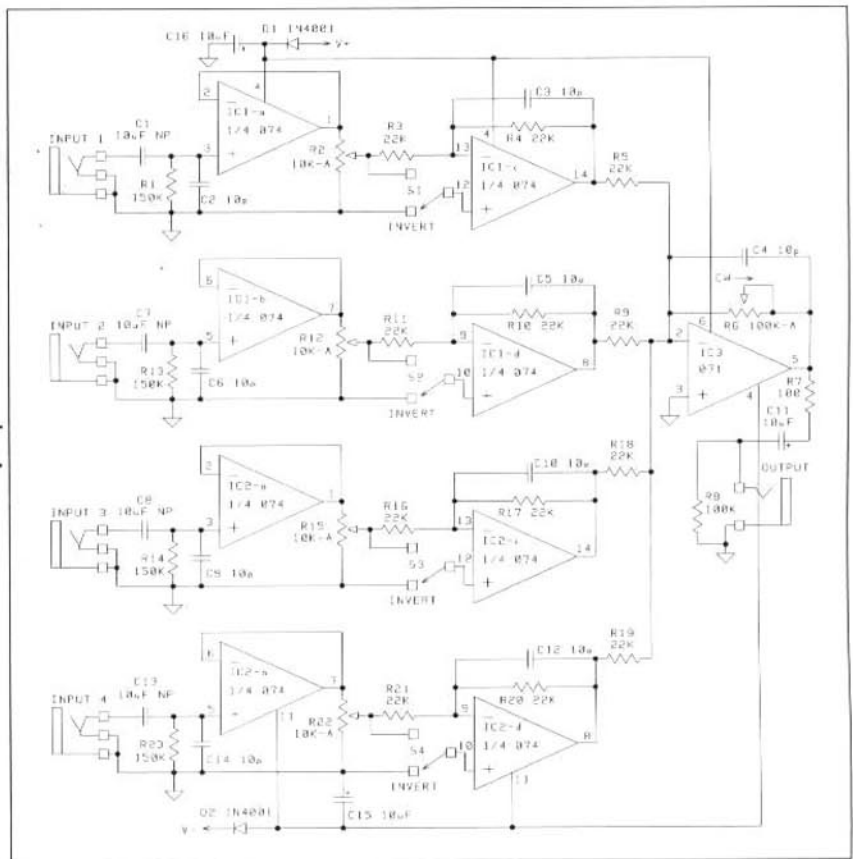
Fig. 16–4. Mix-O-Matic layout & wiring diagram.

Project No. 17
Distort-O-Matic V

DM5 almost got billed 'Swirl-O-Matic.' The box is a distortion device that harnesses the power of a four-quadrant multiplier (4QM) to create harmonic and intermodulation distortion.

Circuit Description

Instrument feed couples through C1 to preamp IC1-a, whose gain varies 1–11 by pot R2. IC1-a output couples through C10 to IC2-a, half an LM13600 configured as a 4QM that multiplies the input signal against itself. R3 trims multiplier balance. IC2-a output couples to a gain-restoration block made up of IC1-b/-c & associated components, which brings level up to match that of the clean feed and blocks a large DC offset. IC1-c output couples to one side of a panning network made up of R10-14; IC1-a output couples to the other side of this network. R12 varies the signal from clean (fully CCW) to distorted (fully CW). Output of the panning network couples to IC1-d, an inverting amp that controls the output level by R15. Signal couples through R16-C7 to the output path.

Use

Pots & switches have these functions:

S1	effect/bypass
R2	preamp gain, 1–11
R3	multiplier balance trim
R12	clean/distorted continuous pan
R15	output level

First, balance the multiplier. Set preamp gain at minimum, output level straight up, R3 in center position, R12 fully CW to give a 100%-distorted output. Feed the input a 1-KHz sinewave @ $3V_{p-p}$. Connect output to oscilloscope. Trim R3 to give a clean, even sinewave at twice the input frequency (see Fig. 17–5D).

If no signal generator and oscilloscope are avail-

able, balance the multiplier by ear. This procedure takes two people. First, connect unit to axe and amp. Center R3. Set preamp gain at maximum, R12 fully CCW. Set the output level at desired volume. Now rotate R12 fully CW. While a helper cleanly picks the low-E–string at the 15th fret, trim R3 for minimum volume. The tone at that point might sound an octave above its expected pitch.

Initial settings: S1 effect in, R2 fully CCW, R3 centered, R12 fully CCW, R15 straight up. In this condition the box acts as a preamp with gain near unity. R2 varies distortion from none to complete. Because the axe output is usually not a pure sinewave, the 4QM output is not frequency-doubled; rather, it contains very high levels of distortion products. The output is doubled in frequency when the input is a pure sinewave, which sometimes happens with cleanly picked notes above the 7th fret. Ring-modulator–like sounds occur when the player picks two notes together.

Fig. 17–1. Distort-O-Matic V prototype board.

DISTORT-O-MATIC V PARTS LIST
Resistors
R1 1K
R2 10K audio-taper pot
R3 200-ohm trimpot
R4, 5, 8 4.7K
R6 10K
R7 6.8K
R9 15K
R10, 11, 13, 14 22K
R12 10K pot
R15 250K audio-taper pot
R16 100
R17 150K
Capacitors
C1 10μF nonpolar electrolytic
C2, 7, 10, 11, 12 10μF aluminum electrolytic
C3, 6 100pF
C4 0.001μF
C5 2.2μF aluminum electrolytic
C8, 9 10pF
Semiconductors
D1, 2 1N4001
IC1 MC33174 quad op amp
IC2 LM13600 dual transconductance amp
Miscellaneous
DPDT switch
wire, pots, knobs, circuit board, enclosure, etc.
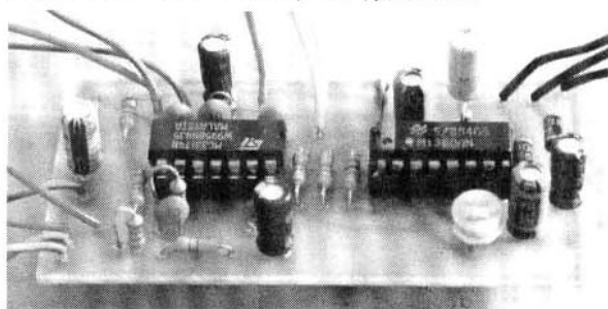
Fig. 17–2. Distort-O-Matic V schematic.
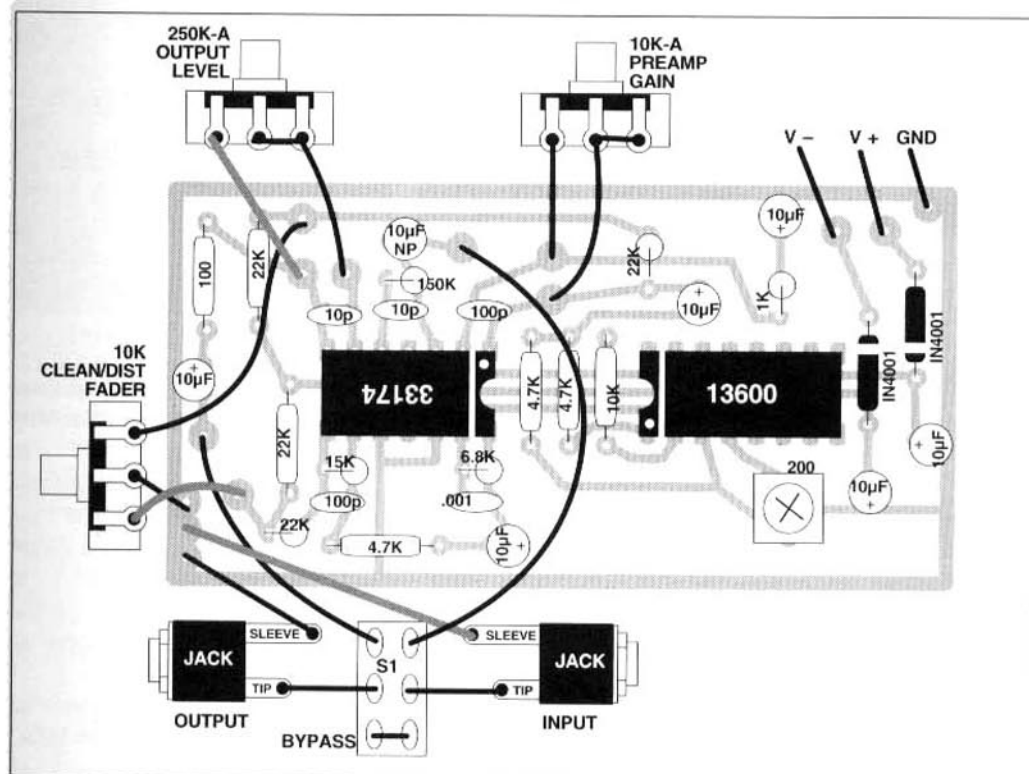
Fig. 17–3. Distort-O-Matic V layout & wiring diagram.

Notes

Making R3 a 1K panel pot lets the player vary multiplier balance. This gives a different range of sounds compared to mixing wet with dry. With R3 at one extreme, the signal coming off the 4QM is almost clean, taking on a bit of tube-like distortion as R3 is rotated away from the limit. If R3 is taken the other way, past the multiplier's balance point, the signal retains a high distortion quotient but distorts at a lower threshold.
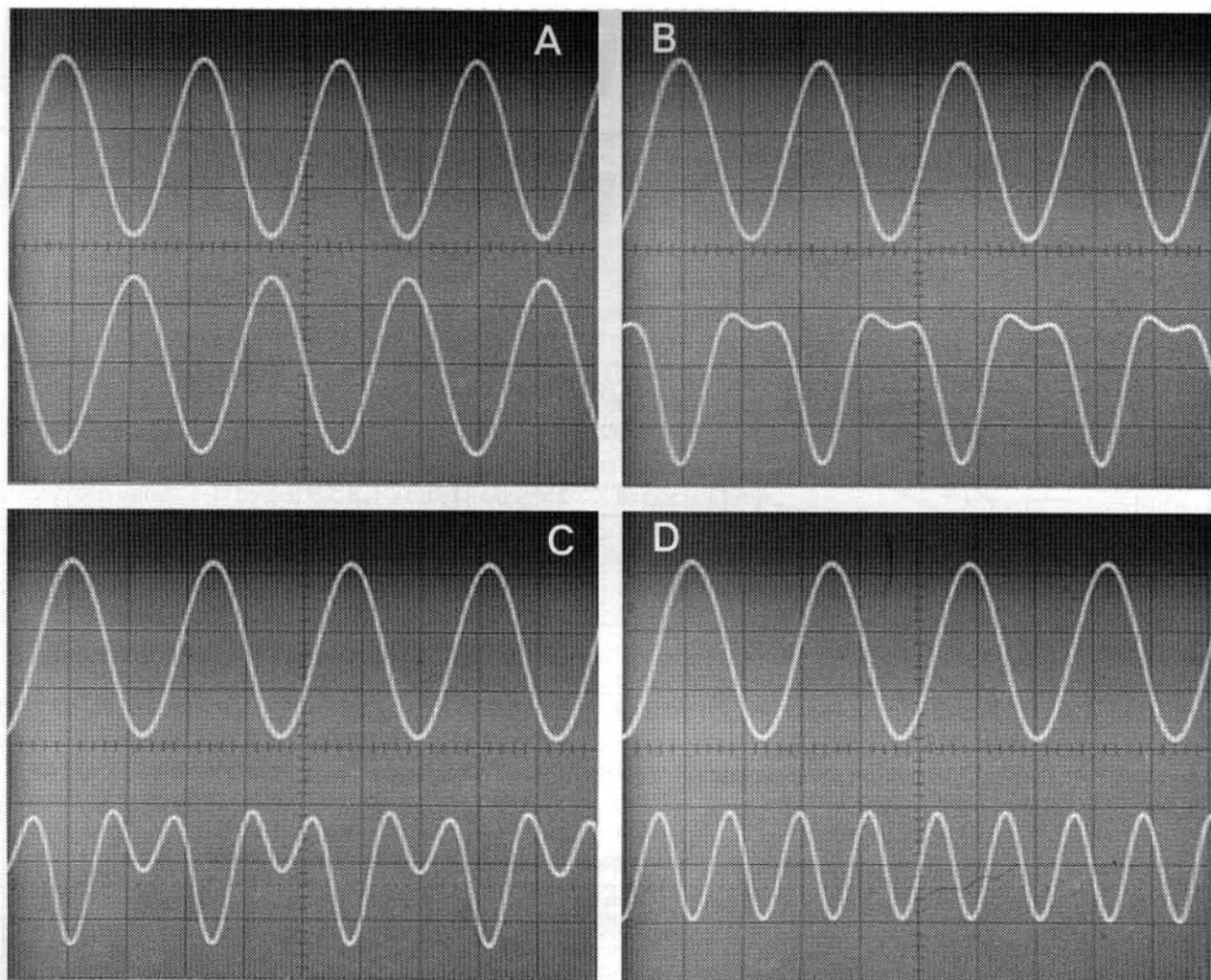
Fig. 17–4. DM5 I/O. Scale 1V/div., 1 KHz sinewave input (top trace), DM5 output (bottom trace), preamp gain @ minimum. A—Output level trimmed to match input; distortion control fully CCW, giving purely clean feed. B & C—Level of second harmonic rises as distortion control is advanced clockwise. D—Panning control fully CW, 100% distortion. Scope display should look this way when R3 is properly trimmed. Complex sinewave sums do not emerge frequency-doubled, but extremely rich in distortion products.
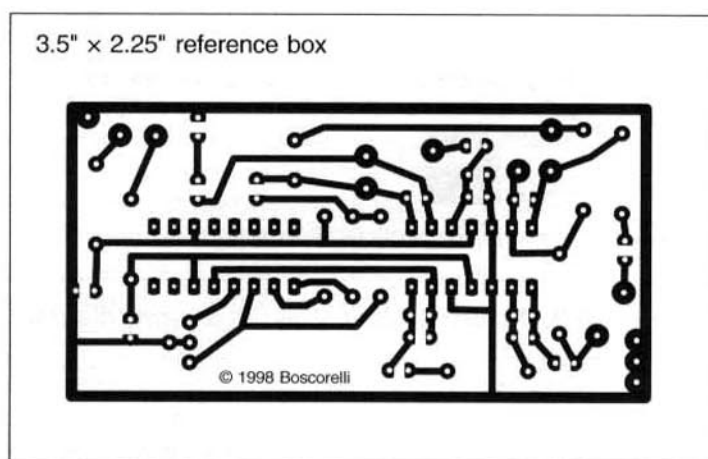
Fig. 17–5. Distort-O-Matic V circuit board.

Project No. 18
Envelo-Matic

Envelope followers use instrument dynamics to tune a voltage controlled filter. Envelo-Matic (EM) fleshes out the concept in a unique and versatile way.

Circuit Function

Signal Path: Instrument feed couples through C9 to noninverting preamp IC1-a. IC1-a output couples directly to a voltage-controlled parametric equalizer comprised of IC5-a, IC2, IC3, and their associated components. IC5-a output couples to highpass filter IC5-c, which attenuates subsonic control feedthrough artifacts; thence to IC5-b and the output path R17-C6. R16 controls output level.

    Control Path: IC1-a output couples to negative full-wave rectifier IC1-b & -c, thence to variable inverting gain block IC1-d, thence through D1 to variable decay network Q1-C3, buffered by IC4-a; thence to variable attack network D4-R9-C4, buffered by IC4-b; thence through D5 to inverting buffer IC4-c. S2 selects straight or inverted voltage feed from this network; applies it to inverting buffer IC4-d, whose DC offset is variable by pot R15. IC4-d output couples through R29 to the control port of the voltage-controlled filter.

    Power Supply: EM incorporates the Append-A-Board Power Supply (Project No. 27), whose details are discussed under that project's heading.

Construction Notes

Keep the 5VDC ground separate from circuit ground. A metal case will require an insulated power jack. The metal tab of the TO-220 regulator case ties to the 5VDC ground, so isolate the heatsink from the metal case. If this is not possible, insulate the heatsink from the tab using a mica washer and silicone heatsink compound.

    The wall wart must supply no less than 8 volts nor more than 12 volts, DC or AC, at a current capacity of at least 500 ma. Observe proper polarity when wiring a DC wart.

Use

Switches & pots have these functions:

S1	effect/bypass
S2	frequency direction up/down
R5	sensitivity
R8	decay (20 ms – 3 sec.)
R9	attack (<5 ms – >500 ms)
R15	static center frequency

ENVELO-MATIC PARTS LIST
Resistors
R1, 2, 3, 7, 10, 11, 12, 22, 23, 31, 32, 33, 35 10K
R4 2.2K
R5, 16 100K audio-taper pot
R6 10M
R8 250K pot
R9 250K audio-taper pot
R13 20K
R14, 29 15K
R15 20K pot
R17, 19, 30, 34 100
R18 36K
R20 22K
R21 68K
R24 10K pot
R25 10K audio-taper pot
R26, 28 1K
R27 150K
R36, 37, 38, 39 4.7K
R40 50K dual pot

Capacitors
C1, 2, 10 100pF
C3 0.0022μF
C4 2.2μF aluminum electrolytic
C5, 12, 15 10pF
C6, 11, 16, 17 10μF aluminum electrolytic
C7, 8 0.1μF
C9 10μF nonpolar electrolytic
C13, 14 0.0033μF, 5% Mylar® or similar
C18 330μF 25V aluminum electrolytic
C19 470μF 10V aluminum electrolytic

Semiconductors
D1, 2, 3, 5 1N914
D4 1N34A
D6 1N4004
IC1, 2, 4, 5 TL074 quad op amp
IC3 LM13600 dual transconductance amp
IC6 LM7805 5V positive regulator (TO-220)
Q1 2N3904 NPN transistor

Miscellaneous
5VDC to ±15VDC converter, 2 watt, Mouser
    p/n 580-NMH0515S or equivalent
S1 DPDT switch
S2 SPDT switch
heatsink for IC6
circuit board, ¼" jacks, knobs, wire, solder, wall wart step-down transformer (see text), etc.

	(100 Hz – 6000 Hz)
R16	output level (0 – ×3)
R24	boost/cut (±15–20 dB; amount depends on bandwidth setting; greater boost/cut is available at wider bandwidth)
R25	preamp gain, 1–11
R40	bandwidth, <¼ octave to >1 octave

EM is a level-driven parametric equalizer covering ~100–6000 Hz. As a tool of subtlety and extremes, harnessing its potential calls for an understanding of the controls. All eight pots are so responsive that it's possible to build a functional box that seems nonfunctional due to improper control settings. Give yourself the better part of a weekend to learn this box. Labeling the controls is a good idea.

Initial settings: S1 effect in, S2 either position, R5, R8, R9 fully CCW; R15, R16, R24, R25, R40 centered. Connect unit to axe and amp, establish desired volume.

First, explore the range of the parametric EQ working under manual control. Turn boost fully CCW, take R15 through its range to vary frequency. You should note strong cut that shifts over the approximate range 100 Hz to 6 KHz. Vary bandwidth to see how this affects adjacent frequencies.

Next, take R24 into boost territory. Exercise care, because EM can boost up to 20 dB.

With R24 applying full boost in a conspicuous way, ease R5 open while plucking a string. The filter's center frequency will move in response to the input signal. How far it moves depends on preamp gain and setting of R5; which way it moves depends on the position of S2.

At this point you may note distortion, due to very fast attack and decay. Advance decay to about 9 o'clock; barely open the attack control, and this distortion should subside. Once you have a handle on how the box works, take attack and decay through their ranges and note the effects of delaying each.

'Wah'-type frequency emphasis occupies a small part of EM's range. The user will find it rewarding to explore EM's effects pre- and post-distortion, to accentuate or mute distortion products under dynamic control.

DC offsets in the filter limit internal headroom to about $22V_{p-p}$. Overload is possible: $3V_{p-p}$ out of the preamp, boosted by 20 dB in the filter, will clip the device.

If you find EM too responsive, change R4 to 10K or 22K, to reduce the gain of IC1-d, which lowers the sensitivity of the level detector.

5.25" × 3.5" reference box

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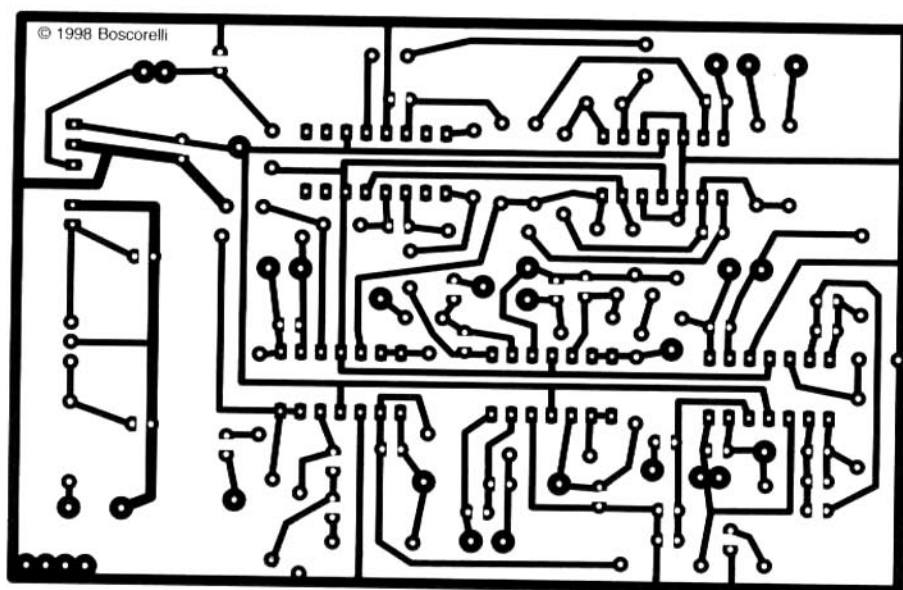
Fig. 18–1. Envelo-Matic Circuit Board.
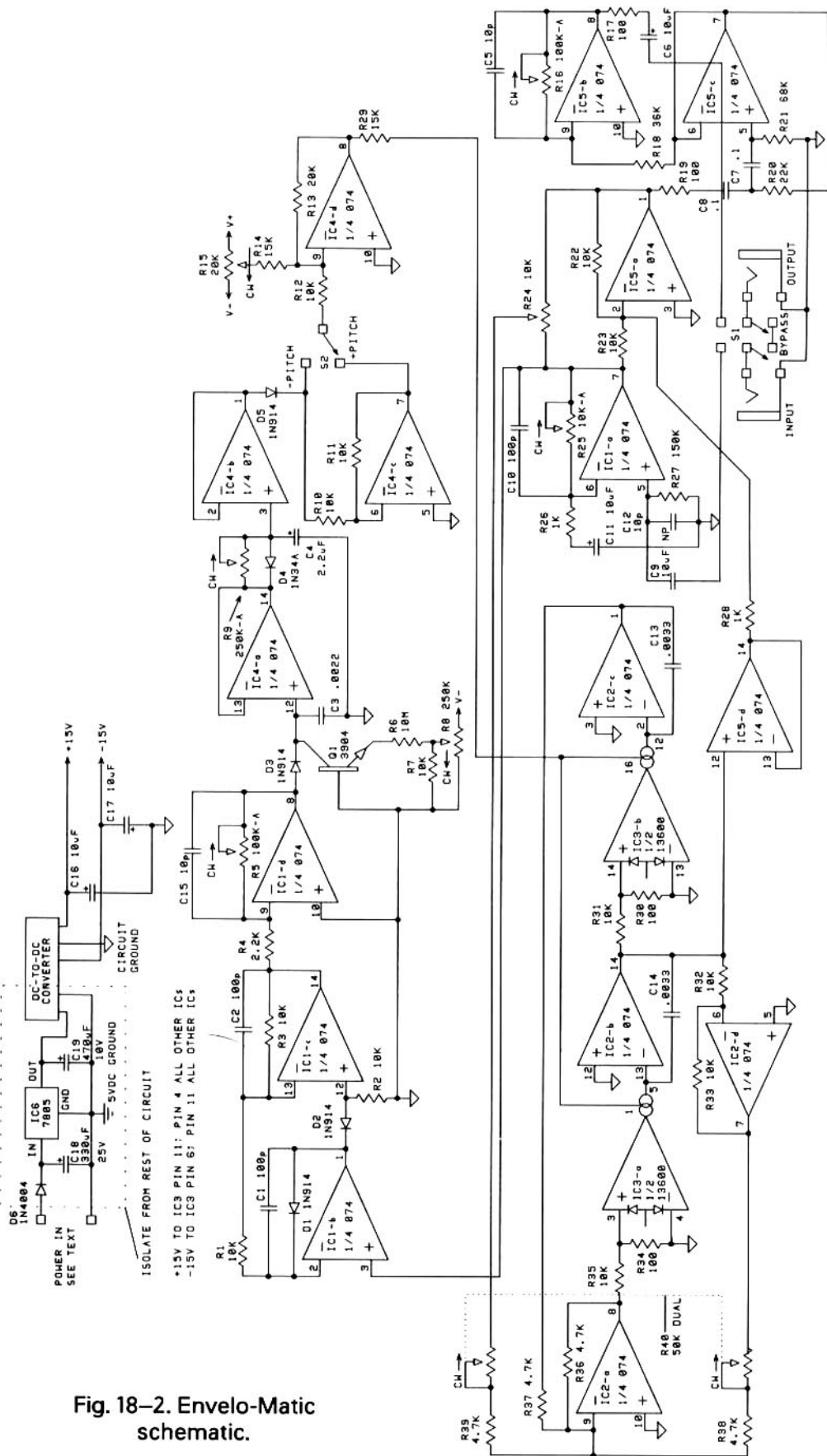
Fig. 18-2. Envelo-Matic schematic.

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Fig. 18–3. Envelo-Matic layout & wiring diagram.

Fig. 18–4. Envelo-Matic prototype board.

Project No. 19
Distort-O-Matic VI

Besides 151-proof distortion, DM6 introduces sounds new to the pedal scene. Not exactly a stomp box, but too cool to omit.

Circuit Function

Transmitter: Audio feed couples through C1 to inverting buffer IC1-b, whose gain varies 0–10 by R2. IC1-b output couples through R3-C4 to the modulation port (pin 5) of IC3, a CMOS 555 timer configured as a frequency modulator whose carrier approximates 40 KHz, and which can be trimmed by R6. C5 helps keep the carrier from feeding back through the FM port. IC3 output drives an ultrasonic transducer.

Receiver: The ultrasonic sensor couples directly to input of IC1-d, a noninverting preamp whose gain varies under control of pot R16. IC2-a output couples directly to a precision fullwave rectifier made up of IC1-a, IC2-b, and their associated components. IC2-b output couples directly to a lowpass filter made up of IC2-c and associated components. IC2-c output—the recovered audio—couples directly to a variable highpass filter made up of IC2-d and associated components; dual pot R23 varies low frequency cutoff over the range ~15–330 Hz. This filter lets the player attenuate low-frequency artifacts as described below. IC2-d output couples directly to inverting amp IC2-a. The audio signal couples through R26-C16 to the output path. R25 varies the output level.

Use

DM6 is an ultrasonic FM transceiver. Encoding, sending, and decoding an axe feed creates two effects. First, overdriving the frequency modulator yields rich distortion whose personality varies with carrier frequency and drive level. In addition, the ultrasonic signal picks up spuriae in transit that decode in ways ranging from mildly interesting to practically cool.

Pots have these functions:

R2	modulator drive level
R6	carrier trim
R16	ultrasonic receiver gain
R23	highpass corner frequency
R25	output level

Initial settings: R2, R6, R16, R25 straight up; R23 fully CCW. Connect unit to axe and amp; set amp volume at minimum. Place the transducers on a table, apertures facing, a few inches apart. Secure them with tape if necessary. Power up DM6, *slowly* turn up the amp's volume; trim R6 until audio is heard. At this point the audio should sound distorted.

Take modulation drive control R2 through its range. The lower the drive level, the less distortion and sustain; the higher the drive level, the greater the distortion and sustain.

Next, explore the effect of altering carrier frequency (R6). The device exhibits a fairly wide sweet spot, as well as marginal areas at the extremes of pot rotation where the distortion threshold rises, giving a punch-through effect.

Test the sound of altering ultrasonic receiver gain

DISTORT-O-MATIC VI PARTS LIST
Resistors
R1, 5, 13, 14, 15, 18, 24 10K
R2 100K pot
R3, 21, 22 4.7K
R4 2.2K
R6 10K pot
R7, 8, 9, 11, 12, 20 22K
R10, 19 39K
R16 10K audio-taper pot
R17, 26, 27 100
R23 100K dual pot
R25 50K audio-taper pot
Capacitors
C1, 7, 16 10μF aluminum electrolytic
C2, 15 470pF
C3 220μF aluminum electrolytic
C4, 13 1μF aluminum electrolytic
C5 0.01μF
C6 0.0015μF, 5% or better
C8, 9 0.001μF, 20% or better
C10, 11 100pF
C12 10pF
C13, 14 0.1μF, 20% or better
Semiconductors
D1, 2 1N914
D3 1N4001
IC1, 2 MC33174 quad low-voltage op amp
IC3 CMOS 555 timer
Miscellaneous
ultrasonic transducers (see text)
shielded cable
enclosure, phono jacks, ¼" jacks, wire, knobs, mounting hardware, 9V battery, etc.
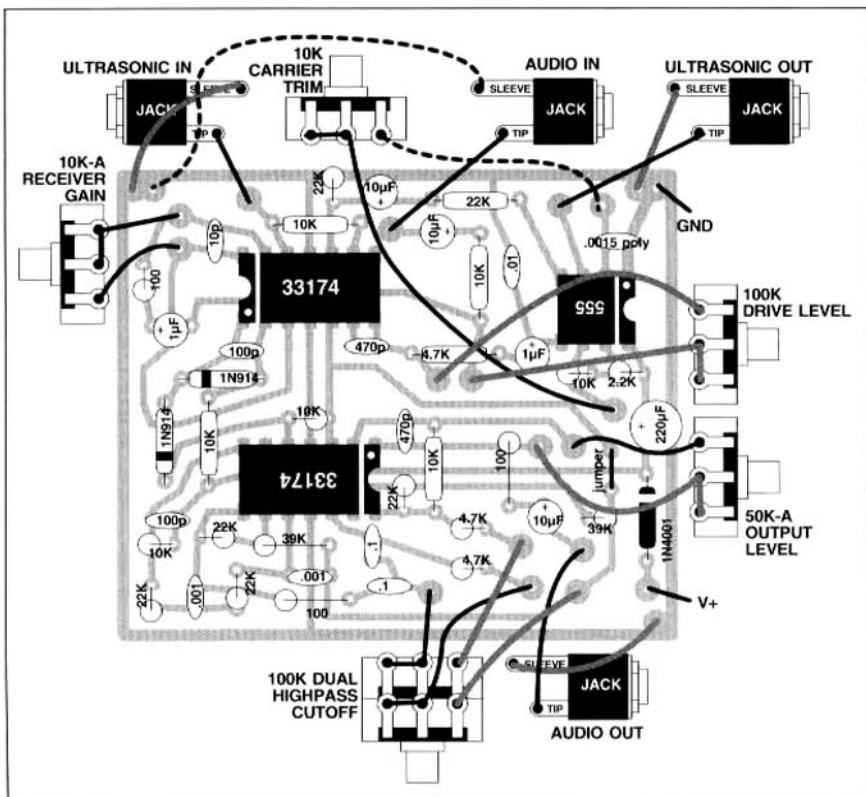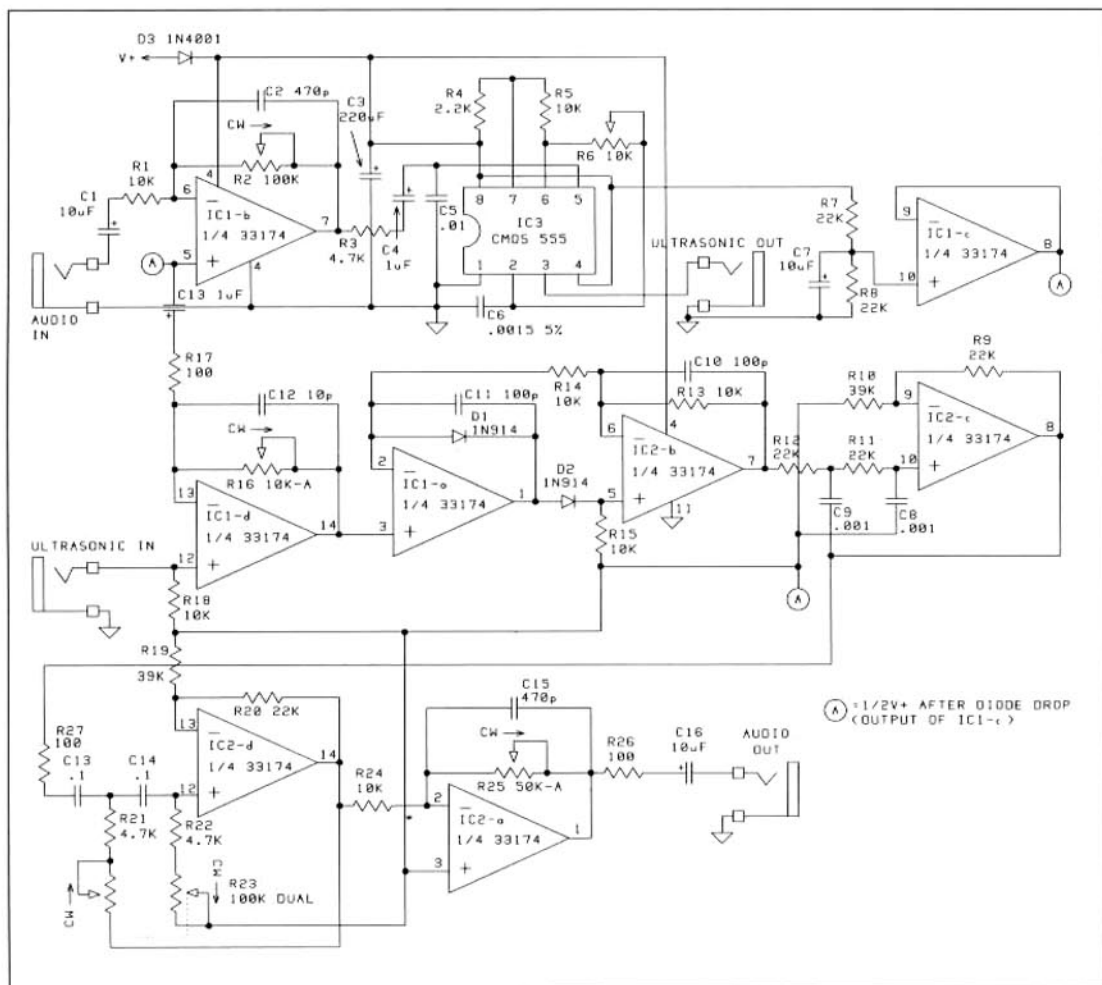
Fig. 19–1. Distort-O-Matic VI layout & wiring diagram.

Fig. 19–2. Distort-O-Matic VI schematic.

by turning R16 fully CCW; this gives minimum receiver gain. Move the transducers to the point they're almost touching; vary receiver gain and note the effect on sound. Now move the transducers about a foot apart. Take R16 through its range and note the effect of putting distance between the transducers.

With transducers about 6" apart and a strong signal established, move your hand over the pair about 2" above them; note the *whoosh* due to spuriae from ultrasonic echoes off the moving hand. Turn R23 fully CW and note the change in tone. It may be necessary to turn up the amp's bass control to perceive the subsonic element of this signal.

The transducers don't have to face each other. Put them about 3" apart but facing the same direction. Now you have a tiny sonar system capable of detecting movement of objects in the beam. Place the pair 3–12" from, say, a speaker cone. Movement of the cone registers in the ultrasonic mix.

Notes

The circuit runs off as little as 5V or as much as 15V, the limit for a CMOS 555. Transducers' leads can solder directly to the board if desired. The transmitter feed should measure 6" or less; the receiver lead can measure up to 3', and should consist of shielded microphone cable or RG174/U coaxial cable.

This system works with 40-KHz resonant transducers, such as Mouser p/n 255-400SR16 (receiver) and p/n 255-400ST16 (transmitter); or Jameco p/n 136653 or p/n 139491 (each Jameco p/n designates a transmitter/receiver pair). Similar transducers sometimes surface in surplus channels; most transducers described as "40-KHz resonant" should work in DM6. Non-resonant transducers, or those that resonate at a frequency other than 40 KHz, will not work. The transducers may register other sources of ultrasonic energy, such as video scan oscillators.

Fig. 19–3. Top: Typical 40-KHz ultrasonic transducers. Lead soldered to case should tie to circuit ground. Bottom: Distort-O-Matic VI prototype board.

Fig. 19–5. Wiring ultrasonic transducer (transmitter or receiver) to coaxial cable & plug.

Fig. 19–4. Distort-O-Matic VI circuit board.

3.25" × 2.75" reference box

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Project No. 20
Modu-Matic

Modu-Matic places the dynamics of one audio feed under control of a second audio feed, or lets an instrument 'modulate' itself.

Circuit Description

Signal Path: Primary instrument feed couples through C14 to preamp IC1, whose output couples through R24-25 to mixer IC6-a; and also to one throw of S1. The external input couples to S1's other throw; S1's pole feeds the signal through C10-R18 to the input of IC4, half an NE570 configured as a voltage controlled amplifier whose output is taken off pin 10 and couples through R17-C7-R19 to IC6-a. Signal couples through R21-C9 to the output path. The net signal path is non-inverting.

Control Path: IC1 output couples directly to a precision fullwave rectifier made up of IC2-a, -b, and associated components. The negative output of IC2-b couples through R4 to inverting amp IC2-c, whose gain varies by R5. The positive output of IC2-c feeds a variable-decay network made up of Q1, D3, C4, and R6-8, buffered by IC2-d. IC2-d output feeds a variable-attack network made up of D4-R9-R29-C5, in turn feeding IC3, whose output is held at the negative limit by bias network R10-11. IC3 output—the VCA control voltage—couples through R13 to IC4's control port.

The result of the control path is for the output of IC3 to approach ground in the absence of a primary input signal. In this state the gain of IC4 is essentially 0, so any signal present in the secondary audio path does not reach the output. A primary input signal generates a positive voltage conducted through IC3 to IC4, causing gain to rise to a maximum of 1 when IC3's output swings to its positive limit. The setting of R5 controls

gain applied to the control voltage and thus varies sensitivity. By this means the dynamics of the primary

Fig. 20–1. Modu-Matic prototype board.

MODU-MATIC PARTS LIST

Resistors
R1, 2, 3, 7, 24 10K
R4 4.7K
R5 100K audio-taper pot
R6 10M
R8 250K pot
R9 500K audio-taper pot
R10 100K
R11, 12, 22, 23 39K
R13 49.9K (±500 ohms)
R14 100K multiturn trimpot
R15 220K
R16 20K single turn trimpot
R17, 25, 26 10K audio-taper pot
R18 36K
R19, 20 22K
R21, 29 100
R27 470
R28 150K

Capacitors
C1, 2, 8, 13 100pF
C3 220pF
C4 0.0033µF
C5, 9, 11, 12, 14, 16 10µF aluminum electrolytic
C6, 15 10pF
C7 1µF nonpolar electrolytic
C10 10µF nonpolar electrolytic
C17 0.1µF
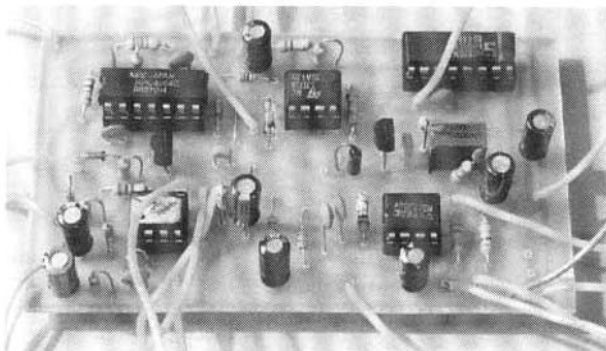
Semiconductors
D1, 2, 3 1N914
D4 1N34A
D5 1N4001
IC1 OP-27 op amp
IC2 TL064 quad low-power op amp
IC3 MC33171 low-power op amp
IC4 NE570 or NE571 dual compander
IC5 78L05 positive 5V regulator
IC6 TL072 dual op amp
Q1 2N3904 NPN transistor

Miscellaneous
S1 SPDT switch
enclosure, ¼" jacks, wire, knobs, mounting hardware, etc.

Fig. 20–2. Modu-Matic schematic.
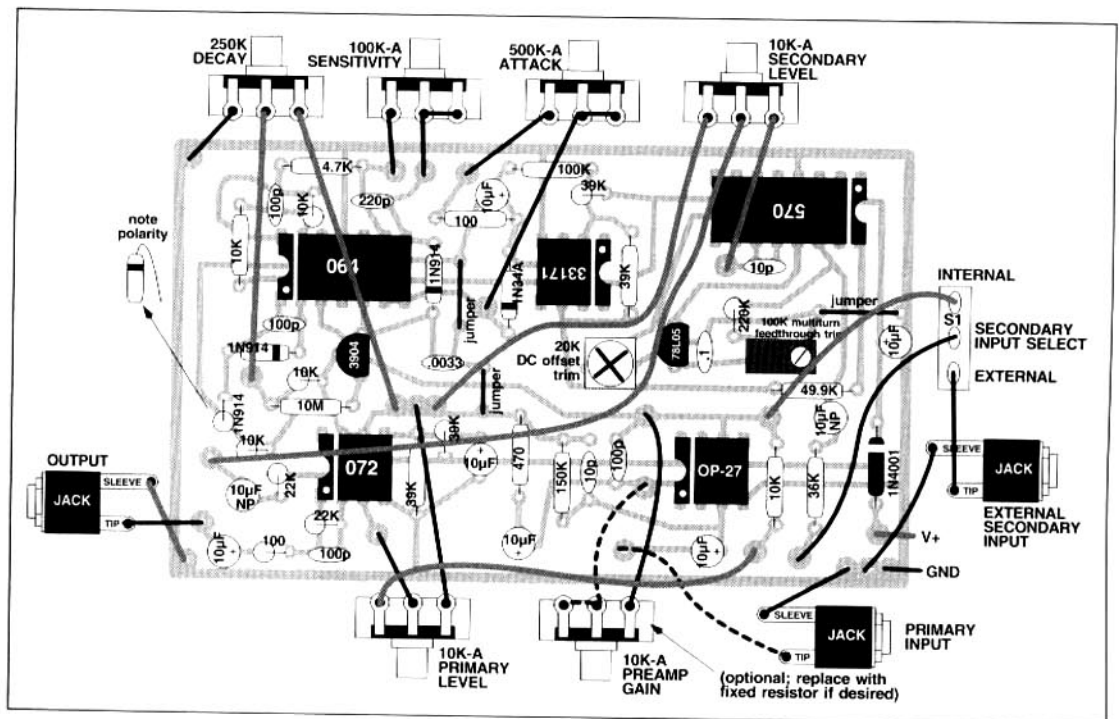
Fig. 20–3. Modu-Matic layout & wiring diagram.

input modulate those of the secondary input.

Use

Switches & pots have these functions:

R5	sensitivity
R8	decay (50 ms – 4 sec.)
R9	attack (1 ms – 4 sec.)
R17	secondary output level
R25	primary output level
R26	primary input gain
S1	secondary input select

Initial settings: all pots fully CCW, S1 either position. First, trim the DC offset of IC4. Apply power, measure voltage at IC4 pin 13; trim R16 to give 1/2V+ at IC4 pin

10. A significant change in supply voltage will require retrimming R16.

[Feedthrough trim is optional. Untrimmed feedthrough (i.e., R14, R15, C17, & IC5 not installed) measured ~30mv$_{p-p}$ in the prototype. Performance was not affected, because the unit usually operates with attack or decay past minimum, which puts feedthrough well into the infrasonic. Users wishing to trim feedthrough can follow this procedure: Set attack and decay at minimum, primary and secondary output levels at zero. Apply a 20-Hz sinewave to the primary input; adjust preamp gain to give 5V$_{p-p}$ at IC1 output. Place scope probe on output of IC2-c, adjust R5 to give maximum amplitude short of clipping. This gives about 3V$_{p-p}$ of ripple in the control voltage. Move scope probe to pin 10 of IC4, trim R14 for minimum feedthrough.]

Connect unit to axe and amp; connect a line-level feed, such as a signal generator, or an FM tuner or CD player, to the secondary input. Trim R26 to give desired preamp gain. Turn R25 fully CCW to kill the primary feed, turn R17 fully CW to open the secondary audio path; slowly open R5 while plucking a string and note that the secondary feed is heard only while a primary input exists. Take R9 through its range and note the effect of delayed attack. Maximum delay is ~4 seconds. Return attack to minimum. Take R8 through its range to hear the effect of delayed decay. At maximum decay, the secondary feed continues to be heard for ~4 seconds after cessation of the primary signal. Turn amp volume down, switch S1 to 'internal' (a pop accompanies switching), return amp volume to desired level. In this state, attack and decay operate on the primary input.

Here's a partial list of what Modu-Matic can do:

► Restore dynamic range to fuzz effects. The nature of clipping strips the signal of dynamic range. By feeding the line-level output of the fuzz box into Modu-Matic's secondary input, fuzz can be made to track instrument dynamics as tightly or as loosely as desired.

► Create special effects. Connect the secondary feed to some odd sound source: a digital sample, FM interstation hiss, etc. The secondary effect takes up to several seconds to emerge after playing commences, or as long to decay once playing ceases.

► Treat the main instrument as the secondary instrument, such that the drum kit modulates the guitar; or use the vocal to modulate the horn feed; etc.

► By using a digital delay or spring reverb as the external input, and with Modu-Matic set for long attack with fast decay, a delay precedes commencement of playing and the onset of reverb, which stops abruptly when the player mutes the strings.

► By switching S1 to 'internal,' and with the primary feed at zero, the instrument modulates itself. Advancing attack creates a delay between picking the string and the emergence of sound from the amp—an automatic guitar swell. Fast decay heightens the effect.

Notes

Modu-Matic is optimized for a 15V supply. Significant departure form this voltage is not recommended. Use of a regulated 15V supply allows omission of D5, and replacement of R16 with a 7.5K fixed resistor; the prototype used both options, as well as a 10K fixed resistor in place of R26.

While the primary input accommodates instrument-level to line-level feeds, the secondary input is designed only for line-level feeds, and high ones at that, preferably averaging at least 2V$_{p-p}$. The higher the level of the raw secondary feed, the better Modu-Matic's S/N ratio.

4.25" × 2.75" reference box

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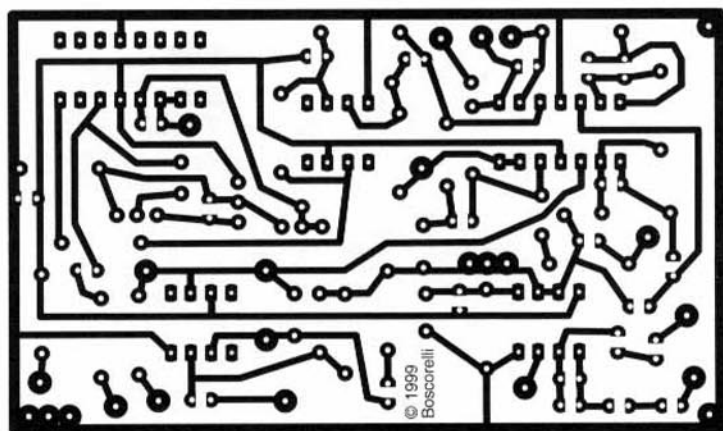
Fig. 20–4. Modu-Matic circuit board.

Project No. 21
Quad Parametro-Matic

QPM is a four-band parametric equalizer.

Circuit Function

Signal feed couples through C3 to noninverting pre-amp IC1-a, whose gain varies 1–11 by pot R15. IC1-a output couples to IC1-b, a unity-gain inverting amp. Boost/cut is enabled by placing a state-variable filter in the input path or feedback loop of IC1-b. The settings of R22-25 determine whether each SVF acts in the input path or the feedback loop. If the SVF exists predominantly in the input path, IC1-b sees a reduced input impedance for frequencies passed by the SVF; the amount of boost depends on the ratio of R20 to the input impedance. If R22-25 are changed such that the SVFs exists predominantly in the feedback loop, gain reduction results from reduced feedback impedance for frequencies passed by the SVFs. All four filters possess independently variable center frequency (R5) and bandwidth (R7). IC1-b output couples through R26 to inverting amp IC1-c, whose gain varies 0–5 under control of R27. IC1-c output couples through R28-C8 to the unbalanced output; the signal path at that point is noninverting. IC1-c output also couples through R29 to unity-gain inverting buffer IC1-d, creating an inverted output that couples through R31-C10 to an optional XLR connector.

Use

Pots & switches have these functions:

S1	equalize/bypass
R15	preamp gain
R22	band-a boost/cut
R5-a	band-a frequency
R7-a	band-a bandwidth
R23	band-b boost/cut
R5-b	band-b frequency
R7-b	band-b bandwidth
R24	band-c boost/cut
R5-c	band-c frequency
R7-c	band-c bandwidth
R25	band-d boost/cut
R5-d	band-d frequency
R7-d	band-d bandwidth
R27	output level

Initial settings: R27 and all boost/cut pots centered, all bandwidth pots fully CCW (minimum bandwidth), all frequency pots fully CCW (lowest frequency). In this state the box acts as a preamp whose gain depends on setting of R15. Connect unit to signal source and target output device; trim R15 to desired level.

Using suitable program material, take each band through the full range of boost/cut, frequency, and bandwidth; verify function of each. To avoid interaction, at least at this stage, return each band's boost/cut control to center before testing the next band.

The prototype ran @ ±15V. Boost/cut was ascertained with a constant $1V_{p-p}$ measured at the preamp output, and with R27 temporarily replaced with a 10K pot, set for maximum output.

The prototype used these component values in each band's state variable filter:

Band-a R3, 4 = 6.8K; R5 = 50K; C1, 2 = 0.1µF
Band-b R3, 4 = 20K; R5 = 100K; C1, 2 = 0.01µF
Band-c R3, 4 = 20K; R5 = 100K; C1, 2 = 0.0022µF
Band-d R3, 4 = 15K; R5 = 100K; C1, 2 = 0.001µF

QUAD PARAMETRO-MATIC PARTS LIST

Resistors
R1, 2, 9, 12 4.7K (×4)
R3, 4, 5 (×4; see text for values)
R6 470 (×4)
R7 100K audio-taper pot (×4)
R10, 11 10K (×4)
R13 150K
R14 1K
R15 10K trimpot
R16, 17, 18, 19 1.8K
R20, 21, 26, 29, 30 10K
R22, 23, 24, 25 10K pot
R27 50K audio-taper pot
R28, 31 100

Capacitors
C1, C2 5% or better (×4; see text for values)
C3, 8, 10 10µF nonpolar electrolytic
C4, 7 10pF
C5, 11, 12 10µF aluminum electrolytic
C6, 9 100pF

Semiconductors
D1, 2 1N4001
IC1, 2 TL074 quad op amp
[semiconductors not individually identified on schematic: 3×TL074]

Miscellaneous
pots, jacks, wire, knobs, circuit board, battery snaps, etc.
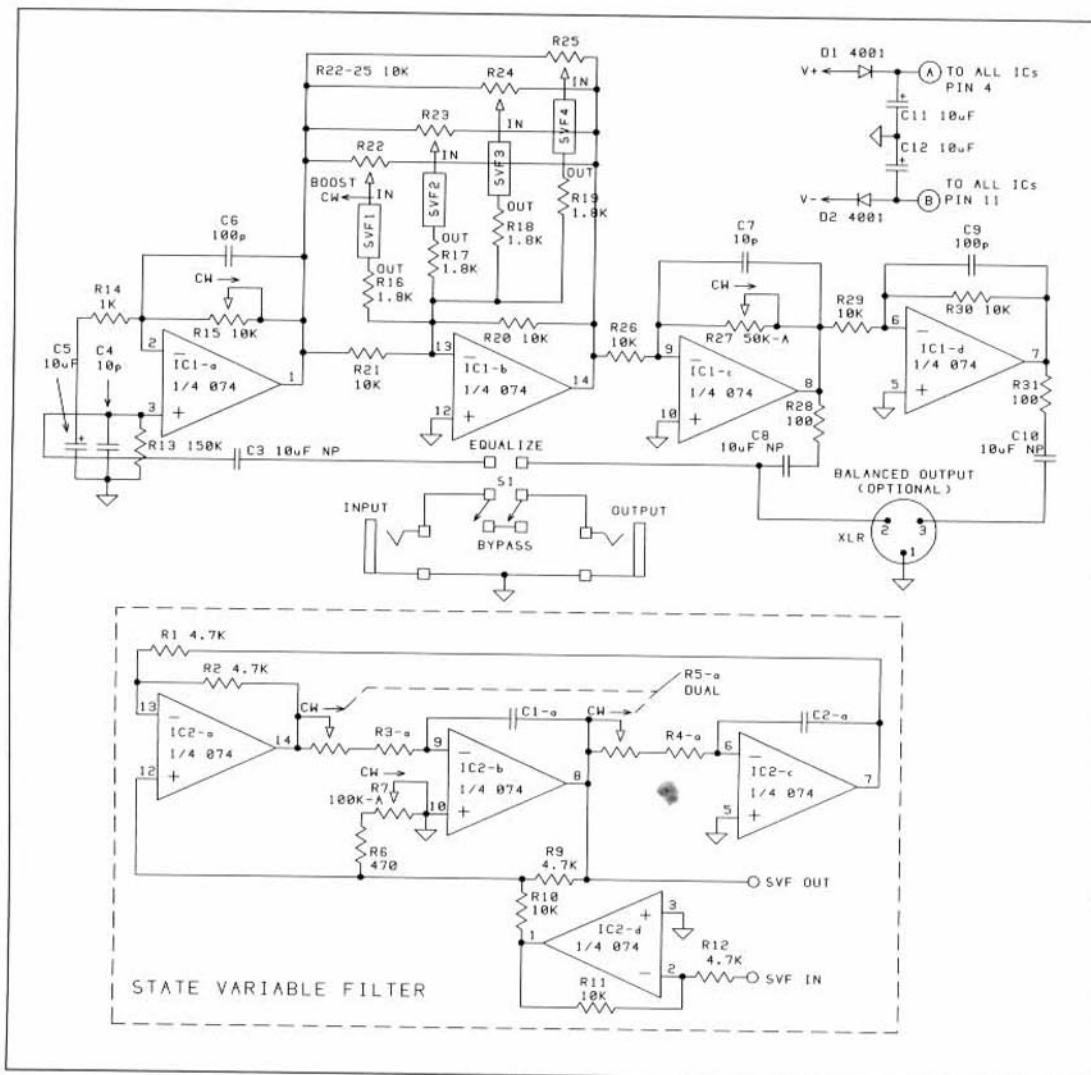
Fig. 21–1. Quad Parametro-Matic schematic.

The prototype's frequency ranges measured as follows:

Band-a	30–240 Hz
Band-b	160–700 Hz
Band-c	600–3000 Hz
Band-d	1500–12,000 Hz

All four bands measured close to ±14 dB of gain with bandwidth at minimum. The maximum boost/cut increased slightly as bandwidth was widened.

Notes

Despite 13 pots, wiring PM2 presents no particular problem if you take your time. Color-coded wire eases the process: one color for all the bandwidth pots, one for all the gain pots, a third for frequency pots. Twisting wire pairs helps, as does labeling each lead with a small piece of masking tape.

While the wiring diagram shows individual leads passing from pads to pot terminals, wiring can be greatly simplified by chaining the ground connection on bandwidth pots R7a/b/c/d, and the connections to IC1-a output and IC1-b output on the end-terminals of boost/cut pots R22-R25.

Instability will result under certain conditions; for example, maximum preamp gain coupled with maximum gain of overlapping bands placed on the same frequency, plus maximum output gain. This totals more than 60 dB. Aside from avoiding such an unmusical setting, separate the input and output leads. Make the pot leads *short*.

The SVFs used in PM2 differ from the one used in Parametro-Matic by sneaking the signal in through the path normally used to control bandwidth, allowing bandwidth control by a single rather than a dual pot (see Ref. 13, p. 191).

Preamp gain resistor R15 can be a trimpot, fixed resistor, or panel pot.

Fig. 21–2. Quad Parametro-Matic layout & wiring diagram.
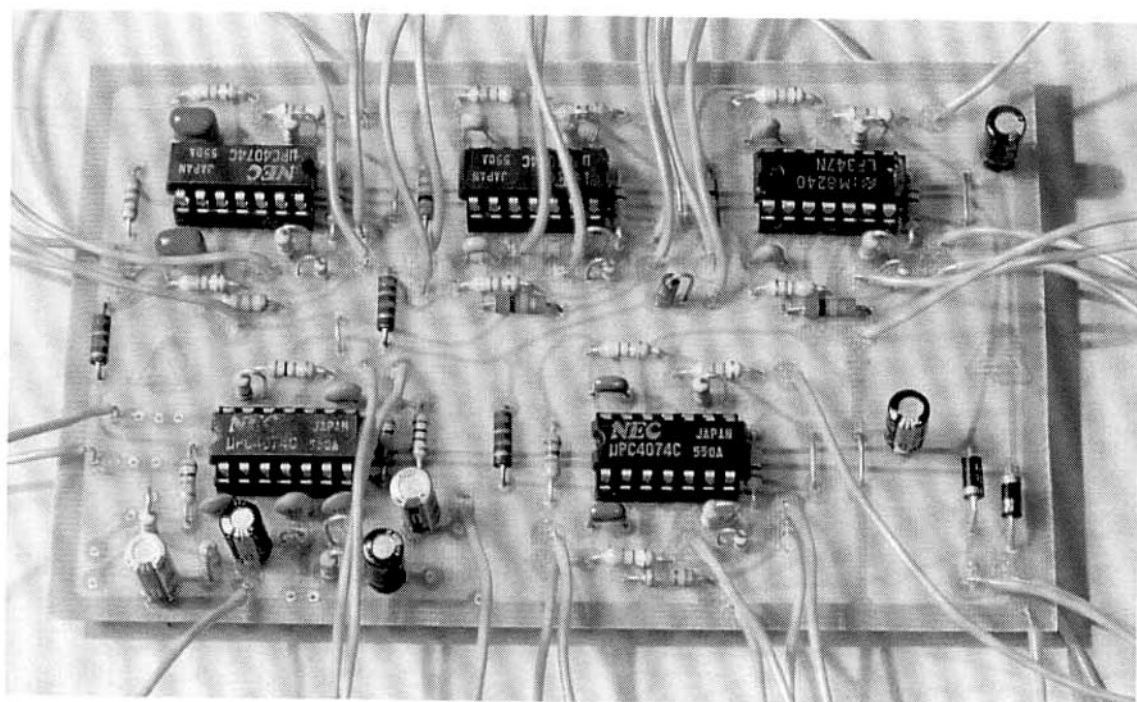
Fig. 21–3. Quad Parametro-Matic prototype board.

5.5" × 3.75" reference box

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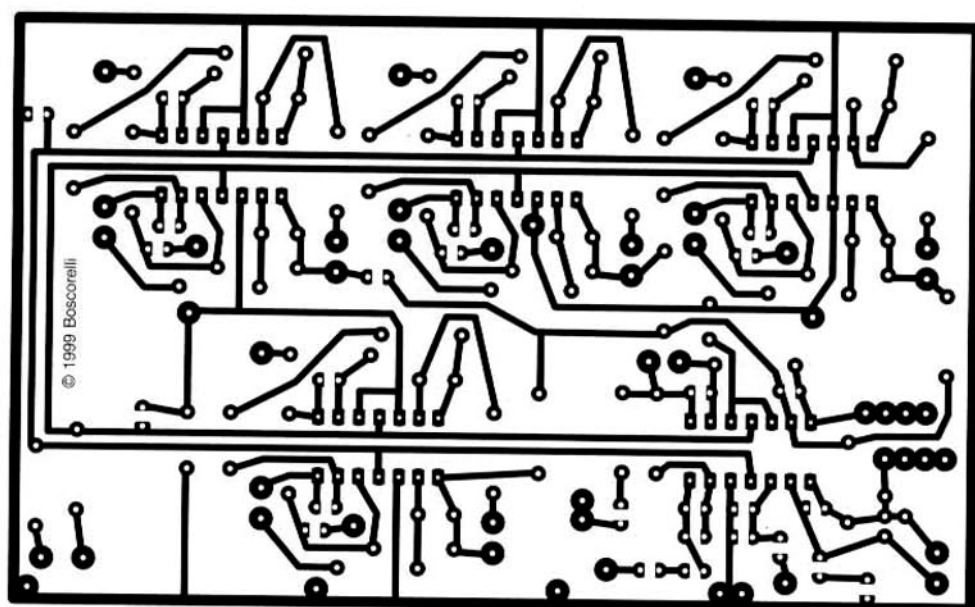
Fig. 21–4. Quad Parametro-Matic circuit board.

Project No. 22
Tremolo-Matic II

Besides low feedthrough and tremendous control range, TM2 contains twin sine generators that produce the sound of "dueling tremolos."

Circuit Description

Signal Path: Audio couples through C17 to IC2, a non-inverting preamp with nominal gain of 22. IC2 output couples through R19-C4 to signal input of IC6, half an NE570 configured as a VCA. IC6 output couples through R21-C6 to inverting amp IC4-b. Audio couples through C8-R23 to the output path; R22 varies the output level. The signal path is noninverting.

    Control Path: IC1-a/-b and associated components form a sinewave oscillator whose rate varies ~1–10Hz by R1, and whose output couples through R9-R10 to inverting summer IC4-a. IC1-c/-d form an identical oscillator whose output couples to IC4-a through R12-R11. R15 varies the DC offset at IC4-a's output, which couples through R16 to the control port of IC6. The result of this control path is for the static DC offset at IC4-a's output to set the resting gain of IC6 at a point between 0 and 1. Sinewaves impressed on the DC offset vary this gain, modulating the instrument volume. Both sinewave paths possess enough gain to clip if their levels are advanced to maximum, giving quasi-squarewave control pulses. IC3-R17-R18 form a feedthrough trim network.

Use

Pots & switch have these functions:

R1	primary rate
R7	primary sine trim
R9	primary depth
R12	secondary depth
R15	static VCA gain

Fig. 22–1. Tremolo-Matic II prototype board.

R18	VCA feedthrough trim
R20	VCA DC offset trim
R22	output level
R28	secondary sine trim
R32	secondary rate
S1	effect/bypass

First, trim the DC offset of the VCA output. Mea-

TREMOLO-MATIC II PARTS LIST

Resistors
R1, 32 2M reverse-audio pot
R2, 31 100K
R3, 4, 8, 29, 30, 33 2.2K
R5, 26 47K
R6, 27 1.8K
R7, 28 1K multiturn trimpot
R9, 12, 22 10K audio-taper pot
R10, 11 33K
R13, 17 220K
R14, 36 150K
R15 20K pot
R16 49.9K
R18 100K multiturn trimpot
R19, 24, 25 36K
R20 20K multiturn trimpot
R21, 35 10K
R23 100
R34 470

Capacitors
C1, 2, 12, 13 1μF 10% tantalum
C3, 7, 15 100pF
C4, 8, 10, 11, 14, 17 10μF aluminum electrolytic
C5, 16 10pF
C6 1μF nonpolar electrolytic
C9 0.01μF
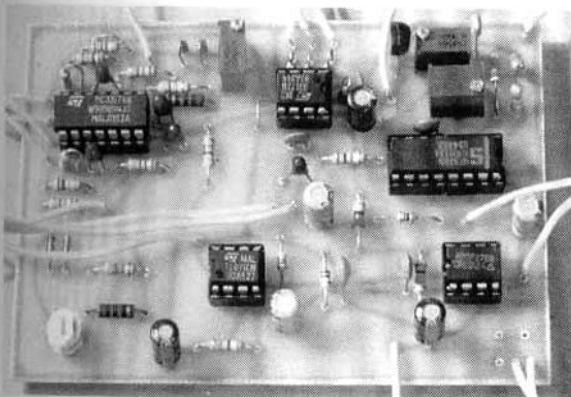
Semiconductors
D1, 2, 3, 4 1N914
D5 1N4001
IC1 MC33174 or TL064 quad low-power op amp
IC2 OP-27 low-noise op amp
IC3 LM78L05 5V positive regulator (TO-92)
IC4 MC33172 dual low-power op amp
IC5 TL071 op amp
IC6 NE570 or NE571 dynamic controller

Miscellaneous
S1 DPDT switch
wire, solder, knobs, 9V battery snaps, enclosure, circuit board, etc.

Fig. 22–2. Tremolo-Matic II schematic.
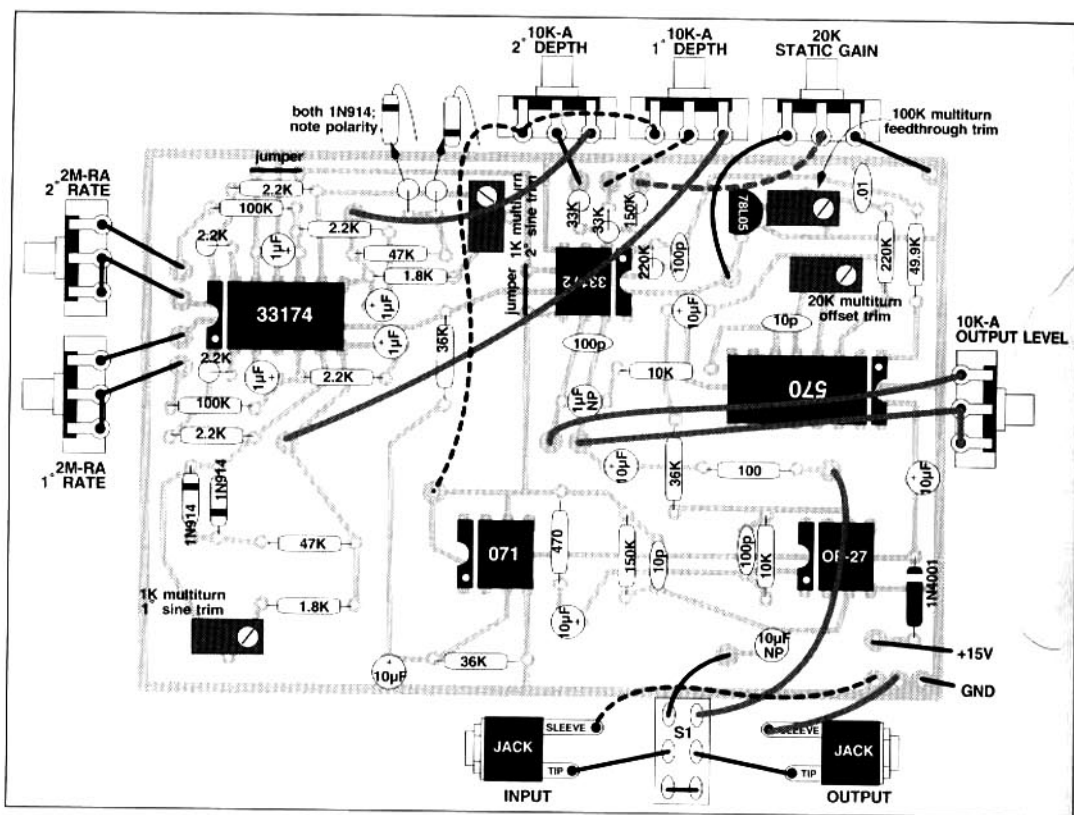
Fig. 22–3. Tremolo-Matic II layout & wiring diagram.

sure supply voltage at IC6 pin 13. Trim R20 to give ½V+ at IC6 pin 7.

Next, trim the primary sine generator. Connect scope probe to IC1, pin 8. Adjust R7 to give a sinewave of ~4V$_{p-p}$. While scope is still connected, take R1 through its range and confirm that rate varies over the approximate range 1–10 Hz.

Move scope probe to IC1 pin 7, trim R28 as above for the secondary sine generator.

Turn R9 & R12 fully CCW; move scope probe to pin 1 of IC4-a; adjust R15 to give ½V+; turn R9 fully CW. This gives a control voltage that clips at both extremes. Move scope probe to IC6, pin 7. Trim R18 for minimum feedthrough.

Connect unit to axe and amp; establish desired volume. Take R9 and R1 through their ranges; vary static VCA gain by R14; note effect on types of tremolo.

Turn R9 fully CCW, advance R12 CW and repeat the checkout sequence for the secondary sine generator.

Once each sine generator checks out independently, test the effect of summing simultaneous control feeds at various rate, depth, and static gain settings.

Notes
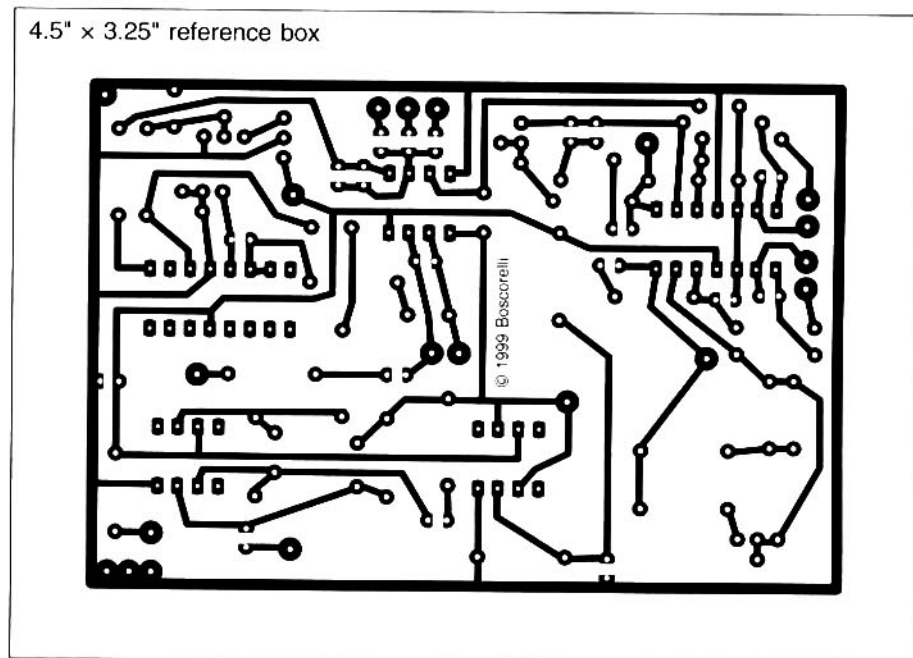
TM2 has been optimized for a 15V supply; significant departure from that voltage is not recommended. The prototype ran off a regulated 15V supply and for that reason omits the polarity protection diode. If you elect to run TM2 off batteries, include the rectifier diode shown on the wiring diagram. The prototype photo shows a single-turn trimpot for R7, but a multiturn pot is recommended.

The preamp is configured for fixed gain suitable for instrument-level feeds. Replace R35 with a 20K audio-taper pot to accommodate inputs ranging from instrument-level to line-level. TM2 should run at a high internal level for best S/N ratio. Average preamp output of 3V$_{p-p}$ leaves adequate headroom.

Adding a second oscillator to a basic tremolo results in more sonic variety than one might expect from the nature of the change. Patient, methodical experimentation rewards the player with sounds not heard on existing recordings. Initially, it's wise not to exceed 50% of either depth setting, because clipping of control feeds tends to mask their interaction. When mixing rates, initially center the static gain control. Once you find an interesting combination, move static gain –15% and +15% off center. In many cases this changes the sound significantly. The most practical way to recall desirable control settings is to provide numbered scales for the controls, and note the settings.

Fig. 22–4. Tremolo-Matic II circuit board.

4.5" × 3.25" reference box

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Project No. 23
Phase-O-Matic

This six-stager's phase-shift method frees depth from dependence on rate, and kills the added noise and feedthrough of the approach through internal companding.

Circuit Function

Signal Path: Instrument feed couples through C9 to noninverting preamp IC7-a. IC7-a output couples through C17 to compressor IC2-a, thence through C18 to a six-stage phase-shift network whose output feeds a highpass filter before being expanded by IC2-b. Expander output couples through C3-R15 to IC8-c, which adds or subtracts clean and phase-shifted versions of the signal, depending on setting of S3. S2 selects expander output, giving vibrato, or output of IC8-c, giving the addition/subtraction characteristic of phase effects. S2's pole feeds output pot R17, whose wiper ties to voltage follower IC7-b. Signal couples through R39-C4 to the output path.

   Control Path: IC6 and associated components form a sinewave oscillator whose rate varies ~0.5–10 Hz by R1. Sinewave output couples to R2, which varies level. R6 varies DC offset at output of IC1, which ties through R7 to the ganged control inputs of six 13600-type OTAs, each configured as a single-ended voltage controlled resistor; each simulated resistor combines with an op amp to give a variable phase-shift network.

   (Note: Do not install R23 & R38. Their need will be determined later.)

Use

Switches & pots have these functions:

R1	rate
R2	depth
R6	static point
R10	sine trim
R17	output level
R19	notch trim
R38	mistracking (optional)
S1	effect/bypass
S2	vibrato/phase

PHASE-O-MATIC PARTS LIST
Resistors
R1 5M reverse-audio pot
R2, 6, 17 10K audio-taper pot
R3, 4, 5 100K
R7 3.3K
R8, 22 150K
R9, 12, 13, 20 2.2K
R10 1K multiturn trimpot
R11 1.5K
R14, 15, 16 10K
R18 4.7K
R19 20K pot
R21 470
R23, 32 470K
R24, 28 30K
R25, 26 47K
R27, 29 62K
R30, 31 36K
R33 39K
R34 33K
R35, 36 330K
R37 47K
R38 50K trimpot
R39 100
[resistors not individually identified on schematic: 12×1K, 24×10K, 6×36K]

Capacitors
C1, 2 1μF 10% tantalum
C3, 5, 18 1μF nonpolar electrolytic
C4, 9, 12, 13, 14, 17, 21, 22 10μF aluminum electrolytic
C6 4.7μF aluminum electrolytic
C7, 15, 23 10pF
C8 470pF
C10, 19 1μF aluminum electrolytic
C11, 20, 29, 30 0.01μF
C16, 24 0.0047μF
C25, 26, 27 220μF aluminum electrolytic
C28 0.1μF
[caps not individually identified on schematic: 6×0.0022μF; see text]

Semiconductors
D1, 2 1N914
D2, 3 3V zener
D5 1N4001
IC1 TL071 op amp
IC2 NE570/571 dual-channel compander
[semiconductors not individually identified on schematic: 3×LM13600, 1×TL074, 1×TL072]

Miscellaneous
S1 DPDT switch
S2, 3 SPDT switch
circuit board, wire, solder, jacks, etc.
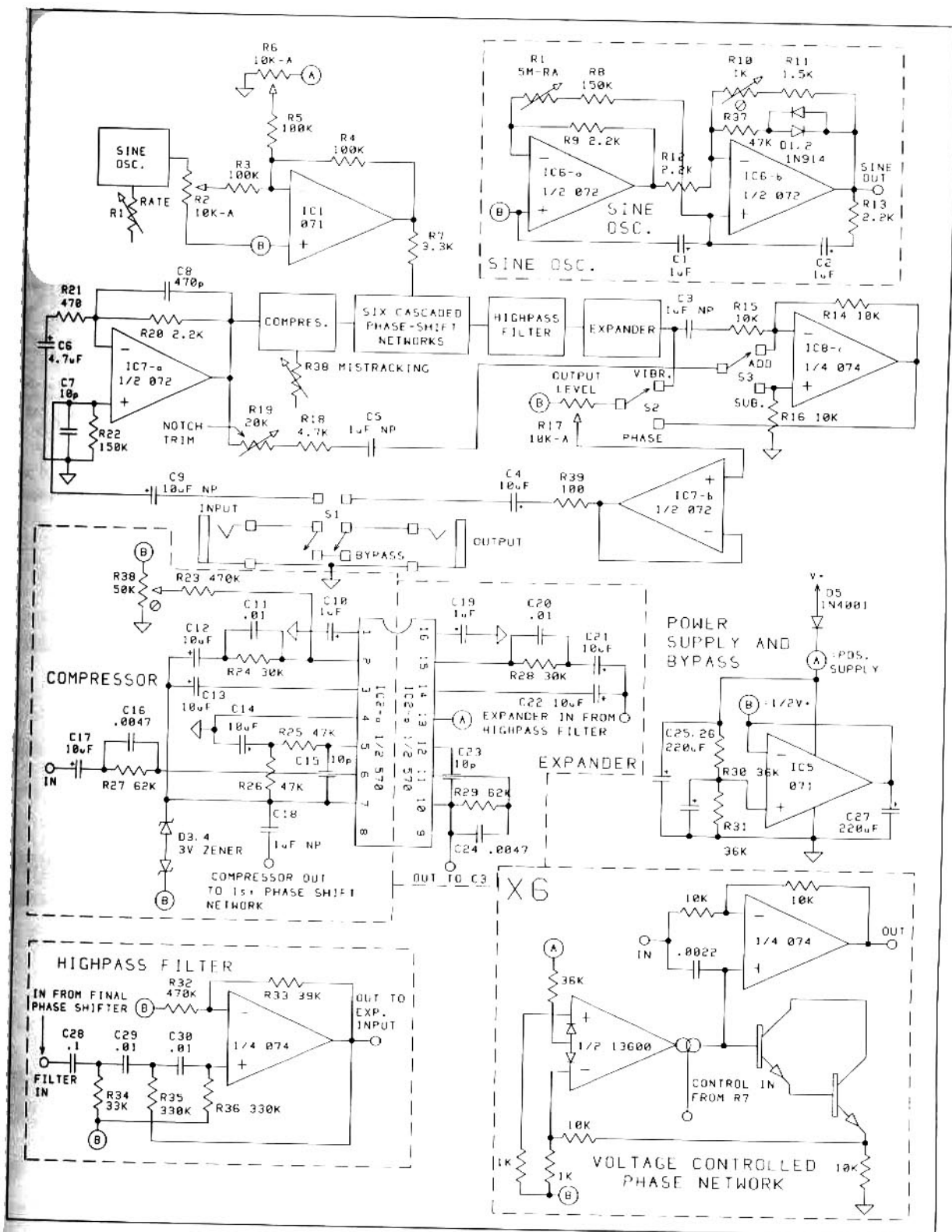
Fig. 23-1. Phase-O-Matic schematic.

S3 add/subtract

First, trim the sine oscillator. Set R1 fully CW, connect scope probe to IC6 pin 7. Trim R10 for sine amplitude ~6Vp-p.

Initial settings: R1 fully CW; R6, R17, R19 centered; R2 fully CCW, S1 effect in, S2 vibrato, S3 either. In this state the box acts as a preamp. Connect unit to axe and amp, adjust R17 to obtain desired volume.

Slowly rotate depth control R2 CW and note the sound of vibrato. Take R1, R2, & R6 through their ranges and note the variety of sounds obtainable by changing the static point and the control voltage amplitude.

6.75" × 3.5" reference box

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Fig. 23–2. Phase-O-Matic circuit board.

Fig. 23–3. Phase-O-Matic layout & wiring diagram.

Fig. 23–4. Phase-O-Matic prototype board; R23 & R38 are not installed.

Turn depth all the way down, switch S2 to 'phase,' S3 to 'add,' set R19 at about 11 o'clock. While lightly picking the open D-string, tune R6 until a null is heard; tune R19 for maximum null. Now turn R2 CW and explore the sound of phase, distinguished from vibrato by addition or subtraction of wet and dry signals.

Notes

POM is optimized for a 15V supply. Significant departure from this voltage is not recommended. The control path makes use of the limited negative output swing of the TL071. Use of an op amp having a greater negative output swing than the TL071's results in feedthrough too severe for the compander to handle.

The preamp's nominal gain of 5.7 suits pickups whose average output does not exceed ~400mv$_{p-p}$. Hotter pickups may require reduction in the value of R20 to avoid clipping, keeping in mind that the addition/subtraction stage adds up to 6 dB of gain for some

frequencies.

POM attains deep modulation at any speed, at the cost of significant feedthrough. Highpass filtering attenuates much of the feedthrough, but quieting the box at rest requires added measures, internal companding in this case. To further reduce noise at rest, POM includes an option to introduce mistracking between compressor and expander (patented technique; see Ref. 45). The prototype was quiet enough at rest that this feature was not needed, so R23 and R38 were not installed.

Most phase boxes feature variable feedback in the phase-shift chain. Companding makes this impractical for POM.

The 0.0022μF caps used in each phase-shift network emphasize a higher band of frequencies than is usual in phase boxes. The caps are easily changed to suit different tastes (compare cap values used in Vibrato-Matic).

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Project No. 24
Compand-O-Matic Lite

Lite is an external device that allows internal companding of many effects.

Circuit Function

Compressor: Audio feed couples through C2 to IC1, a noninverting preamp whose gain varies 1–11, depending on setting of trimpot R2. IC1 output couples through C5 and treble emphasis network R4-C6 to compressor IC4-a's input. Compressor output feeds a second treble emphasis network, R9-C12, feeding the internal level detector, compressing treble more than bass. Compressor output also ties through R10-C13 to the output path. The compression control voltage present at pin 1 ties to voltage follower IC2-a, whose output ties to one throw of S2-b. Zener diodes D2-3 limit maximum compressor output.

Expander: Signal input couples through C25 to IC3-a, a noninverting amp configured for gain of 2. IC3-a output ties to pot R14, feeding buffer IC3-b, whose output couples through C20 to expander signal input pin 14, as well as through C22 to treble emphasis network R13-C21, to one throw of S2-a. When S2 selects 'internal,' the expander's control voltage is a buffered version of the compression control voltage, and the expander's rectifier input is open. When S2 selects 'external,' the rectifier output pin 16 ties to integrator cap C23, and rectifier input pin 15 gets the output of treble emphasis network R13-C21. Expander audio output couples through C17-R11 to the output path.

Use

Switches & pots have these functions:

R2   preamp gain 1–11 (trimpot)

Fig. 24–1. Compand-O-Matic Lite prototype board.

R11   expander output level
R14   expander input level
S1   compand/bypass
S2   expander drive internal/external

Initial settings: S1 compand, S2 internal; R14 centered, R11 fully CW, R2 fully CCW. Feed the compressor output to the expander input.

Connect IC1 output and expander output to oscilloscope. Feed the input a 1-KHz sinewave @ 1V$_{p-p}$. Adjust R14 so that compressor input and expander output levels match.

If no scope is available, connect unit to signal

COMPAND-O-MATIC LITE PARTS LIST

Resistors
R1 1K
R2 10K trimpot
R3, 17 150K
R4, 12 62K
R5, 6 36K
R7, 8 47K
R9, 13 30K
R10 100
R11 10K audio-taper pot
R14 100K pot
R15, 16 4.7K

Capacitors
C1, 5, 7, 8, 10, 11, 13, 15, 17, 20, 22, 25 10μF aluminum electrolytic
C2 10μF nonpolar electrolytic
C3, 26 10pF
C4, 24 220pF
C6, 19 0.0047μF
C9 220μF aluminum electrolytic
C12, 21 0.01μF
C14, 18 10pF
C16, 23 1μF
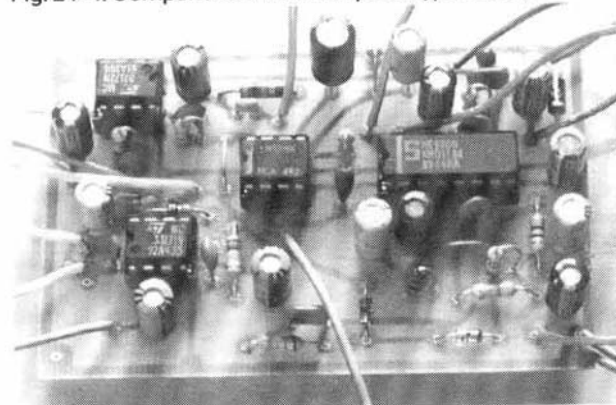
Semiconductors
D1 1N4001
D2, 3 3.3V zener
IC1 MC33171 op amp
IC2 LM358 dual op amp
IC3 MC33172 dual op amp
IC4 NE570/571 dual-channel compander

Miscellaneous
S1, 2 DPDT switch
circuit board, solder, wire, jacks, knobs, enclosure, etc.

Fig. 24—2. Compand-O-Matic Lite schematic.
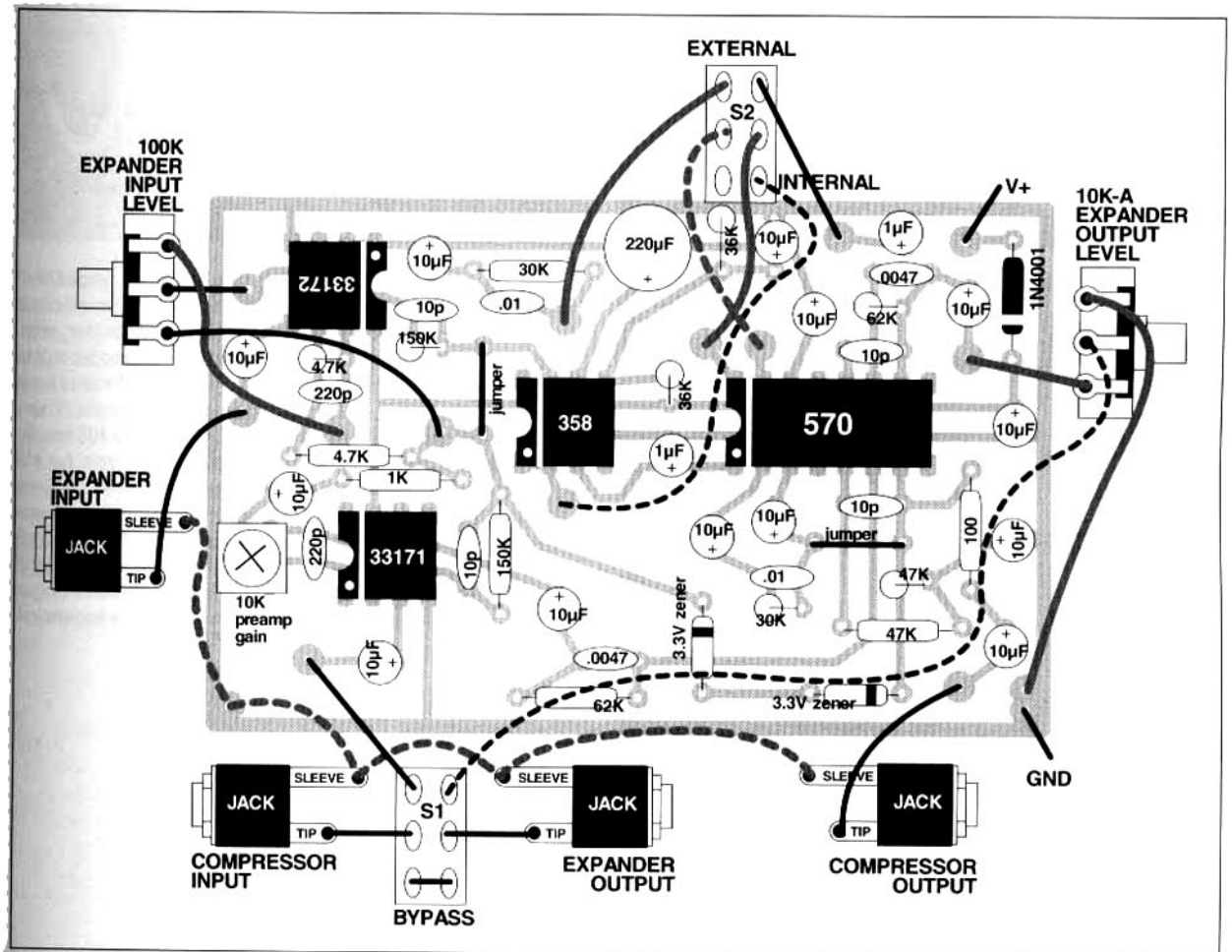
Fig. 24—3. Compand-O-Matic Lite layout & wiring diagram.
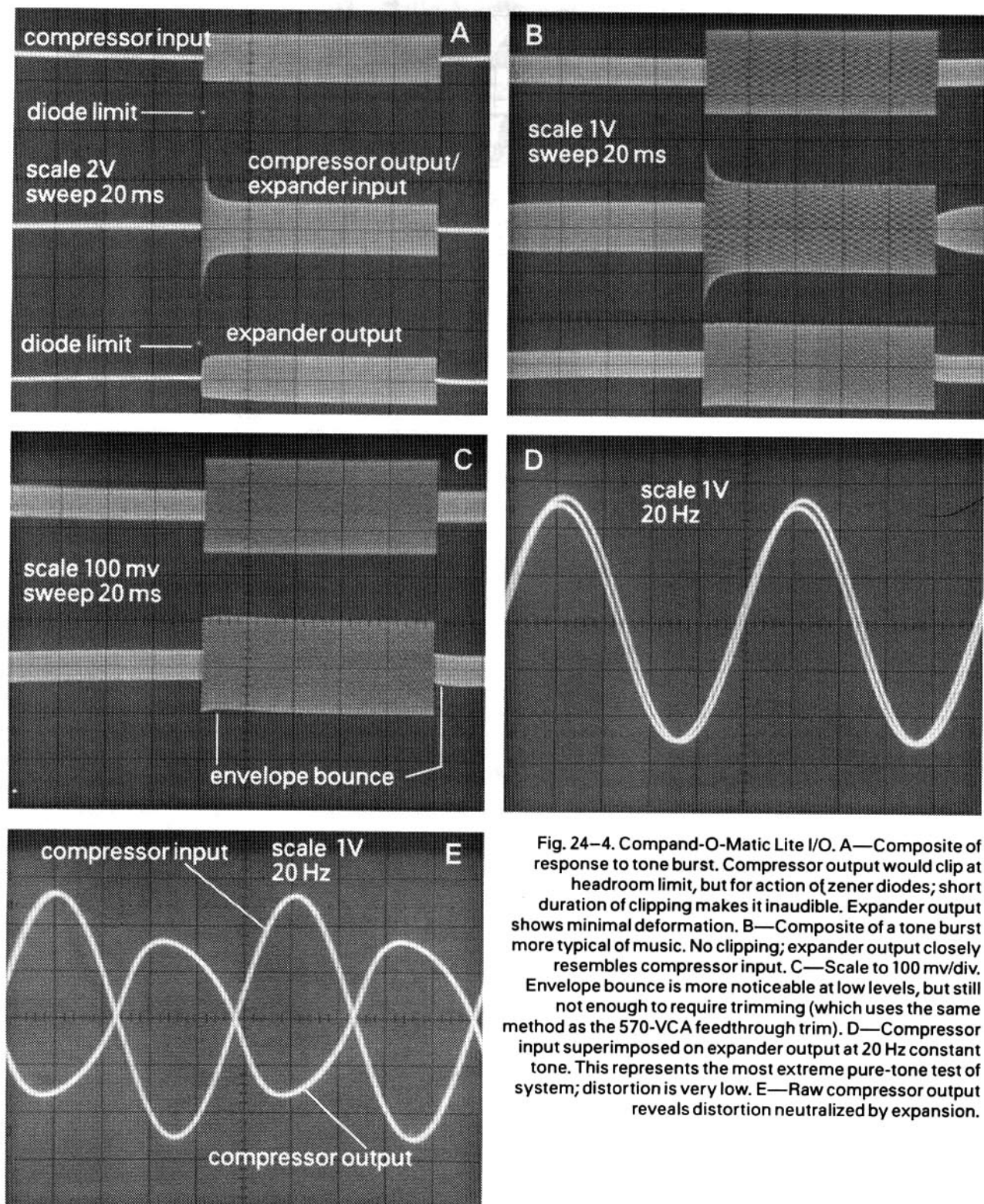
Fig. 24–4. Compand-O-Matic Lite I/O. A—Composite of response to tone burst. Compressor output would clip at headroom limit, but for action of zener diodes; short duration of clipping makes it inaudible. Expander output shows minimal deformation. B—Composite of a tone burst more typical of music. No clipping; expander output closely resembles compressor input. C—Scale to 100 mv/div. Envelope bounce is more noticeable at low levels, but still not enough to require trimming (which uses the same method as the 570-VCA feedthrough trim). D—Compressor input superimposed on expander output at 20 Hz constant tone. This represents the most extreme pure-tone test of system; distortion is very low. E—Raw compressor output reveals distortion neutralized by expansion.

source and line-level amp, such as a home stereo unit. Power up; then, while switching S1 between bypass and compand, trim R14 until no difference in level is heard between bypass and compand.

Turn off amp and Lite. Connect compressor output to input of a noisy stomp box, such as a vintage phase shifter, chorus, or other effect whose gain approximates one. Feed effect output to expander input. Power up amp and outboard gear. Repeat the level-matching sequence, note the reduction in noise at rest.

Notes

Lite is optimized for a 15V supply, but works satisfactorily over the range 12–18V.

Lite's companding system is a chopped version of Signetics' Databook circuit. Noise reduction compares to that supplied by dbx Type II.

Companding works best with unity-gain effects. Effects with gain other than one fare best with S2 in 'external' position, the expander running off its own level detector. Match levels either by measurement, or by trimming R14 while switching between compand & bypass.

Lite can effect record/playback companding with the expander in 'external' mode. A $1.0V_{p-p}$ 1 KHz test tone at the outset of the recording facilitates later level match.

2.5" × 3.75" reference box

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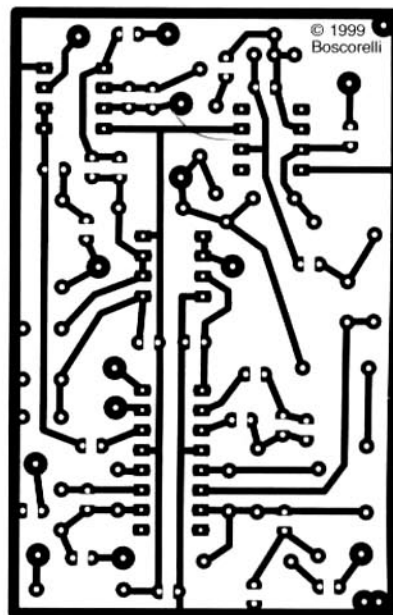
Fig. 24–5. Compand-O-Matic Lite circuit board.

Project No. 25
Tone-O-Matic II

Five-band graphic EQ realized through simulated inductance.

Circuit Function

Signal input couples through C8 to noninverting preamp IC1-a, whose gain varies 1–22 depending on setting of R2. Preamp output couples to IC1-b, a noninverting amp whose gain depends on settings of gain control pots R7-R11. When pots are fully CW, a series resonant network with one end tied to ground appears predominantly at IC1-b's inverting input, raising gain for resonant frequencies. When pots are fully CCW, series resonance forms a divider with R4, reducing gain for the resonant frequencies. The signal output at the end of R4 is buffered by IC2-a; signal couples through C5-R5-R6 to the output path. The signal path is noninverting.

Use

Switches & pots have these functions:

R2	preamp gain trim
R5	output level
R7	64 Hz boost/cut
R8	250 Hz boost/cut

TONE-O-MATIC II PARTS LIST
Resistors
R1 470
R2 10K trimpot
R3, 4 22K
R5 10K audio-taper pot
R6 100
R7, 8, 9, 10, 11 100K pot
Ra, b (×5 see schematic for values)
Capacitors
C1 10pF
C2 10µF aluminum electrolytic
C3, 4 100pF
C5, 8 10µF nonpolar electrolytic
C6, 7 220µF aluminum electrolytic
Ca, b (×5; see schematic for values)
Semiconductors
D1, 2 1N4001
IC1 TL072 dual op amp
IC2, 3, 4 LM833 dual low-noise op amp
Miscellaneous
circuit board, wire, solder, 9V battery snaps, jacks, pots, etc.

3.25" × 3.5" reference box

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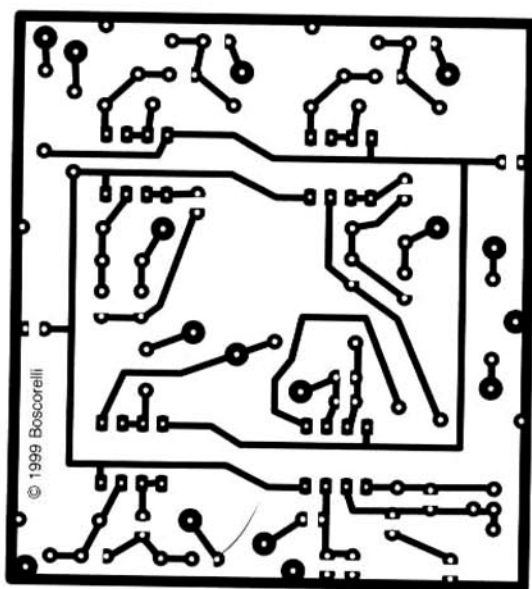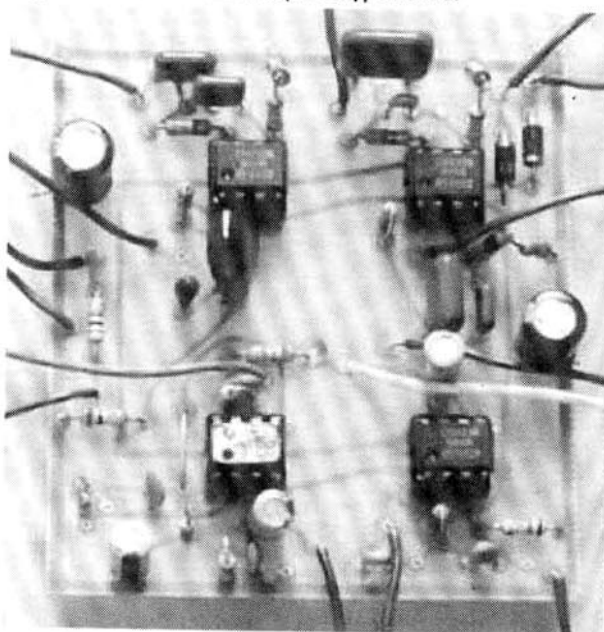
Fig. 25–1. Tone-O-Matic II circuit board.

Fig. 25–2. Tone-O-Matic II prototype board.
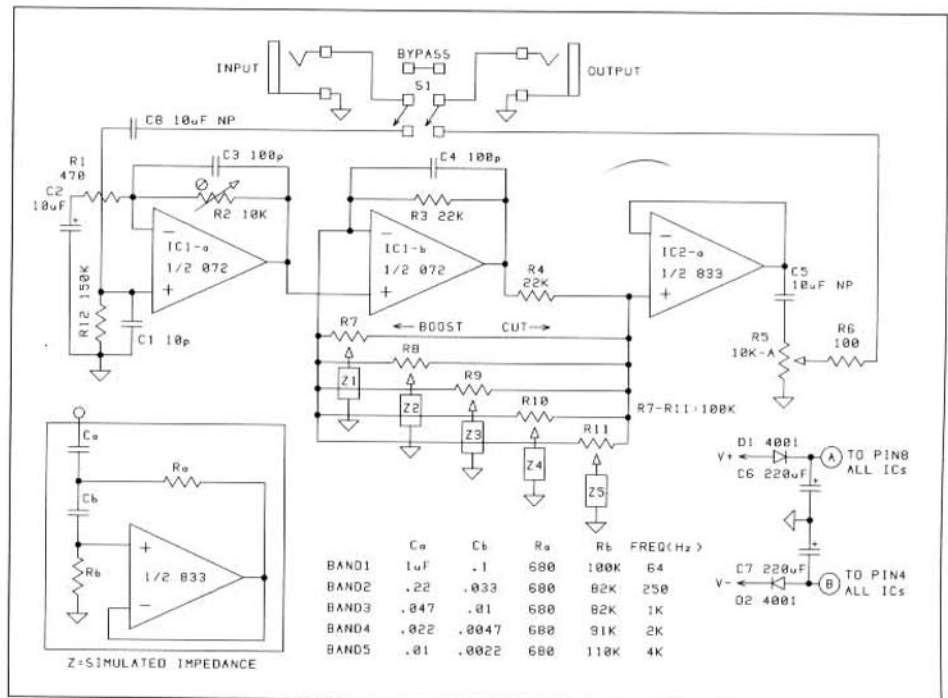
Fig. 25–3. Tone-O-Matic II schematic.

Fig. 25–4. Tone-O-Matic II layout & wiring diagram.
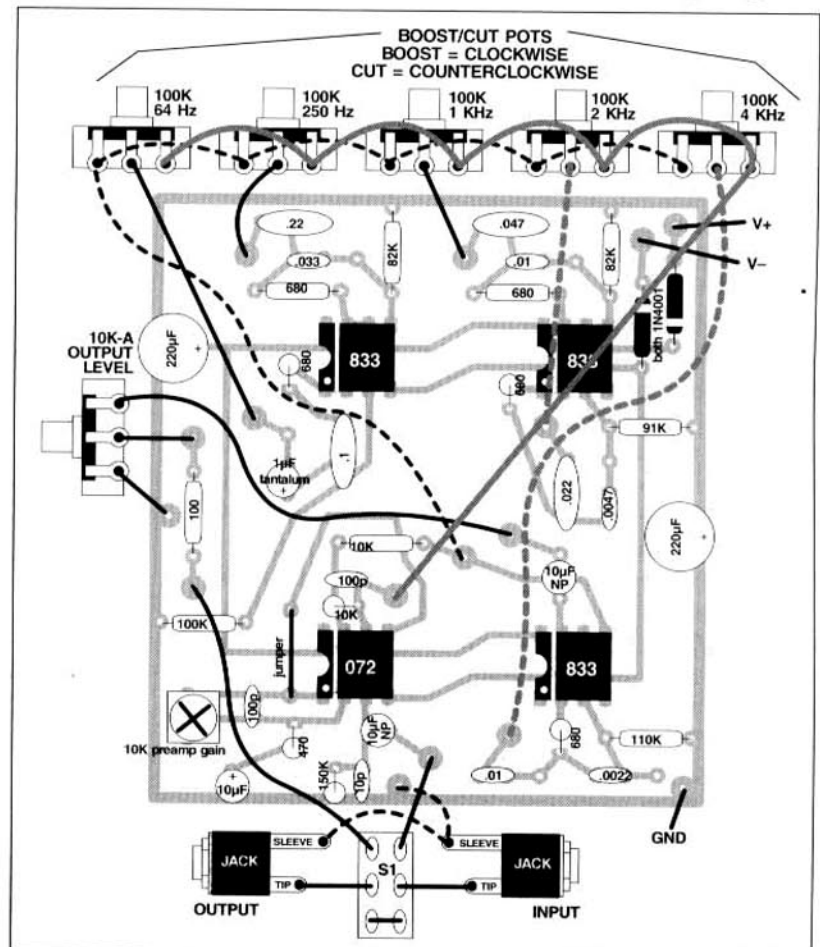
R9 1 KHz boost/cut
R10 2 KHz boost/cut
R11 4 KHz boost/cut
S1 equalize/bypass

Initial settings: R7-11 centered; R2, R5 fully CCW. Connect unit to audio source and target device, such as amp or recorder. Adjust R2 & R5 to give desired monitoring level. Take each boost pot through range and note effect on tone.

Notes

Preamp output should not exceed 1V$_{p-p}$ to avoid overloading the simulated inductors. Boost/cut in the prototype measured ~±8 (±18 dB) one band at a time; a few dB more when adjacent bands are engaged.

The ideal caps for Ca and Cb are polyester types rated 100V and with 5% or better tolerance. Pads on the circuit board accommodate two cap widths. R2 can be a trimpot, a fixed resistor, or an externally mounted pot; the prototype used a fixed resistor.

85

Project No. 26
Tremolo-Matic III

TM3 gives the player control of waveform symmetry and duty cycle. The result, in a sense, spans the gap between tremolo and analog synth.

Circuit Description

Signal Path: Instrument feed couples through C19 to input of IC1, a noninverting preamp whose gain is fixed at ~48, nearly 34 dB. Preamp output couples through C20-R38 to signal input of IC3, half an NE570 configured as a VCA whose gain varies 0–1, depending on the control voltage applied through R46. IC3 output couples through R43-C23 to inverting buffer IC8-a, thence through R45-C25 to the output path. R44 varies the output level. The signal path is noninverting.

Tremolo Control Path: IC2 is an XR2206 function generator configured to generate squarewaves and triangle waves. S2 selects waveform shape; R28 varies the squarewave duty cycle over the range ~10–90%, or varies the triangle function smoothly from ramp to triangle to negative ramp. Oscillation rate is controlled by the capacitor selected by 12-position rotary switch S3. Function output couples to amplitude control pot R31, thence through R30 to inverting buffer IC9. R32 varies the DC offset at IC9's output, which varies the VCA's resting gain. Control voltage couples through R46 to the control port (pin 1) of IC3.

Noise Gate Control Path: IC1 output couples through R1-C1 to modified AC log amp IC5-a, whose output is rectified in IC5-b/-c; boosted in IC5-d, where a variable DC offset is applied by R8 to vary the gating threshold. IC5-d output feeds a positive peak detector made up of D5-Q1-C6, which gives a linear decay of about 2 seconds when R49 sees ½V+; decay accelerates if voltage applied through R49 drops below ½V+. IC6-a output feeds an auto-variable attack network made up of R13-D6-R14-C7. Output of the attack network is buffered by IC6-b, whose output is held at the negative limit by the positive potential applied through R10. IC6-b output couples through R12 to the base of Q2. In the absence of a positive voltage flowing through R12, Q2 is held ON by R39, shunting IC3 pin 1 to ground, muting the VCA. When the preamp

TREMOLO-MATIC III PARTS LIST
Resistors
R1, 3, 4, 5, 12, 24, 25, 43 10K
R2, 13, 18, 33, 34 100K
R6 1K
R7, 22, 23, 29, 30, 37 22K
R8, 44 10K audio-taper pot
R9, 10, 11, 15, 20 39K
R14, 16, 26, 27, 39 4.7K
R17, 19 1M
R21 2.2K
R28 100K pot
R31, 32 10K pot
R35 470
R36 150K
R38 33K
R39 220K
R40 100K multiturn trimpot
R41 220K
R42 7.5K
R45 100
R46 47K
R47, 48 36K
R49 10M
Capacitors
C1, 5, 7, 8, 9, 11, 14 0.1μF
C2 220pF
C3, 4, 10, 17, 24 100pF
C6, 21 0.001μF
C12, 16, 19, 25, 27 10μF aluminum electrolytic
C13, 15 0.01μF
C18 10pF
C20, 23 1μF nonpolar electrolytic
C22 22pF
C26 220μF aluminum electrolytic
Ca–Cl see text
Semiconductors
D1, 2, 3, 4, 5, 8, 9 1N914
D6, 7 red LED
D10 1N4001
IC1 OP-27 low-noise op amp
IC2 XR2206 function generator
IC3 NE570 compander
IC4 LM78L05 5V positive regulator
IC5, 7 TL064 quad low-power op amp
IC6 TL062 dual low-power op amp
IC8 TL072 dual op amp
IC9 MC33171 op amp
Q1 2N3904 NPN transistor
Q2 2N3906 PNP transistor
Miscellaneous
wire, solder, circuit board, jacks, knobs, etc.
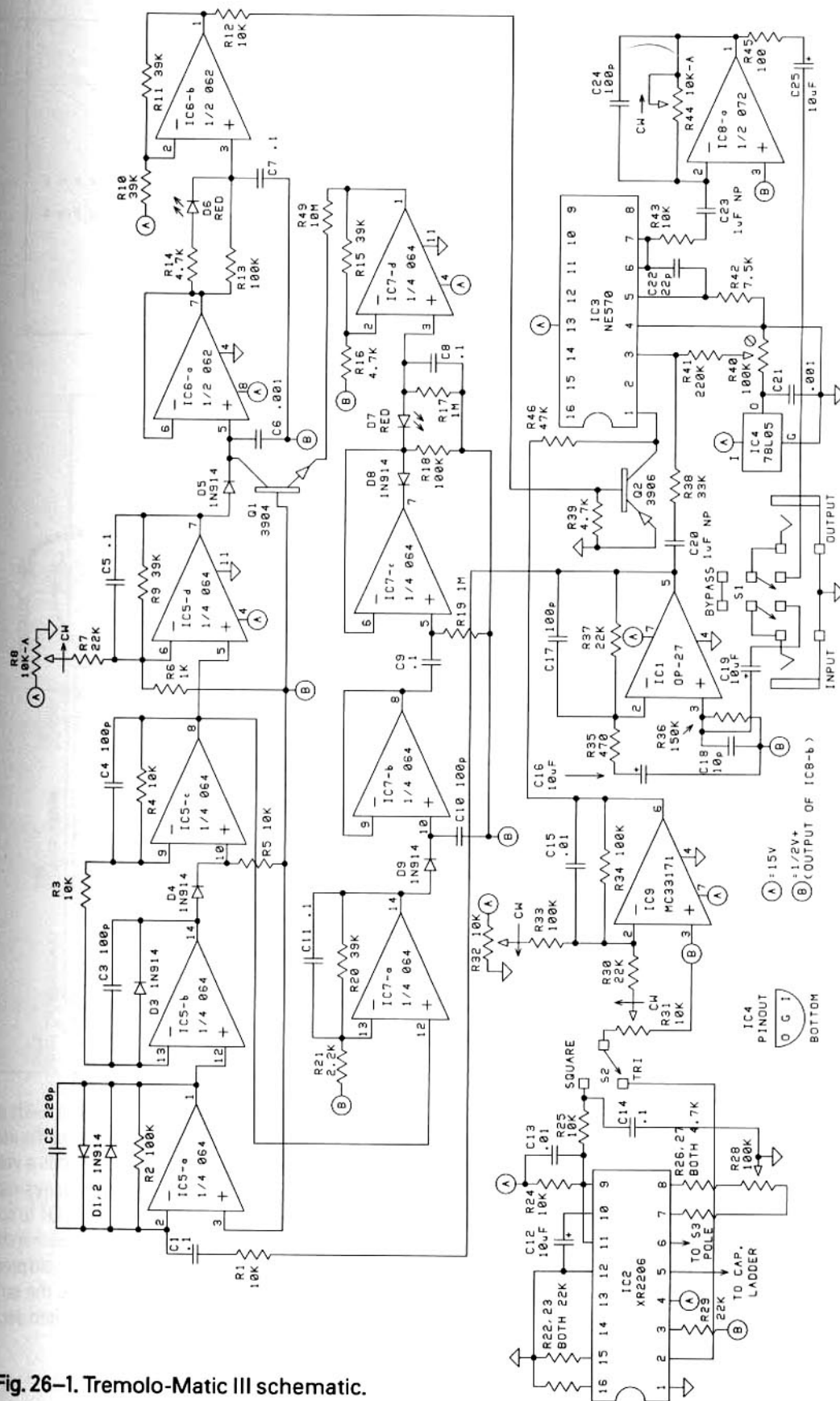
Fig. 26–1. Tremolo-Matic III schematic.
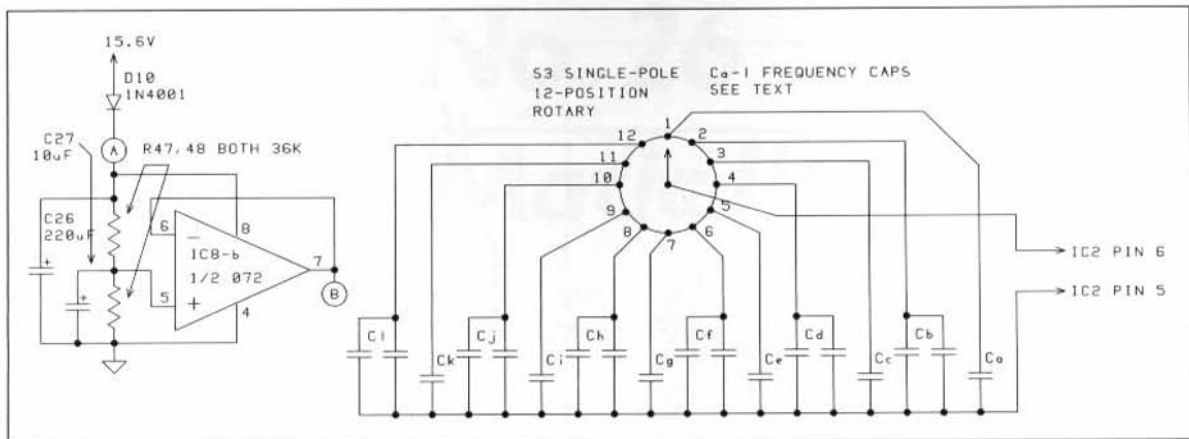
Fig. 26–2. Tremolo-Matic III supplemental schematic.

5.5" × 3.75" reference box

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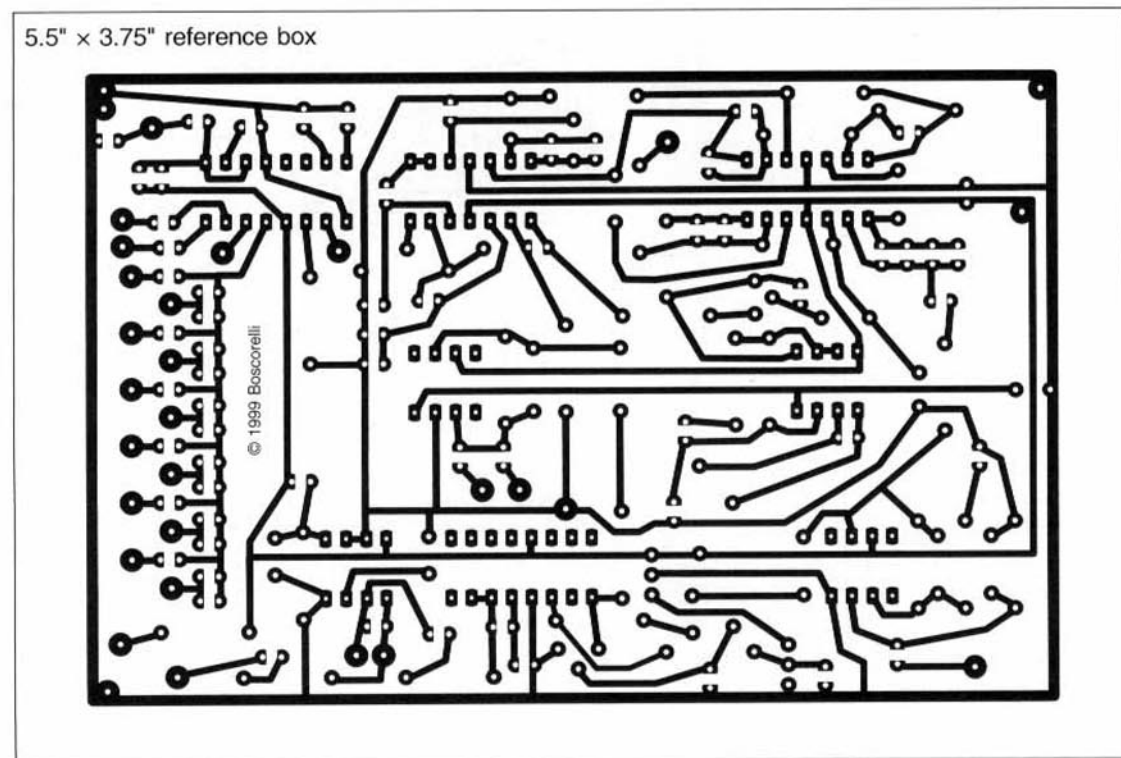
Fig. 26–3. Tremolo-Matic III circuit board.
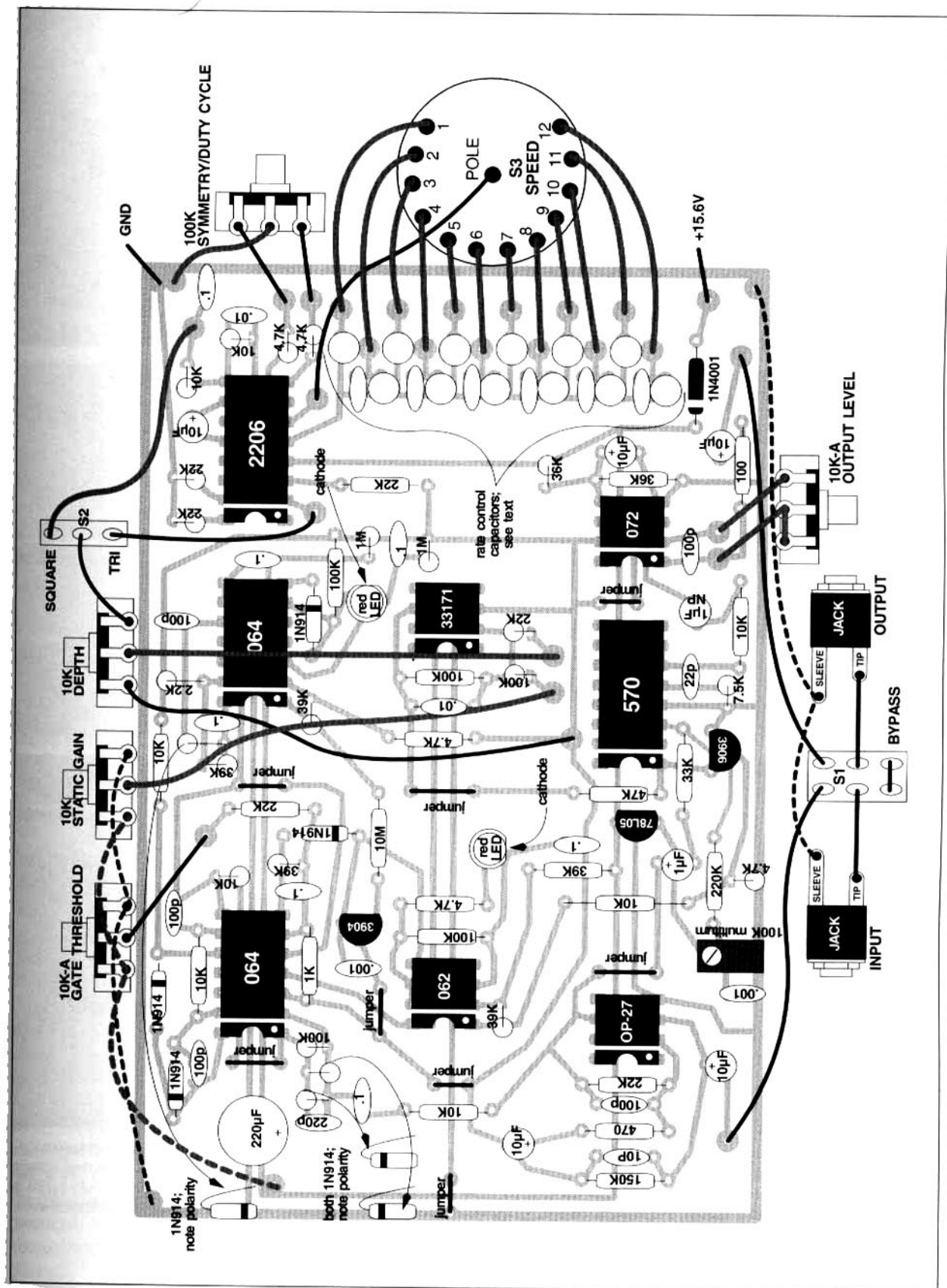
output reaches ~200 mv$_{p-p}$, the output of IC6-b swings positive enough to turn Q2 OFF, and VCA gain become a function of the static/dynamic voltages conveyed through R46. When preamp output drops below ~200 mv$_{p-p}$, the output of IC6-b swings negative enough to turn Q2 ON, muting the VCA.

Slow decay of the primary level detector gives low distortion as notes (especially bass) decay; but beating may be heard for several seconds before Q2 turns ON. This makes it desirable that the gate close swiftly when the player mutes the strings. TM3 includes a second level detector to this end: IC5-c output feeds scaling amp IC7-a, feeding an integrator made up of

D9-C10, buffered by IC7-b, whose output feeds a falling-edge detector made up of IC7-c/-d and the associated components. This detector generates a voltage below $\frac{1}{2}$V+ only when the program decays rapidly. The voltage flows through R49, causing Q1 to source a negative current that neutralizes the positive charge in C6 in less than 40 ms. With gate threshold properly set, TM3 mutes when the player mutes the strings, with no audible lag, yet allows uninterrupted decay of held notes and chords.

Use
Switches & pots have these functions:

Fig. 26–4. Tremolo-Matic III layout & wiring diagram.
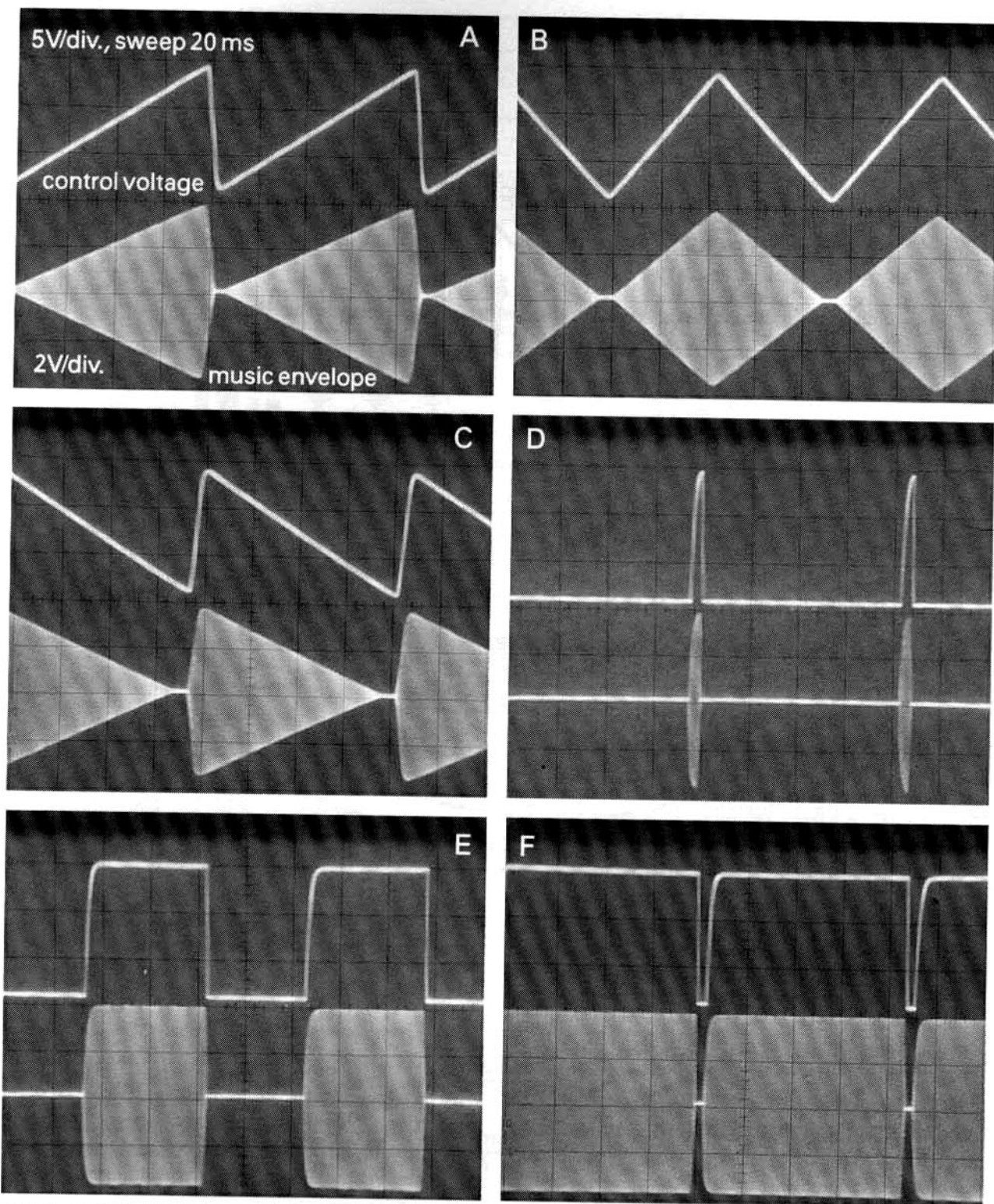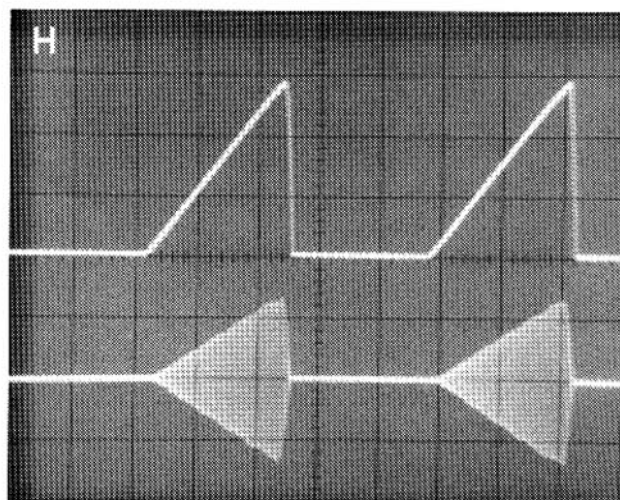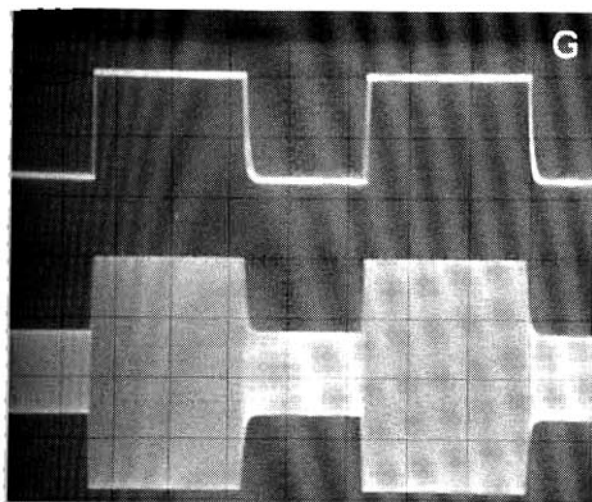
Fig. 26–5 (this and facing page). TM3 control voltage top trace, music envelope bottom trace. A—Ramp, B—triangle, C—negative ramp. Ramp and negative ramp sound significantly different, especially at slow rates. D/E/F—Control voltage switched to quasi-squarewave, duty cycle varies 10%/50%/90%. These, too, change the sound distinctively. Duty cycles in the 30–70% range sound musical; those outside that range better suit special effects. G/H/I (facing page)—Additional waveforms generated by TM3.

R8	noise gate threshold
R28	triangle symmetry/squarewave duty cycle
R31	tremolo depth
R32	VCA static gain, 0–1
R40	VCA feedthrough trim
R44	output level
S1	effect/bypass
S2	waveform shape (square/triangle)
S3	tremolo rate
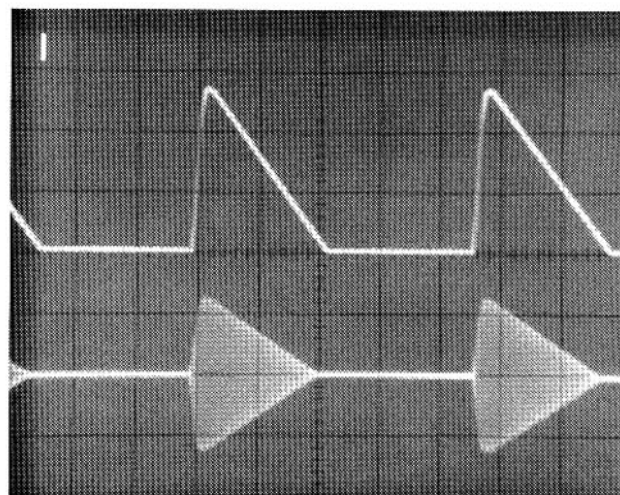
First, trim feedthrough. Short the input. Toggle S2 to triangle, turn R8 fully CW to deactivate the noise gate; center R28 and R32, turn R31 fully CW. Set S3 for maximum tremolo rate. In this state the control voltage clips at both extremes. Attach scope probe to IC3 pin 7. Trim R40 for minimum feedthrough.

If no scope is available, trim feedthrough by ear. Configure settings as above, connect unit to amp whose volume is turned all the way down. *Slowly* advance the volume until feedthrough artifacts are heard. Depending on the initial setting of R40, these could measure up to several volts$_{p-p}$. Trim R40 for minimum feedthrough.

Without changing the settings, connect axe to unit; establish desired amp volume. In this state deep tremolo should be noted. Take rate control S3 through its range. With a slow rate selected (e.g., rate cap 10µF), explore the sounds of ramp to triangle to negative ramp.

Toggle S2 to squarewave (actually, a quasi-squarewave; rising and falling edges are slowed by C13/14/15). Maximum depth occurs with R32 fully CCW, R31 fully CW. In this state gain varies 0–1 with each pulse. Vary the duty cycle and note the different types of sounds produced. Minimum duty cycle gives a percussive sound; maximum duty cycle gives the effect of double-picking.

Next, adjust the noise gate threshold. Turn R8 CCW until gating occurs. Properly set, the gate lets notes or

chords hang as long as the player likes, yet mutes instantly when the player mutes the strings.

Notes

TM3 is optimized for a 15V supply. The nature of the circuit makes departure from this voltage imprudent. The prototype was tested with 15.0V on the positive supply bus (15.6V into D10).

The 2206 configuration allowing variation of squarewave duty cycle and ramp slope happens to make continuously variable rate difficult to achieve, so changing speed requires a bank of switched capacitors. Half the pads on the circuit board accommodate two caps, to let the builder realize marginal values. The prototype contains only two caps, 2.2µF to 10µF, for the approximate rates 4.5 Hz & 1 Hz. The board holds space for caps giving up to 12 speeds.

Feedthrough trim of the 570 VCA is so good that gating is not needed when using the triangle waveforms, including those that clip. Squarewave control results in audible artifacts. Gating proved the most practical means to quiet the effect at rest while retaining squarewave-like modulation capability.

The unit gives best S/N ratio when the level coming
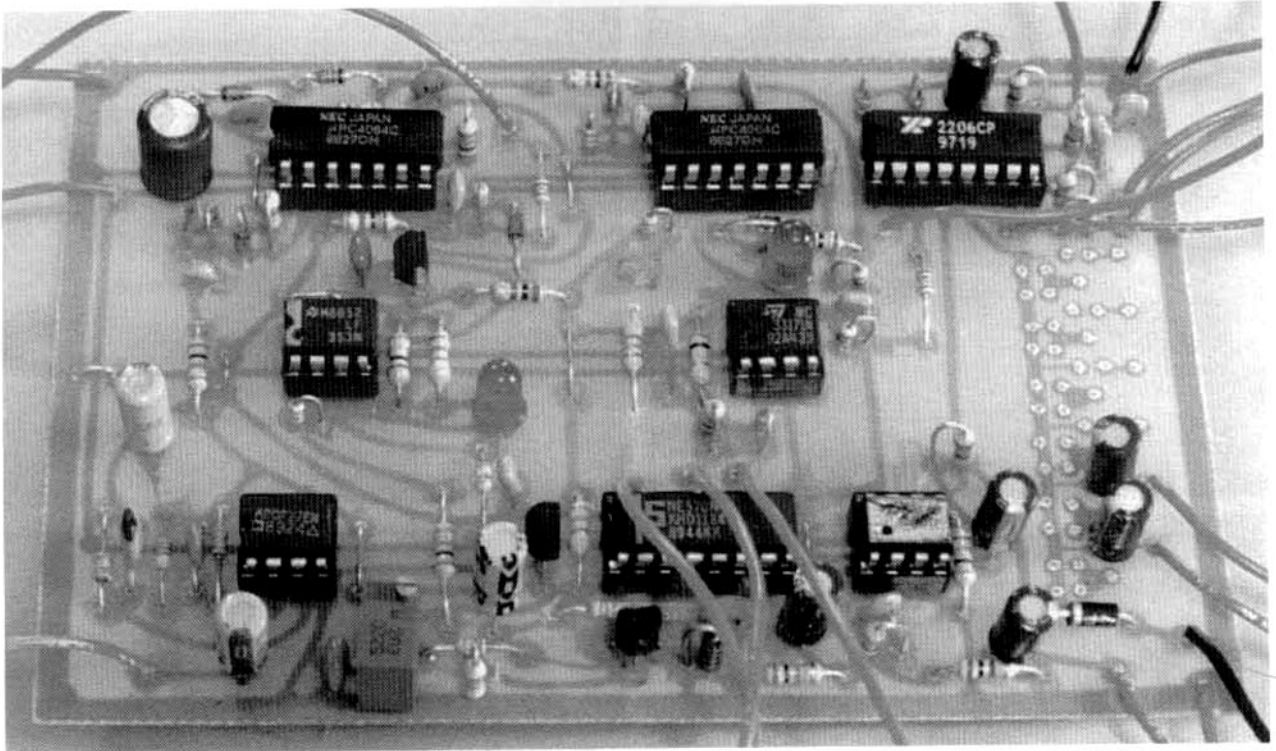
Fig. 26—6. Tremolo-Matic III prototype board.

off the preamp averages at least $3V_{p-p}$. The inputs of most amps will require a good bit of gain reduction by R44 to avoid overload, adding a quasi-companding aspect to the system. The prototype's preamp gain was fixed at a very high level to suit a specific guitar. Lower the value of R37 to suit other instruments.

Net gain of cascaded signal and control stages exceeds 60 dB, so keep wiring short and neat. The prototype did not require shielded input or output leads, but this could change depending on the builder's mounting arrangement. Input and output signals are in phase, with enough gain to cause oscillation if their leads are juxtaposed.

Appendices
Stomp Box Ingredients

Builders cooking up their own effects need the raw fixin's for tone shaping, compression, distortion, and a lot more. While nothing special surrounds these ingredients, they seldom surface in a heap, cut and sorted for the stomp-box scene. Libraries bulge with circuit compendia, but few circuits advertise their stomp-box attributes. Only a seasoned builder sees that a frequency-doubling ramp generator belongs in a thrashmetal pedal.

To help adventurous builders advance their craft, the text has gathered a nucleus of stomp-box building blocks and related lore. The presentation assumes the reader to have reached the stage of breadboard debugging and fine tuning.

Good luck—and great stomping!

Anatomy of a Stomp Box

Stomp box defines a discrete, floor-dwelling effect for electric guitar or bass, typically housed in a wedge-shaped case and powered by batteries. No one seems to know when the term took hold, but it had to have trailed the dawn of fuzz in the fifties, and prefaced the proliferation of effects in the seventies. By 1980 the flying door-stop had given way to signature pedals whose aspect declared their pedigree. Evolution continues. Some of the current crop can't seem to decide whether they belong on stage or in the Museum of Modern Art.

In the early days a single box fit between axe and amp. As new effects appeared, players tested the sound of fuzzed wah and wahed fuzz. So began the practice of chaining. Manufacturers encouraged this market by adapting their pedals to series wiring. Modern guitar setups routinely chain effects.

Though born of distortion, *stomp box* grew to embrace any effect, preamp, or switching function. Evolution has come full circle in a reissue craze fueled by the quaint view that at least some of the old boxes had it right.

Analog functions now available include:

- ► compression/sustain
- ► direct/isolation box
- ► distortion
- ► envelope follower/auto-wah
- ► EQ
- ► flange/chorus/phaser
- ► frequency divider/multiplier
- ► mixer
- ► noise gate
- ► pedal volume
- ► pedal wah
- ► preamp
- ► reverb
- ► splitter
- ► tremolo/vibrato
- ► tube-sound emulator
- ► boxes combining effects

Culling a recent issue of a guitar buyer's guide, the author counted 111 boxes by 10 manufacturers. The profile comprised 30% distortion, 22% flange/phase/chorus, 13% wah-type effects, 9% compression/sustain, 5% EQ, 21% 'other', including preamps, switches, and miscellaneous effects. Most run on 9V batteries; power from a wall-wart is a common option.

The outward hallmarks of a stomp box flow from convenience in use and conformity with ritual. Internally, pedals differ little from any small electronic device, enabling use of standard construction techniques, with the general rule to keep wiring as short and direct as possible. Low gain, short signal paths, and shielding offered by a metal box obviate shielded cable for most internal connections.

The nature of the stomp-box signal path inflicts the issue of clickless switching. Clicks that haunt CMOS and JFET switches stem partly from leakage of the control voltage into the signal path. 'Clickless' electronic switches, like the SSM2402/2412, combine low charge injection with networks that slow the edges of the squarewave control pulse, generating much less audible energy than typical electronic switches.

Bleeder resistors prevent a charge from accumulating in coupling capacitors, avoiding a plug-in pop.

Most builders will find it prudent to hardwire the bypass using a DPDT rather than a SPDT switch. The DPDT switch takes the effect completely out of the
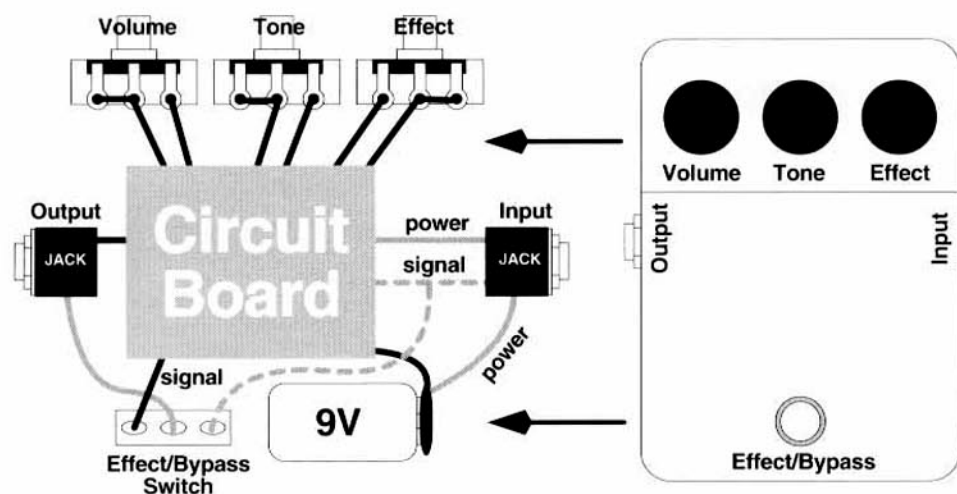
Fig. A1. Diagrammatic rendering of exterior and interior of a typical stomp box. Signal enters through input jack, which usually contains the power switch. The stomp-switch selects the output of the circuit board, or bypasses the input to the output.
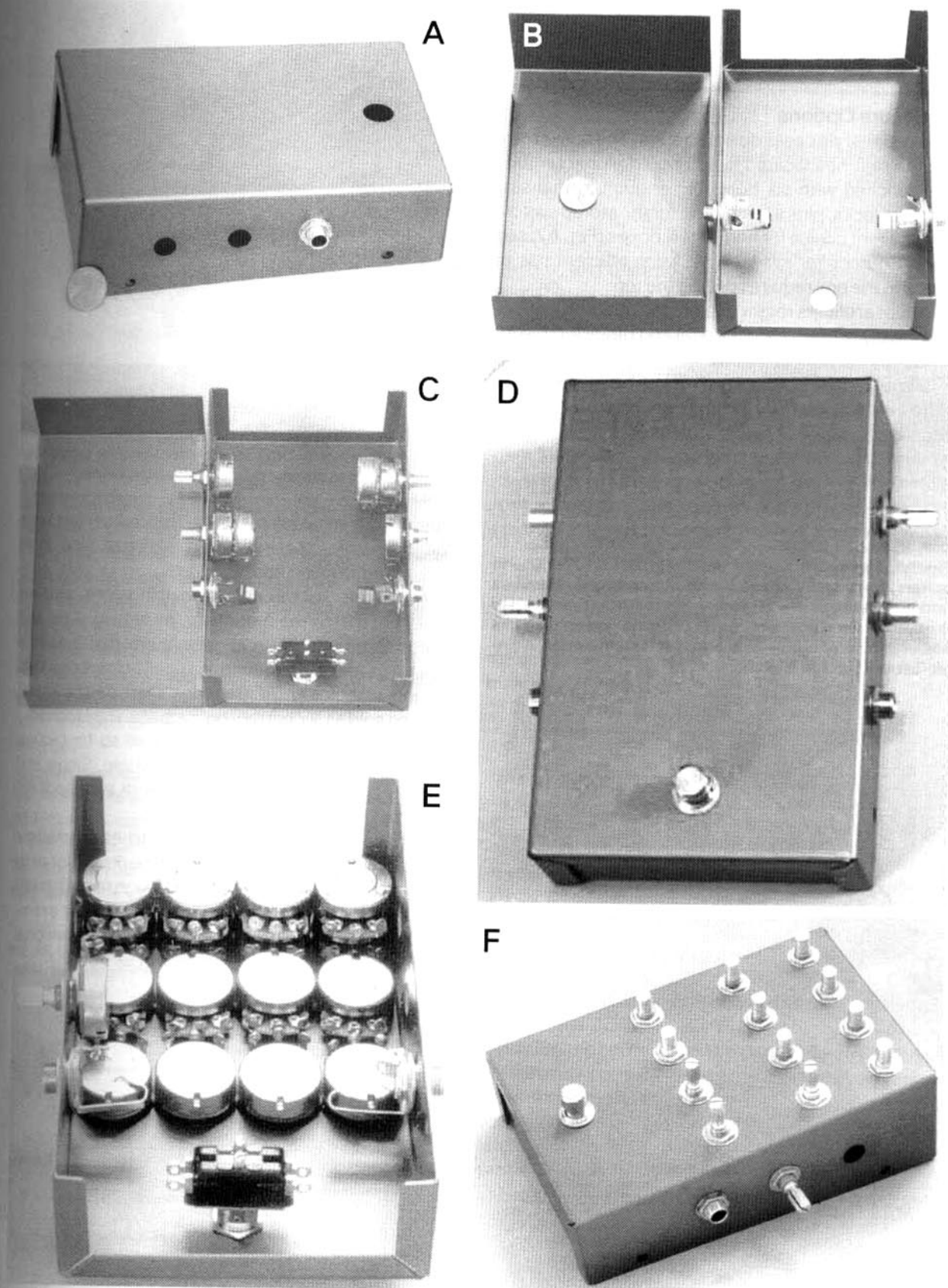
Fig A2. Generic stomp-box enclosure. A & B—Blue-toned aluminum pre-punched with six holes big enough for full-size pots, plus one over-size hole for a large stomp switch. Jacks are not supplied with the enclosure. C & D—Pots & stomp switch mounted. E & F—Drilling the top of the enclosure allows more efficient use of space; in this case, 13 full-size pots, two jacks, and the stomp switch are mounted, with room left for the Quad Parametro-Matic circuit board, a power switch, and maybe a pair of 9V batteries. See Parts Sources for availability.

signal path, replacing it with straight wire. This lets the player bypass the box even if the batteries fail, and eliminates loading from the effect's input circuit.

Enclosure Options

Raw pedal-style cases do not abound. The only example the text has located is an angled aluminum box pre-punched with six holes big enough for full-size pots and jacks, plus a larger hole for the stomp switch; an excellent choice for homebrew boxes (Fig. A2; see Parts Sources for availability). Most efficient use of space in the box requires drilling the top.

Adept artificers might explore the option of nondestructively gutting a stock pedal and replacing the innards with custom electronics.

Generic project boxes will never be mistaken for retro classics, but make cheap, serviceable options. Steel boxes stand up under stomping, aluminum boxes tend to crumple, but steel usually costs more than aluminum, and is difficult for those with no metalworking experience to work. While plastic boxes make convenient testing platforms, lack of shielding leaves

Fig. A4. Heavy-duty DPDT switch (left) ideal for stomp boxes, compared to standard-size pushbutton switch.

them iffy for sessions and gigs. Thick rubber feet keep the box from slipping when stomped, and from scratching wooden floors.

As for internal mounting, anything goes, with the caveat to match the mounts to the setting. A gig box needs to be a lot sturdier than a box that lives in a project studio. Spring-loaded D-cell holders do a better job of holding 9V batteries than do the metal clamps sold as 9V battery holders.

The enclosure options mentioned so far require opening the box for battery replacement. Single and dual 9V battery holders equipped with swing-out access make handy options (Fig. A3).

Locations of pots, jacks, and switches are matters of taste. The stomp-switch should be placed for easy stomping, and should be heavy enough to hold up under the stress. Fig. A4 illustrates one suitable example; see Parts Sources for availability. Conventional push-on/push-off switches suffice in light duty.

Fig. A3. Swing-out dual-9V battery holder sold for use in guitars, but readily adaptable to stomp boxes. Because nothing but the user's diligence guards against insertion of the battery with incorrect polarity, circuits powered off this holder should incorporate polarity-protection diodes. A single-battery version is available.

Power Supply Options

Batteries

Traditional stomp boxes run on one or two '9V' batteries, something of a misnomer, for no 9V battery sustains a 9V output over its working life. Voltage falls with use. Typical battery discharge curves illustrate this trait (Fig. A5). While a stomp box must be built with a nominal supply voltage in mind, practical designs incorporate tolerance to voltage changes.

Nine-volt batteries come in four key categories. Carbon-zinc types offer such low power density as to constitute a waste of materials; high cost offsets lithium batteries' great energy density, leaving alkaline and nickel-cadmium as the viable options.

A fresh alkaline 9V measures about 8.4V under load, ~7.8V after the drop across a polarity-protection diode. (That diode prevents reversed voltage should the battery terminals briefly touch the battery snap wrong-way, as could occur while changing batteries in the middle of a gig. Reversed polarity can destroy some semiconductors, and cause tantalum bypass capacitors to short.)

With alkaline batteries hovering near $3 apiece, nicad rechargeables make the sensible option for long-term portable power. One rechargeable replaces hundreds of dollars worth of alkaline, saving resources in the process.

Late-model nicads no longer suffer the memory effect, in which failure to drain the battery fully prevented it from delivering a whole charge in later cycles.

Common nicads read 8.4V hot from the charger; 7.5–8.1V after idling for a few days, 7.2–7.5V under load, leaving ~6.6V after passing through a polarity-

protection diode. Nicads have a self-discharge rate of 1–2% per day, and their energy density is substantially less than alkaline batteries'.

Whatever the battery choice, the builder will find it instructive to measure the average current drain of the box, and thereby project likely working life with a given battery type.

Single Supply vs. Dual Supply

Traditional stomp boxes run on a single 9V battery, but the text finds nine volts wanting. Nine-volt boxes run out of gas just as things get interesting. An alkaline Duracell® measures close to 7.8V by the time the electrons hit the circuit, which loses another 2V of headroom because common op amps do not swing rail to rail, leaving headroom of ~6V$_{p-p}$, tops. Careful design makes this enough, but ploys should not be necessary in homebrew boxes. Several high-end chips won't run on less than 10 volts, and don't truly have room to do their thing at that level.

Low voltage limits the designer severely enough to make the second battery worth its while. And a box built for a dual supply can later step up to the chips' limits (usually ±18–22V) with no circuit mods.
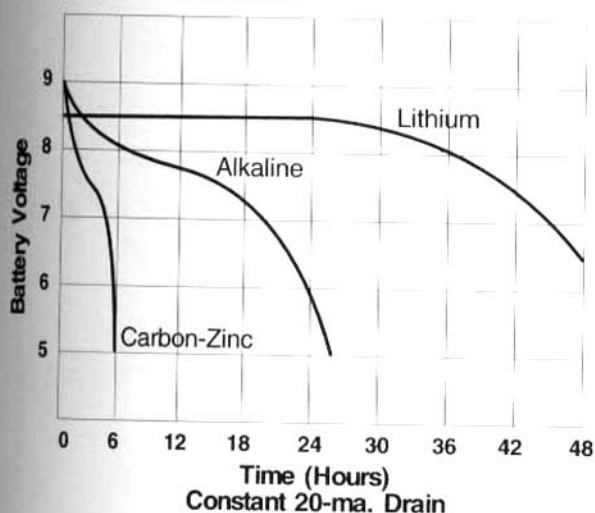
Deriving a Dual Supply from a Single Supply

The simplest way to get a dual supply from one battery is known as rail-splitting. Many circuits get by with a simple resistive divider; high-gain circuits, or those having many active components, usually use the output of an op amp as the artificial ground, because it both sources and sinks current.

A voltage mirror creates a negative image of the positive supply using a charge pump; just about any ultrasonic oscillator juices the pump (Fig. A6–1). Common examples include the 555 timer, and chips built for the purpose, like the 7662. This approach yields limited current capacity, and a negative potential that may not match the positive. Whatever current flows through the negative output must come from the battery, shortening its life. The output is unregulated; voltage falls with load, making the approach practical only in selected circuits.

Second- and third-generation switching converter chips run more efficiently, providing a symmetrical dual output from a single supply. The prospective user should read these chips' data sheets carefully. While the chips indeed switch in the ultrasonic range, some of them pulse on and off in bursts that generate large audio-band artifacts.

Fig. A5. Generic discharge curves typical of carbon-zinc, alkaline, and lithium 9V batteries.

AC-Derived Supplies

Wall warts are encapsulated transformers that plug into the AC wall socket and furnish low-voltage AC or DC. They eliminate the risk of dealing with the full AC line voltage (though no voltage should be handled carelessly), and distance a strong hum-source from the signal path. DC models supply anything from a raw rectifier output to filtered and regulated voltage.

A box running off an AC wart must include an on-board rectifier and filter (Fig. A6–2, 3); sometimes a regulator. AC-powered boxes suffer greater risk for hum than battery-powered boxes, and should use single-point grounding.

Because one wart per box is cumbersome, and since one wart can power many boxes, a common tactic groups all the pedals on a board and runs them off a common supply. The builder choosing this approach should avoid daisy-chaining boxes on the power bus. Power leads from each box (ground especially) should return to a single point.

Small DC–DC converters make attractive options for wart-powered boxes. These modules step 5VDC up to ±5, 7, 10, 12, 15, or 17V, and come in 1W and 2W versions. Output current is inversely proportional to voltage. A 1W ±5V module supplies 100 ma; a 1W ±17V type only 30 ma. The converters are smaller than the end of your little finger, and offer 100% isolation from their 5VDC source. Even with bypass caps, they occupy no more space than a conventional AC-to-DC supply. Best of all, incorporating a converter module does not demand redesigning the board. Simply append a converter supply to the main board (See Project No. 27).

When used in a rack powered by a single wall wart, converter-powered boxes can be daisy-chained without fear of ground loops, because the 5V power bus is isolated from each converter's output ground. The 5V supply needs to be robust: 1W converters take up to 250 ma; 2W models may need up to 500 ma apiece.
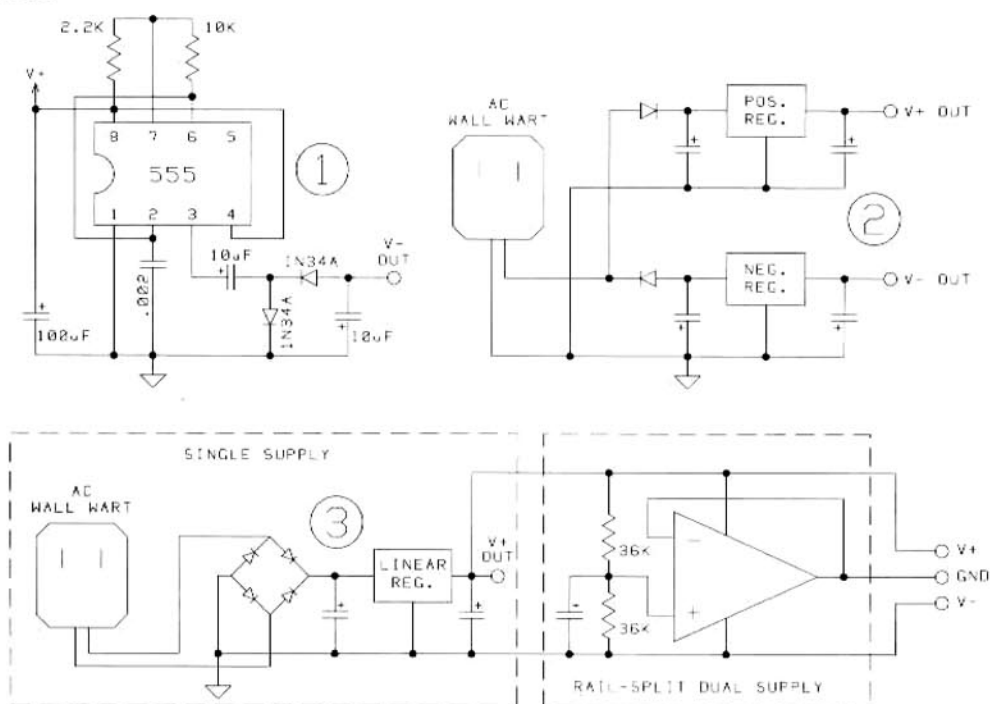
On the downside, converters contain inductors, leaving them prey to hum from electromagnetic fields. This tends not to be a problem so long as they don't sit within a foot of a power transformer. The output contains ripple at several hundred kilohertz, usually not noticed in the audio band, but a potential source of aliasing in digital gear. Converters demand care in the breadboard stage, for accidentally shorting an output can destroy them. Efficiency falls in the 70–85% range. Recent prices: $9 for 1W modules, $14 for 2W.

Gray Areas

Followers of the axe/tone scene may have read anecdotes claiming a backdoor link between battery type and stomp-box tone; specifically, accounts citing better—or at least altered—sound when using rechargeable vs. carbon-zinc vs. alkaline batteries; and fresh vs. aging samples.

These claims suggest several possibilities. First,

Fig. A6. 1—Typical ultrasonic charge-pump voltage inverter. Output magnitude is >1V less than input; highly load-sensitive; 10 ma makes a practical limit. 2—Typical AC wall-wart–derived dual supply. 3—Another AC wall wart-derived supply using a bridge rectifier; raw single supply output, or rail-split to give an artificial dual supply. Circuits #2 & #3 assume the supply to be hardwired to effect.

sundry battery types furnish different voltages. This most demonstrably affects headroom. Players may be hearing distortion and compression due to reduced headroom.

Second, different battery types, including different brands within a type, exhibit disparate internal impedances. The ideal battery has zero impedance; it acts as an immovable current/voltage source. A real-world battery's finite impedance leads to voltage drop under load. Furthermore, this trait shifts as the battery ages, or with state of charge in a nicad.

Third, some circuits exhibit instability when the available voltage or current falls below a certain point. This instability could manifest as desirable distortion.

Since the point of the exercise is to help players take control of their sound, and to resolve a hazy issue, let's concoct a failing-battery simulator. Fig. A7 shows one speculative rendering. Battery pack B1 must have stout capacity, say, 12 nicads at least AA size in series; a solid 15V. This feeds a three-terminal regulator to vary voltage from 1.2V to the supply limit (by R1), less the regulator's dropout voltage.

A wirewound pot in series with the regulator output (R3) simulates various battery impedances. The pot
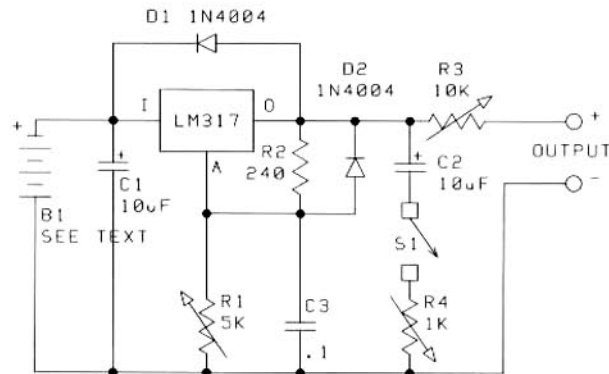
Fig. A7. Schematic of putative failing-battery simulator. See text.

must pass the necessary current without sustaining damage. A 2W rating should be adequate.

Since battery aging places heavier demands on bypass caps, a simulator needs the ability to impair bypass, or to remove it completely. S1 & R4 serve this end, which may cause instability in some circuits. (Experimenters should be aware that powering a box off this device could damage the box, and probably voids the warranty.)

Project No. 27

Append-A-Board Power Supply

Batteries make an impractical option when a box needs the ±15V necessary for serious headroom. That voltage is available from DC–DC converters now selling for $8–$14. These devices step 5VDC up to ±5 to ±17VDC, at current capacities that power most stomp boxes.

Circuit Function

Raw power input couples through halfwave rectifier D1 to input of 5V regulator IC1, whose output feeds DC converter module IC2. C1 filters the rectifier output, C2 filters the output of IC1.

Use

The wall wart should be rated 8–12 volts, AC or DC, @ 500 milliamps. Voltage coming off the rectifier must equal or exceed 7.6V for full regulation. If the power input comes from a DC source, connect the positive potential to the anode of D1.

Choice of converter determines the output voltage, ±5V to ±17V. One-watt modules supply up to 100 ma at ±5V; 30 ma at ±17V. Two-watt modules approximately double this current. Determine the proper current rating by measuring current drawn by the project the module will power. Most projects in this book can get by with a 1W module.

IC1 requires a heatsink whether it powers a 1W or a 2W module.

You can use this board three ways. First, incorpo-

rate it into the design of another project. This requires modifying the circuit board layout to have the module's outputs tie directly to the effect's power bus. Envelo-Matic (Project No. 18) gives an example of this approach.

A second option is to 'append' the circuit board pattern to that of another project, building two separate boards at the same time and on the same blank. Power and ground buses connect through short, insulated jumper wires.

Finally, you can build the supply as a freestanding board, perhaps with an extended skirt to give room for mounting holes, and mount it apart from the project being powered.

Hardwiring obviates the polarity-protection diodes needed when using batteries.

APPEND-A-BOARD POWER SUPPLY PARTS LIST
Capacitors
C1 330μF 25V aluminum electrolytic
C2 470μF 10V aluminum electrolytic
Semiconductors
D1 1N4004
IC1 LM7805 +5V regulator, TO-220 case
IC2 DC-to-DC converter module (see text)
Miscellaneous
heatsink for IC1
wire, solder, circuit board, etc.

Fig. 27–1. Append-A-Board Power Supply schematic.
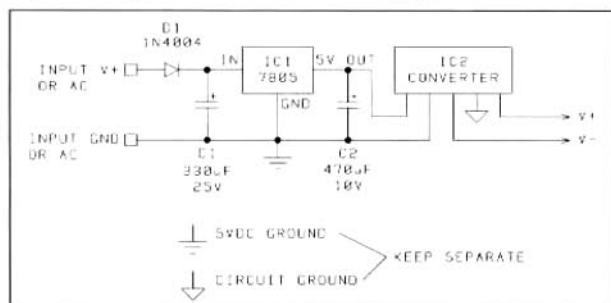
Fig. 27–2. Append-A-Board Power Supply circuit board.

2.5" × 1" reference box

© 1998 Boscorelli
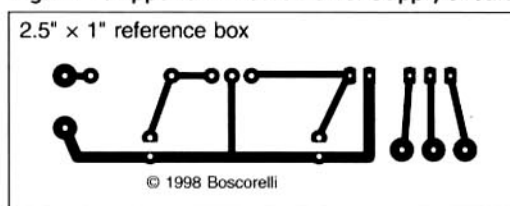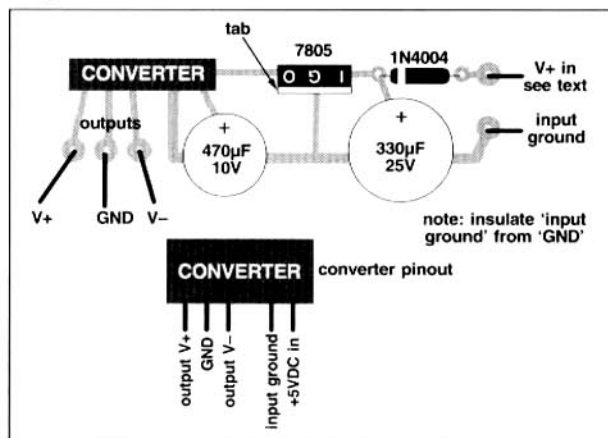
Fig. 27–3. Append-A-Board Power Supply layout & wiring diagram.

Switching
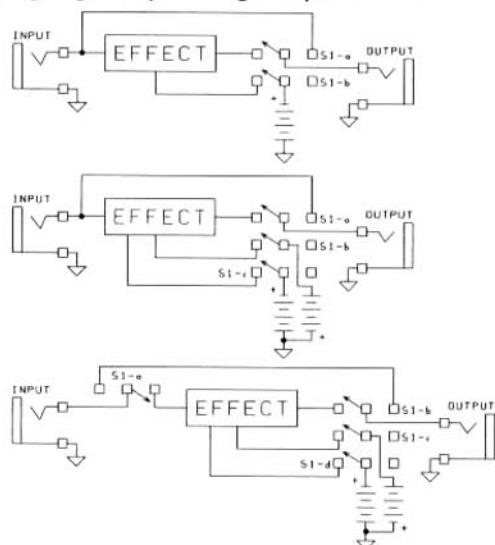
Separating the signal switch from the power switch proves worthwhile in a pedal, if only to avert the massive thump that often comes with powering up a dead circuit. Turn-off thump can be equally bad. Projects shown in this book assume independent power and signal switches, the stomp switch being reserved for the signal path. Builders who prize noiseless switching, or who plan alternative switching setups, may wish to consider options beyond discrete manual switches.

Switches include eponymous devices, as well as electrically controlled channels for which switching is only one job. They're designated in terms of connections called poles and throws. A single-pole-single-throw switch opens or closes two contacts. A single-pole-double-throw switch connects one path alternately to two others; etc. Acronyms evolved to truncate the names: SPST, SPDT, DPDT, 3PDT, etc. Switch diagrams show arrows originating from poles and terminating in throws.

Manual mechanical switches come in toggle, pushbutton, slide, rotary, pull, rocker, pot-mounted, and board-mounted DIP types. Pedals favor a pushbutton stomp switch, supplemented by slide switches if necessary. They generally avoid protruding levers likely to be tripped accidentally, or bent by rough handling.

Switches having two or more throws possess tim-

ing properties. Those that make the new contact before breaking the existing one are said to make-before-break (MBB). Those that break the existing contact before making the next are said to break-before-make (BBM). Timing matters in cases where switching momentarily connects the output of a gain stage to its input, and in circuits carrying enough current that a brief short could damage the switch or the circuit.

Indirect mechanical switches include *relays*, electrically operated switches whose metal contacts move in response to an electromagnet. Common versions hold in one position only on application of power, returning to a second position on removal of power. Latching relays hold in either position, switching state on brief application of power. Relay switching time typically falls in the range 5–30 milliseconds. Time to make often differs from time to break, allowing the builder to define this by choice of relay.

Relay coils are built for specific supply voltages, 5V, 12V, and 24V being common. Coil resistance predicts current. For instance, a 12V relay with a 1300-ohm coil draws more than 9 ma. A 5V relay with a 250-ohm coil draws 20 ma. Generally, the more elaborate the switch, the stronger the field required, and the greater the holding current. Miniature, multi-pole relays would be ideal for pedals, but for the fact that their holding current may dwarf circuit current. These relays are practical only in AC-powered boxes. Latching relays make good options for battery-powered boxes, because they draw current only for a split-second during transition.

Solid state relays are electronic switches optimized for current. They use the same terminology as mechanical relays but have no moving parts. Their operating current is usually less than mechanical relays', but still not trivial in a battery-powered box.

Bipolar and JFET transistors can serve as switches. The collector-emitter path of a BPT, or the drain-source path of a JFET, acts as an open circuit when OFF, a low-impedance path when ON. BPTs switch fully ON or OFF from a single supply; most FETs pinch off several volts below ground.

Fig. A8. 1—Simplest single-switch power/signal arrangement. One DPDT switch controls power and signal. 2—Going to a dual supply means a costlier switch. 3—Dual supply with superior signal bypass means a 4PDT switch. All three approaches could be subject to loud switching transients, arguing for separate signal & power switches.
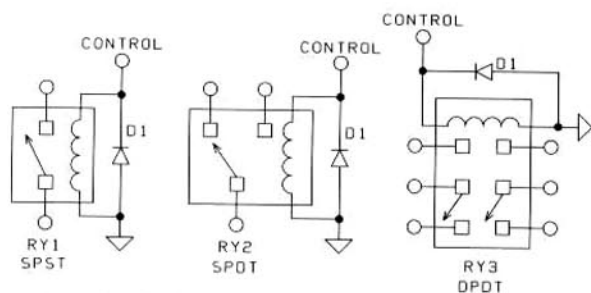
Fig. A9. Typical relay schematics. Diode in parallel with coil is needed to shunt the reverse current generated when power is withdrawn from relay; magnetic field collapses, induces voltage that can damage other circuit components unless shunted by D1, typically 1N4004.
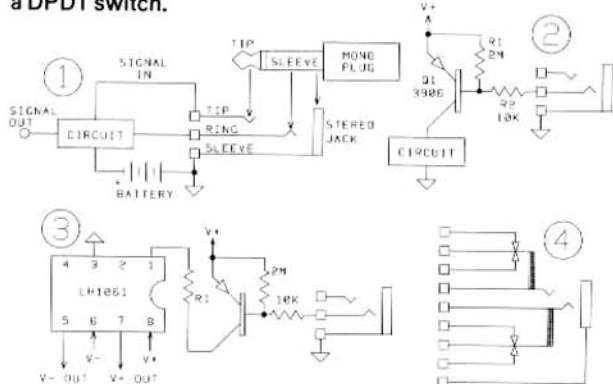
Power Switching

Natural selection has thinned countless options to a few modes of switching power in pedals. Probably the commonest of these is making the input or output jack the power switch. A two-conductor plug in a three-conductor jack ties the jack's center (ring) terminal to its ground (sleeve) terminal. This shunts power directly to the circuit, or drives a secondary switch, such as a transistor.

Just about anything that presents a low-resistance path can switch power. Mechanical switches typically present less than one ohm. Bipolar and MOSFET transistors present less than one ohm to a few ohms. Common FETs (MPF102, 2N5457) present several hundred ohms; specialized FETs exhibit ON-resistance less than 10 ohms (J105, J108, 2N5434).

For power switching, most circuits treat the collector-emitter or drain-source path as a series element.

Fig. A10. 1—Using stereo jack as stomp-box power switch. Insertion of mono plug shorts ring to ground. Circuit is not grounded until plug is inserted. This arrangement has a drawback in that axe being plugged in will momentarily connect power supply through circuit. 2—Using a PNP transistor as a power switch. Q1 is held off by R1; leakage current is typically less than the battery's self-discharge rate. Insertion of plug grounds base of Q1 through R2, which turns the emitter-collector path ON, powering the circuit. 3—Same technique powers LED in LH1061 dual electronic power switch; choose R1 to give at least 2 ma into chip for reliable switching. Example shows use in dual supply; each switch has maximum capacity of 200V @ 200 ma. 4—Additional options are implicit in the schematic of a jack equipped with a DPDT switch.

Extremely high OFF-resistance means negligible leakage. For boxes drawing less than 20 ma, there is no reason to use anything beyond common 3904/3906 types. Jobs switching higher current might move up to, say, a 2N2907 or 2N2219A in a TO-39 case.

Other electronic power switches include members of the thyristor family, such as silicon controlled rectifiers, diacs, triacs, etc., whose details lie outside the scope of the text.

Signal Switching

Because line-level signals tolerate up to several thousand ohms of series impedance, just about anything that provides a low-to-medium-impedance path can act as a signal switch. This includes the options mentioned so far, as well as those whose high resistance ill-suits power switching.

The 'stomp' switch is usually an effect/bypass switch. The DPDT arrangement is preferred because it takes the effect's circuitry completely out of the signal path. A SPDT switch simplifies wiring and lowers cost but leaves the axe connected to the effect's input all the time. While not a problem with properly designed input stages, this method leaves a needless channel by which spuriae can enter the signal path.

The pedal scene created a need for specific switches that themselves became pedals. An A/B selector box toggles between two inputs feeding one output. Effect-order switchers place two effects in series, but allow reversal of their order with single 4PDT switch.

Unless all points in the signal path exist at the same potential, switching yields a transient whose strength varies with speed of connection and magnitude of the voltage difference. Clicks are minimized by biasing contact points at the same potential or, if that is not possible, by making the connection over some tens of milliseconds, to give disparate potentials time to equalize, thereby avoiding a sonic impulse. Voltage differences commonly develop by the accumulation of charge in coupling capacitors. Bleeder resistors limit this charge. Another sticking point arises if input and output briefly meet, a common risk when bypassing an effect. In-phase gain causes feedback oscillation. The typical switch acts fast enough not to let the box howl; but a millisecond of feedback manifests as a pop. This problem is solved by using BBM switches.

Electronic switches suffer varying degrees of charge injection, a quantity of current coupled from the control path to the signal path. This current drops a voltage across the impedance in series with the switch terminals. The higher the impedance, the greater the voltage developed and the louder a click.

Discrete-transistor signal switches comes in series, shunt, and combination circuits. Though charge injection can be significant in such switches, they're
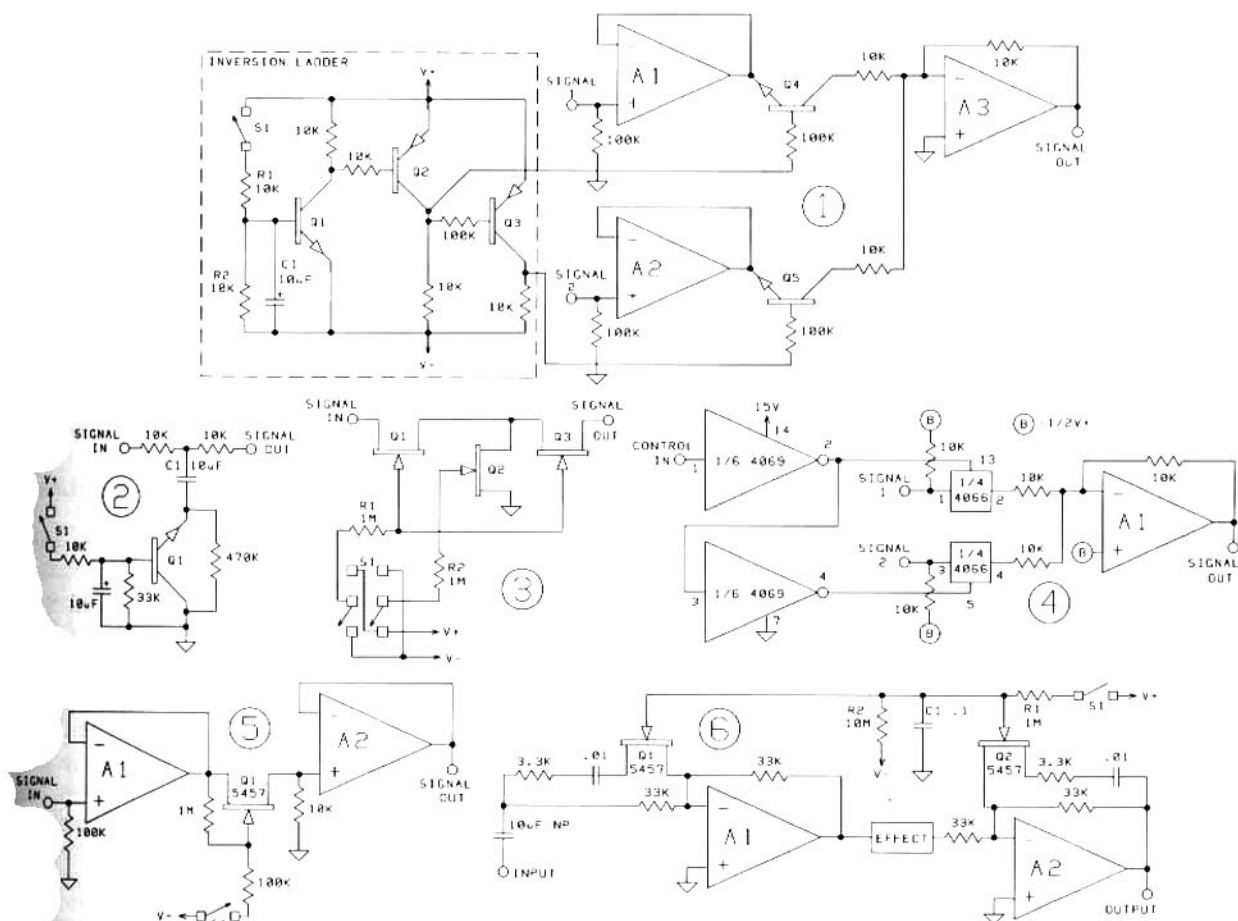
Fig. A11. 1—NPN transistors Q4/5 serve as signal paths controlled by logic inversion ladder Q1-3. When control input to R1 is HIGH, Q1 turns ON, bringing collector LOW, turning Q2 ON, bringing its collector HIGH, finally turning signal pass transistor Q4 ON. At the same time, the HIGH state of Q2's collector turns Q3 OFF, bringing its collector LOW, turning pass transistor Q5 OFF: Signal 1 is selected and appears at output of A3. When positive control voltage is released from R1, Q1 turns OFF from the unopposed ground potential applied through R2, and the states just described reverse; Signal 2 appears at A3 output. Time for C1 to charge and discharge slows the switching process, mitigating potential clicks. 2—BPT shunt switching method. When Q1 is OFF, it acts as a high impedance that does not materially affect signal traversing the two series 10K resistors. When Q1 turns ON, it acts as a low-impedance shunt to ground, effectively blocking the signal by divider action. Signal path is bidirectional. DC blocking cap C1 might or might not be needed, depending on the remainder of the signal path. This configuration applies to dual-supply systems, i.e., audio swings above and below ground. 3—Classical 'tee' switch using FETs. Q2 is OFF when Q1 & Q3 are ON, offering signal a low-impedance path through Q1 & Q3. When Q1 & Q3 are OFF, Q2 is ON, offering a near-complete shunt to ground. Isolation is so great that crosstalk depends more on circuit layout than switch leakage. 4—Electronic A/B signal switch using CMOS parts. When control input is HIGH, pin 2 is LOW, pin 4 is HIGH; so Signal 2 is selected. When control input goes LOW, switching state reverses, Signal 1 is selected. Key point is that this single-supply system uses $\frac{1}{2}$V+ as an artificial ground; switch inputs are biased to $\frac{1}{2}$V+ through 10K resistors, and switch output terminals are biased to $\frac{1}{2}$V+ as well, since inverting input of A1 is a virtual ground. Little voltage difference should be present between switch input and output terminals; resistors on switch inputs prevent accumulation of charge should coupling caps be needed on signal inputs. This switch has proven 'clickless' in author's pedals. 5—Single FET used as series switch. When S1 is open, Q1 is ON by ground bias from op amp output. Closing S1 pinches Q1's drain-source channel OFF, blocking signal. Switching can be slowed by a cap tied between gate and GND. A1 and A2 represent preceding and following stages, which do not have to be op amps. 6—Use of FETs as series elements to switch a pre-emphasis type noise reduction system in or out of the circuit using a single control input. When S1 is open, both gates are held LOW by R2, both FETs are pinched OFF, and emphasis/de-emphasis networks drop out of the circuit. When S1 closes, both FETs turn ON, giving low resistance relative to 3.3K. Time taken to charge/discharge C1 mitigates switching transients.

easily configured to switch slowly enough that no pop results. Transistors used for signal switching should be fully ON (saturated) or fully OFF. Marginal bias lets the audio voltage change the switching state, causing distortion.

Integrated-circuit switches come in JFET and CMOS types, commonly two or four to a chip. CMOS switches thrive in pedals because they're cheap and widely available, and run off as little as 5V while drawing scant current. The 4016 and 4066 exhibit nonlinearity in the form of ON-resistance that varies with signal level. The resultant distortion isn't audible in a typ-
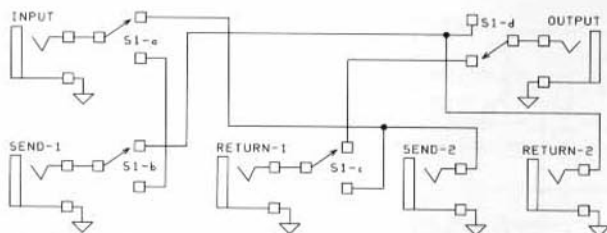
Fig. A12. Schematic of 4PDT switch wired as an effect-order changer. With switch in position shown, instrument input passes first to Send-2, comes back through Return-2, then passes out Send-1, comes back through Return-1, then passes to the output. When the switching positions change, order of effects reverses.

ical pedal, but can be reduced, if the builder wishes, by raising the external impedances seen by the switch. Divider action dilutes the nonlinearity, but the higher impedance acts at cross-purposes with click reduction.

JFET switches include older LF11331/13331 series, and the new SSM2042/2142 types built to be clickless by virtue of minuscule charge injection, and slow rise and fall times of the control pulse. The SSM switches are built for low noise and low distortion, and deliver greater headroom than CMOS switches. A built-in ramp function effects slower rise than fall, so break always precedes make. The SSM switches are clickless only in that the switch injects no audible transient. A click or pop may result if the points being switched exist at greatly different potentials.

Despite CMOS switches' 18V supply limit, they can

Fig. A14. Example of using LDRs as signal switches; combines shunt with series technique. When all LDRs are illuminated (low resistance), LDR1 & 2 act as shunts in path #2; LDR3 & 4 present a negligible resistance in path #1, so signal #1 predominates. When all LDRs are in darkness (>20 megohms each), path #2 is unimpeded, signal in path #1 sees practically an open circuit. Technique is found in amps made by Acoustic Control and Mesa Engineering.

Fig. A13. 1—LTC1043 A/B switch used as distortion multiplexer. Preamp output feeds comparator configured as zero-crossing detector, whose squarewave output (optionally passing through ÷2, ÷3, etc.) flips the 1043's signal path alternately between Distortion 1 and Distortion 2. Chip contains an oscillator that could be used to make switching rhythmic. 2—Example of using a 4066 to switch signal in a ±15V system. The 4066's supply must not exceed 18V; signal traversing it is limited to something less than 18V_{p-p}. Principle also applies to other electronic switches.

interface with op amps running off higher voltage; Fig. A13–2 illustrates the technique.

Electronic switches that open one path while closing another usually require an inversion ladder to provide simultaneous, reciprocal outputs from a single control input. The inversion ladder is generated by CMOS inverters, comparators, op amps, or discrete-transistor networks (Fig. A11–1). The ladder also gives a point to implement break-before-make, by controlling attack and decay of the output pulses. Attack (make) is made slower than decay (break), so break always precedes make.

LDRs' built-in time constants makes them 'soft' signal switches that have been used in a number of guitar amps (Fig. A14).

Hum

Hum happens. Shielded cable, humbucker pickups, balanced lines, ground-lift tabs, iso boxes and, lately, DSP engines tuned to quash hum all witness a constant struggle. Mixed measures also speak to sundry roots. No remedy has met 100% success, making an expectation of hum-free homebrew gear a tad optimistic. Still, nothing keeps the builder from creating pieces as resistant to hum as anything sold commercially.

Hum flows through copious modes, including:

- ► AC power fields
- ► supply hum
- ► ground loops

AC Power Fields

Hum originates in fact that electricity travels as alternating current, 60 Hz in America, 50 Hz in some other countries. Electromagnetic fields occur near power transformers and AC wiring. Magnetic and voltage components of the field couple by disparate mechanisms that dictate different countermeasures.

A magnetic field couples by means of induction. Movement of a conductor in a magnetic field, or movement of a field through a conductor, induces a voltage in the conductor. Induction equations reveal coupling efficiency to be proportional to magnetic permeability of the medium, orientation of field relative to conductor, and the number of turns of conductor experiencing the field. The tip-off to this mechanism is sensitivity to location and orientation of the axe, often with respect to an obvious source of 60-Hz radiation.

The field's voltage component couples capacitively, inducing a current in a conductor. The current drops a voltage across the conductor's impedance, and any impedance in series with it.

By these factors, a probe designed to sense hum could hardly better the single-coil pickup: thousands of turns of copper wire on a ferromagnetic core whose whole point is high permeability. Coupled with their high impedance, pickups ought to hum worse.

Shielding, as by a metal box, or a foil-lined electronics cavity in an axe, staunches mainly the voltage field, which cannot induce a current on the interior of a closed conductor. The magnetic field penetrates all but impractical thicknesses of iron or mu-metal. More effective anti-magnetic measures include reorienting the axe and putting space between the axe and the source of the magnetic field.

EM fields induce hum elsewhere in the signal path. Axe cable passing close to power cables pick up hum by magnetic induction. While stomp boxes don't usually contain spools of copper wire on an iron core, many do contain high-impedance nodes in the signal path. This leaves them prey to the AC voltage field. Say the field couples 100 nanoamps @ 60 Hz to the lead of a 100K resistor that happens to be in the signal path. That current drops 10 mv across the resistance, growing with each gain stage. The hum may be hard to ignore once it reaches the Jensens. This mechanism responds well to shielding, and is the reason stomp-box cases are made of metal.

Use of inductors or transformers in a stomp box introduces susceptibility to the magnetic component of the AC field.

Big filter-cap cans in the power supply may radiate enough energy to induce hum in nearby signal paths. Certain nonpolar electrolytic capacitors sport large aluminum cans that act as hum antennas when used as input coupling caps.

Supply Hum

Supply hum presents a triple threat. Most AC-powered gear uses a transformer, probably the commonest and strongest 60-Hz radiation source. Shielding the transformer blunts some of the magnetic field, much of the voltage field. Toroidal transformers confine their fields to the core, effectively shielding themselves. They see much use in high-end audio gear where self-shielding makes a selling point. Toroids tend to weigh less, for a given wattage, than conventional transformers, and to command a modest premium.

A second supply-related artifact manifests as buzz due to high-frequency pulses at 120 Hz. Rectification generates sharp transients at twice the line frequency. Even at 120 Hz, the folding points contain harmonics in the KHz. Shunting each rectifier with a cap helps but does not eliminate this problem (Fig. A15).

Finally, residual AC in the DC output, known as ripple, leads to hum if severe, or if the circuit suffers poor ripple rejection. Ripple rarely causes hum these days because modern audio circuits show a high degree of ripple rejection. Hum traced to ripple often indicates a faulty power supply.

Wall-mounted AC-to-DC stepdown converters sometimes give the user pot luck, for the manufacturer's idea of direct current may embrace filtered, regulated voltage as well as the raw output of a rectifier. The only way to tell what you've got is to examine the output on a scope. Even if excessive ripple is present, the circuitry lives sealed in plastic, impossible to open
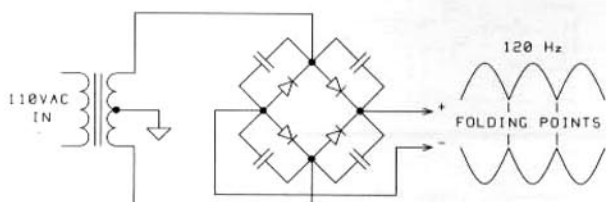
Fig. A15. Rectification in power supply creates high-frequency folding points at 120 Hz; buzzing sometimes bleeds into signal path. Caps in parallel with rectifier diodes help, but do not eliminate this problem.

without nullifying the UL® listing. The author advises against attempting to open or modify a wall wart. A sample tainted with ripple can sometimes be saved by adding a filter or a regulator to the circuit it powers (Fig. A16).

Grounding

Ground has several meanings. *Earth ground* refers to a neutral potential available at a cold water pipe or the third (round) terminal on the AC wall jack. It ties literally to the earth, a source/sink for electrons that can be considered locally infinite.

Circuit ground describes a reference point for positive and negative voltages that power a circuit. Because power supplies isolate their circuits from the AC line through a transformer, circuit grounds in separate devices can exist at any potential relative to each other and to earth ground.

Chassis ground refers to the metal chassis/enclo-

Fig. A17. 1—T1, D1, & C1 form a power supply typical of consumer audio gear. Schematic illustrates three types of ground. Note that, unless deliberately connected, the three grounds can exist at any potential relative to each other. Chassis ground is usually connected to circuit ground, but consumer audio gear is rarely earth-grounded. 2—Guitar amps in the 1950's tied circuit power ground to one side of the power line (selected by S1) through a 0.047µF 600V UL cap. Amp makers added the earth ground connection in the 1970's.
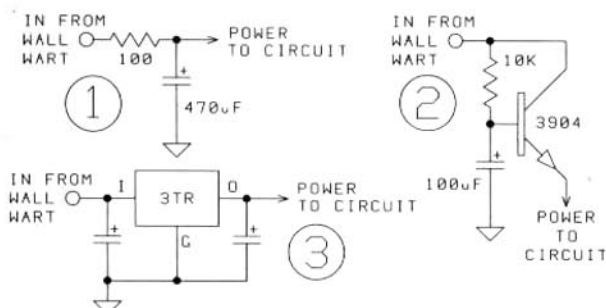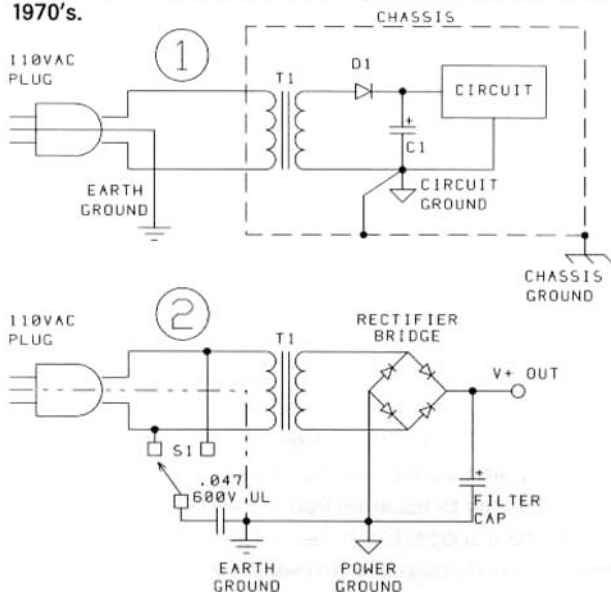
Fig. A16. Means to purge ripple from output of a DC wall wart. 1—Simple RC network cuts off around 3 Hz; ripple attenuation on the order of 25 dB. 2—Capacitor's decoupling action is multiplied by transistor's beta; voltage loses two diode drops across transistor. 3—Using standard 3-terminal regulator. For regulation to hold, input voltage must not fall below output plus regulator's dropout voltage; around 2.6V in the case of the older LM78XX and LM78LXX types.

sure that supports/houses a circuit. It usually ties to circuit ground at one point, and to earth ground if the circuit is earth-grounded. (See Fig. A17–1.)

These three grounds connect in various ways. Some arrangements hum, other don't, and the noisy ones don't always declare themselves before the fact.

The fortitude implied in the notion of 'grounding' spawns false intuition. A natural reaction to hum is to earth-ground the chassis, yet this may exacerbate the matter, or create hum where none was. Take the example of a VCR feeding the AUX input of a stereo receiver. Neither device is earth grounded; the two connected do not hum. If the receiver's chassis ground is then tied to earth ground, the combination may hum badly.

If earth-grounding cured hum, all consumer audio products would use three-prong plugs, which practically none of them do. Guitar amps came late to earth ground, and then out of safety, rather than hum. The polarity switch on vintage amps tied a high-voltage 0.047uF cap from circuit ground to either side of the AC line, selected through a SPDT switch (Fig. A17–2). That cap's 56K reactance at 60 Hz was low enough to reduce hum; but if the cap ever shorted, the full line voltage appeared on the chassis. Consumer audio gear uses a related approach to reduce hum, a resistance in the megohms tied from circuit ground to one side of the power line.

When using a stomp box to drive a home stereo amp, hum is usually reduced by tying the ground screw on the back of the amp to earth ground (the screw holding the cover plate on the wall socket; do not insert ANYTHING into the socket holes).

Ground Loops

Consider the knotty fact that two hum-free AC-powered pieces can connect by shielded cable and, voilà, they hum. The mechanism usually cited is the dread and fabled ground loop. Supposedly, no signal exists in the ground bus; but a signal—hum—may exist be-
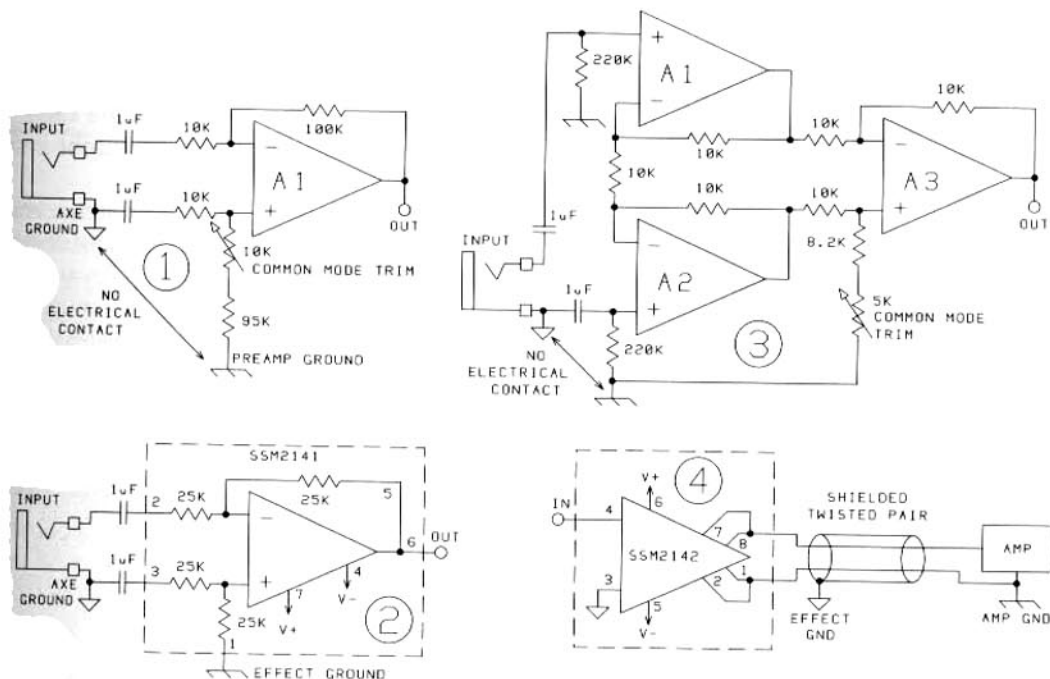
Fig. A18. 1—One means to isolate two ground systems is to treat an unbalanced source as if it were balanced. Differential amp rejects common-mode noise, but also isolates input ground system from amp ground system. This single-amp design suits line-level feeds due to high series resistance. 2—Similar circuit is available in a single IC, the SSM2141 differential line receiver; all parts inside dotted line are on chip. 3—Alternative differential amp allows high input impedance without the noise of series resistance. Net input impedance roughly equals source impedance. Common-mode rejection takes place in subtractor A3. 4—SSM2142 single-chip balanced line driver, here used to isolate effect ground from amp ground by treating amp's unbalanced input as if it were balanced.

tween the ground buses of two independently powered pieces.

The intuitive fix breaks the ground loop. This can be as easy as flipping a switch or unscrewing a wire on studio gear equipped with ground-lift. Gear not so equipped requires a cable whose shield lead has been disconnected from the XLR plug's ground tab at one end only; such a cable should be clearly marked. Try the cable in studio gear whose hum appears due to a ground loop.

For an unbalanced path between stomp box and amp, prepare a cable whose shield lead is disconnected from the plug at one end only. Mark this cable clearly. Unlike balanced lines, stomp-box ground must connect to amplifier circuit ground at some other point, or the setup won't work.

Intractable ground loops call for an isolation transformer. This measure applies to balanced and unbalanced lines. While a transformer breaks the loop, it introduces a latent hum source in that the transformer might as well be a pickup. If transformer isolation fails, the hum probably isn't due to a ground loop between the two isolated pieces.

Star Ground vs. Perimeter Ground
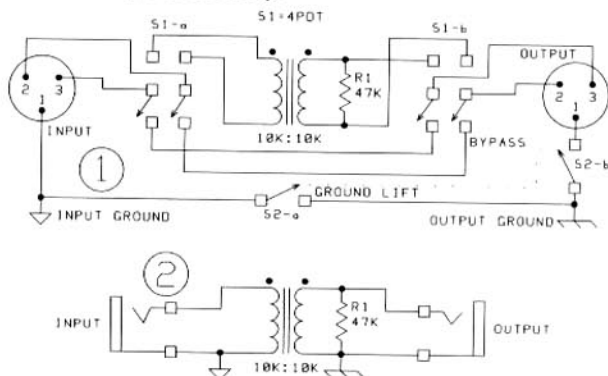
Ground loops can occur within an amp or a stomp box. Prevention of these loops popularized single-point ground, a wiring practice in which all grounded circuit points reach ground independently, avoiding a daisy-

chained ground bus. The resultant look of the ground point suggested the name 'star' ground.

Boards in this book use a perimeter ground because the power supply is assumed to consist of batteries, or a fully isolated supply, such as a DC–DC converter. Neither of these give 60 Hz entree to the system.

The builder looking to star ground to quash hum in a battery-powered stomp box is likely to be disappointed. Well shielded, battery-powered, star-ground gear still hums when connected to certain line-powered

Fig. A19. 1—Iso box schematic for balanced I/O. 4PDT switch bypasses transformer. A DPDT switch can easily be added to enable polarity reversal. Choose transformer impedance to suit the need. R1 loads secondary winding. 2—Iso box for unbalanced feeds, suitable for placement between stomp box and amp.

products.

The builder serious about preventing ground loops must recognize jacks mounted to the stomp-box housing as chief offenders, in part because the aluminum or steel case has high impedance, causing hum current to drop a higher voltage. Elimination of loops requires insulated jacks. The metal box ties to circuit ground at a single point.

Independent ground paths are vital to stability in certain types of solid-state power amps; they minimize noise in hybrid analog/digital circuits. The builder laying out a project meant to run directly off an AC wall wart might well choose single-point ground, because 60 Hz is present in the ground bus.

Input/Output Modes

Much of musical instrument systems' susceptibility to hum resides in the unbalanced connection mode, which naturally chains the ground bus, creating a ground loop. One way to prevent a ground loop is to couple the signal in balanced from. Differential stages need only a voltage difference between their input terminals to sense a signal. Whether their circuit grounds connect is irrelevant. The cable shield requires a connection to the ground of one system or the other, but not both.

Differential stages come in active and passive forms. A transformer is a passive form that completely isolates the two systems. Disadvantages include bulk, susceptibility to magnetically coupled hum, poor common-mode rejection ratio (~40 dB for a 42TL016), and limited choice of impedances.

The active form, a differential amplifier, offers much better common-mode rejection than ordinary transformers, but this doesn't necessarily translate into lower hum, because a perfectly balanced hum signal is rare. The differential amp isolates input ground from output ground enough to prevent ground-loop hum.

A balanced output stage is also useful for isolation, even when feeding an unbalanced stage (Fig. A18–4).

Electronic Noise

Electronic noise manifests as hiss, also called white noise. The builder must contend with three sources:

- ▸ thermal source noise
- ▸ amplifier input noise voltage
- ▸ amplifier input noise current

Thermal Source Noise

The *source* is the resistance of the axe, represented by R_s. Charge carriers bouncing around the humbucker generate electronic noise that stops only at absolute zero ($-459.7°F.$). This thermal noise voltage (E_{th}) is conveniently read off the graph shown below. E_{th} of selected resistances:

Ohms	Noise (nv)
10	.41
100	1.28
1K	4.1
10K	12.8
100K	41

Electronic noise voltage is expressed in nanovolts per root Hertz (nv/rtHz, or simply 'nv'), usually at 1 KHz. While these units initially seem abstract, they provide a means of quantifying noise, and they're the ones used in op amp data sheets.

You can find thermal noise of any resistance at room temperature by remembering that 10K=12.8 nv. Take the root of the ratio of the new resistance to 10K; multiply the result by 12.8. Thermal noise of a 5K single-coil pickup is:
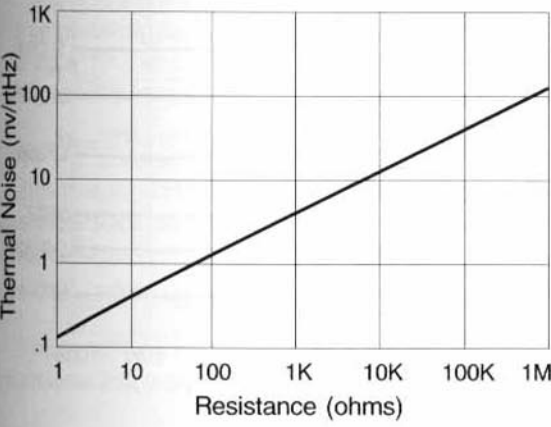
$$(5K \div 10K)^{0.5} \times 12.8 = 9.1 \text{ nv}$$

Thermal noise of a 100-ohm studio mic is:

$$(100 \div 10K)^{0.5} \times 12.8 = 1.28 \text{ nv}$$

Any resistance connected to an input acts as a source

Fig. A20. Graph of thermal noise vs. resistance.

of thermal noise.

Amplifier Input Noise Voltage

Like the source resistance, the preamp input stage generates noise by existing at room temperature. This amplifier input noise voltage (E_n) is stated in the data sheet, in nv/rtHz. It applies to op amps, preamp ICs, and discrete transistors.

Amplifier Input Noise Current

The preamp also generates input noise current, I_n, stated in picoamps per root Hertz (pa/rtHz). I_n matters because it drops a noise voltage across the source impedance, and any other impedance tied to an input. This noise in nv is given by:

$$(I_n \text{ in amps}) \times (\text{impedance in ohms}) \times 10^9$$

and is called "$I_n \times R_s$ noise."

The term 'R_s' has appeared twice, in contexts requiring careful distinction. When speaking of thermal noise, R_s must indicate resistance. When speaking of $I_n \times R_s$ noise, R_s can indicate resistance, reactance, or impedance, depending on the situation. For example, a 10K resistor tied to an op amp input has a resistance and an impedance of 10K ohms; this value is used to calculate thermal noise and $I_n \times R_s$ noise. But one winding of a 10K:10K transformer might show a resistance of 600 ohms; an impedance of 10K ohms. If that winding were connected to a preamp input, 600 ohms defines R_s for thermal noise, 10K ohms defines R_s for the $I_n \times R_s$ calculation.

An ideal capacitor has no resistance and thus makes no thermal noise. It exhibits reactance, which is subject to $I_n \times R_s$ noise. Real-world caps possess equivalent series resistance (ESR), usually low enough to ignore in noise calculations.

Complex RC and RLC networks that typify preamp input wiring require calculating the impedance of the network for $I_n \times R_s$ noise, while separating the resistance component for the thermal term. The picture is cleared somewhat by the observation that a typical pickup's resistance approximates its impedance closely enough to use resistance for both calculations. The text has adopted this convention.

Calculating Total Noise

We now have three noise quantities:

- ▸ thermal source noise voltage, E_{th}
- ▸ amplifier input noise voltage, E_n
- ▸ $I_n \times R_s$ noise voltage
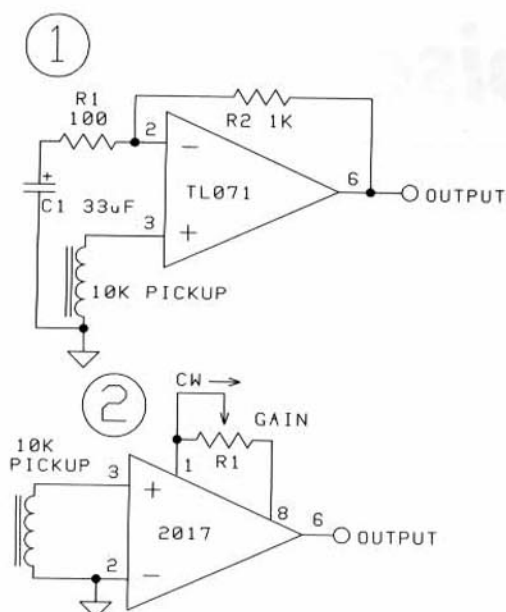
Fig. A21.

These quantities sum, but not by simple addition. Noise combines as a quadrature sum, the square root of the sum of the squares. Total noise at 1 KHz is given by:

$$[(E_{th})^2 + (E_n)^2 + (I_n \times R_s)^2]^{0.5}$$

To get total noise over a given bandwidth, multiply the quantity inside the radical by that bandwidth. This extra step is not needed to derive the relative noise index we'll be using.

Preamp Noise Evaluation

Choosing a chip or a transistor from its E_n alone is folly, because this might happen:

Ace Player built a preamp for his 10K humbucker using the TL071 op amp. Though it performs pretty well, he wants something quieter. Thumbing through a recent databook, he spots the SSM2017, an ultralow-noise studio-grade audio preamp. The chip can vary gain with a single pot, obviating the op amp's gain-setting components, and sports an E_n of 1 nv/rt Hz. This compares to 18 nv for the 071. Ace orders the chip, builds the preamp—only to find it no quieter than his current box.

Had Ace bothered with the calculation, he would have seen this:

For the SSM2017 (Fig. A8–2):

$E_n = 1$ nv
$E_{th} = 12.8$ nv
$I_n \times R_s = (2 \text{ pa} \times 10K) = 20$ nv

Total input noise at 1 KHz is:

$[(1)^2 + (12.8)^2 + (20)^2]^{0.5} = [1 + 163.8 + 400]^{0.5}$
$= 23.8$ nv

For the TL071 (Fig. A8–1):

$E_n = 18$ nv
E_{th} (for 10K) = 12.8 nv
E_{th} (for R1 = 100 ohms) = 1.28 nv
$I_n \times R_s$ (0.01 pa x 10K) = .1 nv
$I_n \times R_s$ (0.01 pa x 100 ohms) = .001 nv

Total input noise at 1 KHz is:

$[(18)^2 + (12.8)^2 + (1.28)^2 + (.1)^2 + (.001)^2]^{0.5}$
$= [324 + 163.8 + 1.6 + .01 + .000001]^{0.5}$
$= 22.1$ nv

Moral: Total noise is a complex function. E_n quoted in isolation may look impressive, but might or might not dominate total noise of the preamp connected to an axe.

Ace would have found the OP-27 a quieter choice. It and similar op amps (LT1007, MAX427) approach the minimum noise possible for a preamp working with a 10K source resistance. Input noise of the OP-27 in the 071 preamp circuit totals 13.8 nv, only a nanovolt more than the pickup's thermal noise.
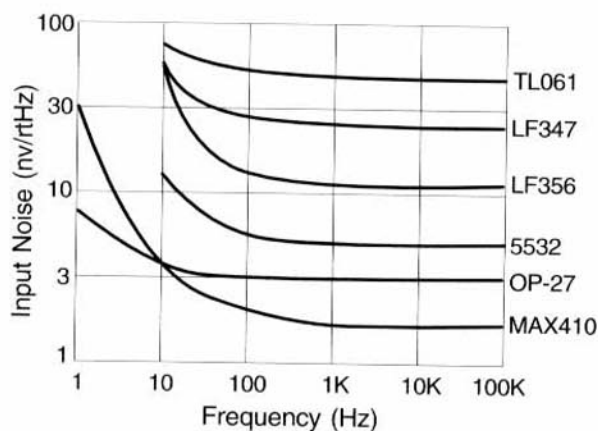
The 2017 makes a fantastic choice for a 100-ohm studio mic, because low source resistance means low thermal noise; low source impedance means low $I_n \times R_s$ noise; total noise below 2 nv.

Two minor points: (1) $I_n \times R_s$ noise is negligible for FET-input op amps. Don't bother to calculate it. (2) In strict terms, the net impedance for noise purposes at the 071's inverting input, is the impedance of [(R1 in series with C1) in parallel with R2]; but this comes so close to 100 ohms to make the difference moot. Use the value of the resistor on the inverting input for most noise calculations.

Noise in Discrete-Transistor Amps

These considerations apply to discrete-transistor pre-

Fig. A22. Graphs of E_n vs. frequency for selected op amps. As a normal trait, noise rises below 100 Hz; curves tend to flatten above 1 KHz. Curves approximated from manufacturers' data.
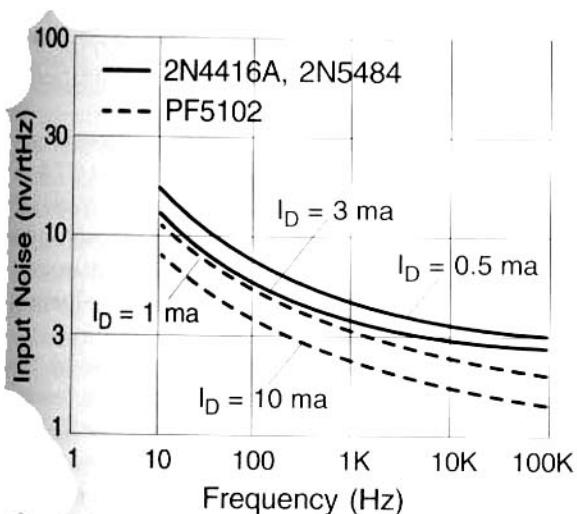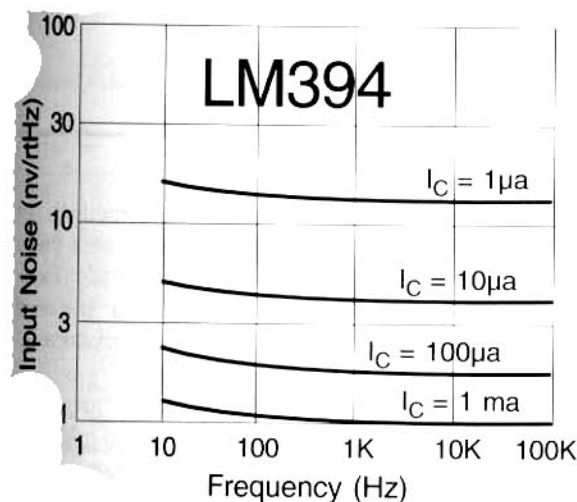
Fig. A23. Top graph plots E_n vs. frequency at selected values of I_C for the LM394 bipolar transistor. Bottom graph does the same for FETs and I_D. Note that FETs require comparatively higher current to achieve E_n as low as bipolar transistors'. FETs' advantage is very low I_n. Curves approximated from manufacturers' data.

amps, except that E_n and I_n change with transistor bias. E_n falls as the root of collector current (I_C), I_n rises as the root of I_C.

Premium low-noise transistors, such as SSM2210 and LM394, exhibit 1 nv E_n only when running high I_C, which also means high I_n. Generally, a bipolar transistor biased to show E_n of 1 nv means at least 1 pa of I_n. For common bipolar transistors working with a typical pickup, I_C (in microamps) at which minimum *total* input noise occurs is:

$$[(h_{FE})^{0.5} \div R_s] \times 0.026 \times 10^6$$

Assume an SSM2210, whose h_{FE} measures 700; axe resistance 5000 ohms. Minimum noise occurs when collector current equals

$$[(700)^{0.5} \div 5000] \times 0.026 \times 10^6 = 138\mu a$$

The plot of total input noise vs. I_C shows a minimum so shallow that a difference of $\pm75\%$ in I_C results in an insignificant change. An I_C of 100μa makes a generally satisfactory choice for guitar preamps.

Field-effect transistors follow a similar relationship between E_n and bias, but use drain current (I_D) instead of I_C. I_D should fall in the range 1–10 ma for lowest total input noise with a typical pickup.

The Big Picture

The preceding is necessary to put electronic noise in perspective. Analysis rarely requires detailed calculation because the major noise sources are obvious and often immutable. Key points to keep in mind:

▶ A guitar pickup's resistance seldom falls below 5K, meaning at least 9 nv of thermal noise. Catalogs are bursting with op amps having E_n lower than this, making pickup thermal noise the dominant noise source of the preamp. A high S/N ratio is still possible because most pickups generate high output relative to their thermal noise.

▶ If the preamp input uses a coupling capacitor, choose a relatively large value; 10μF works fine. A 0.01μF cap has 16K ohms of reactance at 1 KHz, which is subject to $I_n \times R_s$ noise.

▶ Always assess $I_n \times R_s$ noise, unless the op amp is an FET-input type. E_n and E_{th} dwarf $I_n \times R_s$ noise for types such as 074, 356, 347, etc. For bipolar op amps, each picoamp of I_n means 10 nv of noise per 10K of source impedance.

▶ Assess the *relative* contribution of noise sources by calculating incremental noise. For example, take a 6100-ohm pickup (10 nv E_{th}) relative to the E_n of an OP-27 op amp (3 nv). Quadrature sum of the pair is only 10.44 nv. Three nv looks significant next to 10 nv, but in this case is small enough to ignore.

▶ The best point to minimize electronic noise is the axe-preamp interface. Boost the signal as high as possible, as quietly as possible *in the preamp*.

▶ Because resistance alone adds noise, avoid unnecessary resistance in series with an input. While this usually means avoiding the inverting configuration in an op-amp preamp, most pickups generate such high output relative to their thermal noise that an inverting preamp contributes no audible noise, so long as the input resistance is kept below, say, 220K ohms, and so long as the op amp is an FET-input type.

▶ Noise analysis is normally applied only to the preamp. While succeeding stages also generate noise, they're assumed to operate at line level, which isn't always the case, especially in a stomp-box daisy chain. Noise gates should remain handy for the immediate future.

Schottky Noise

Schottky or 'shot' or 'popcorn' noise is not reflected in noise equations. It manifests as pops and crackles, rather than hiss. If shot noise is audible, the offending part is deemed defective and replaced.

Preamp Design

The preamp is the stage that first boosts the axe feed, a key nexus, because noise introduced here will grow in subsequent stages. Unless the builder means for distortion to happen as the first step, he should make the preamp as clean and quiet as possible.

This series of stomp boxes chose an op-amp preamp for many reasons. Op amps cost negligibly more than first-quality transistors, and their inter-unit uniformity eliminates variables that haunt discrete-transistor designs. Op-amp circuits shift less with changes in supply voltage than do discrete-transistor circuits, and they're easy to configure for special needs without departing from the basic circuit. Op amps possess extremely high input impedance, coupled with output impedance low enough to drive most loads found on stage or in studios. Finally, op amps oblige inverting, noninverting, and differential input configurations, each offering special advantages in the pedal realm.

The noninverting preamp (Fig. A24–1) minimizes noise by minimizing the resistance at both inputs, in turn minimizing thermal source noise and $I_n \times R_s$ noise. R3 sets the input impedance at 150K, but this value is shunted by the source impedance, making the net impedance slightly less than the axe impedance. By contrast, an inverting preamp uses brute-force series resistance to attain the 150K impedance, adding more than 49 nv of thermal source noise (Fig. A24–2). A bipolar op amp's I_n drops a significant noise voltage across this resistance. The 0.7 pa I_n of an LM833 times 150,000 ohms is 105 nv; the 10K pickup's normally dominant 12.8 nv looks puny by comparison. Also, the added series resistance causes hum current sensed by the pickup to drop additional voltage. (The inverting amp has a couple of redeeming features. The big series resistance bulletproofs the op amp. Also, the user can vary preamp gain down to zero by

using a pot for the feedback resistor, something not possible with a noninverting preamp. Unless these factors outweigh others, avoid the configuration for instrument-level inputs. It works great for line-level stages, but even there, use the smallest practical input resistance to keep noise low.)
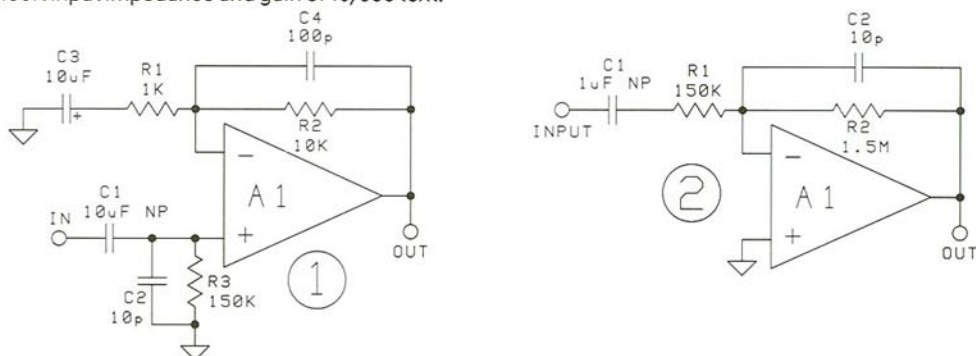
The preamp includes C2 to limit high-frequency response at the input, because local RF can make the chip an AM radio. C2's f_{-3} with a 10K pickup is about 1.6 MHz. In areas cursed with high ambient RF, the builder can raise C2 as high as 220pF without affecting the sound of many pickups. Rectification of RF is less problematic in FET-input op amps than in bipolar types.

Veteran op-amp users routinely compensate the amp for stray capacitance, which is the point of C4. Its f_{-3} with the feedback resistor does not fall below 159 KHz. While neither C2 nor C4 get close enough to the audio band to "suck tone," C2 should be kept as small as practical, because it forms a 12 dB/octave lowpass filter with the pickup's inductance.

Low-frequency response is determined by two poles. First, R1 and C3 corner at about 16 Hz, leaving 7-string basses all the bottom they need. The 1K value of R1 is a compromise. If R1 were reduced to 100 ohms, C1 would have to rise to 100μF to keep the same low-frequency corner. The greater noise contribution of 1K vs. 100 ohms is inaudible at typical preamp gain levels.

The second low-frequency pole is formed by C1 with the source impedance and R3. Assuming R_s to be 10K, this pole cuts off around 1.6 Hz. The builder wishing to weaken low-frequency response should decrease the value of C3, not the value of C1. A value of C1 low enough to attenuate bass response coincides with reactance high enough to add significant $I_n \times R_s$ noise to a bipolar-input op amp. While C3 is also in se-

Fig. A24. 1—The standard preamp chosen for projects in this book; adapts to many special needs; see text. 2—An inverting preamp w/150K input impedance and gain of 10; see text.
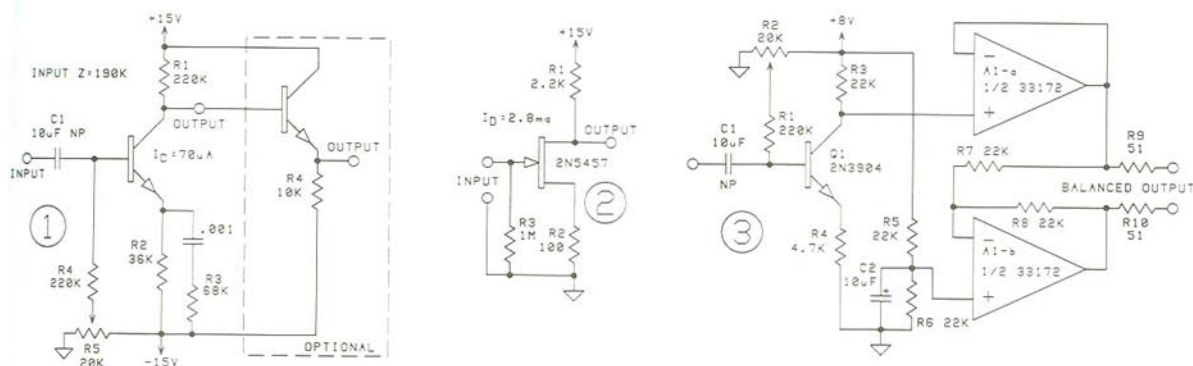
Fig A25. Three of countless alternatives to an op-amp preamp. 1—Bipolar transistor is biased at about 70 μa collector current; close to lowest total noise for 10K impedance. Holds for 2N3904 & similar; much intersample variation occurs among common transistors; gain ~5. Trim R5 to give maximum headroom. Input impedance ~190K. High output impedance usually means adding a buffer. 2—Simple FET preamp; gain = 5; drain current = 2.8 ma, in keeping with need for higher current to minimize FETs' input noise voltage. Clean, low distortion up to ~2V$_{p-p}$ input; exhibits tube-like squashing before hard clipping occurs. 3—A hybrid, coupling a discrete-transistor preamp to a couple of op amps that convert the output to balanced from, gaining 6 dB in the process. Relatively high noise of MC33172 op amps is less problematic, because preamplification takes place in the transistor.

ries with an input, it's a gain-determining element, where C1 is not. C3's reactance rises with falling frequency, but this rise also reduces gain for the noise resulting from the increased reactance, resulting in no audible rise in noise. Reducing the value of C1 leads to an uncompensated rise in reactance, and is the noisiest way to roll off the low end.

Input coupling cap C1 is nonpolar, anticipating the rare event of a DC offset present in the axe feed. Setups that never face this problem can use an ordinary polarized aluminum electrolytic, the '−' lead oriented toward the axe output.

DC coupling an axe to this preamp is practical so long as the axe contains no coupling capacitor and no DC offset. Direct coupling obviates R3, because the amp bias current flows through the axe.

Most preamps, but direct-coupled preamps especially, should use a shorting jack for the input, to prevent instability when the axe cord is unplugged.

The value of R3 sets the preamp's input impedance; 150K does not load the presumed 10K pickup. R3 can be increased to 220K or more, but this may cause problems in op amps having high input bias current or high offset current. FET-input op amps readily tolerate a 1-megohm value of R3.

Gain of a noninverting amp = [1 + (R2÷R1)]. Setting preamp gain requires knowledge of the output of the axe and the supply voltage at which the preamp will run, for these variables define headroom, the maximum signal voltage the preamp can support without clipping. If the axe output peaks at 1V, and if the amp is running on ±6V, voltage gain of about 10 (20 dB) should be the limit, to allow for the fact that common op amps do not swing rail to rail. The gain limit rises in proportion to supply voltage.

Choosing Op Amps
Op amps come in a kaleidoscope of flavors that tends to confuse the inexperienced builder. Choice calls for balance among noise, supply current, cost, and packaging. An op amp used as a preamp should seek minimum noise. Assuming a 10K average pickup impedance, minimum noise occurs in types similar to the OP-27, such as NE5534A, LT1007, and MAX427. The old stand-bys give best value. An OP-27 is second sourced for <$1.50; an NE5534 for about $0.50.

Maximum economy comes with quad packages, which also facilitate compact boards.

Op amps in line-level stages following the preamp do not have to be low-noise types, but circuit design should not ignore the factors affecting noise. That generally means using the smallest interstage resistance possible. The venerable TL07X series (18 nv/0.01 pa) makes a good choice, drawing about 1.4 ma per op amp. Most players will hear the noise difference between 07X types and the low-power 06X series (47 nv/0.01 pa).

Current drain becomes an issue in battery-powered boxes. The builder can trade noise performance for longer battery life. Standard low-power series include TL06X, LF44X (35 nv/0.01 pa), and MC3317X (32 nv/0.2 pa). Each op amp draws about 200 μa. Current drain usually measures higher than this with the amp in a circuit, because supply current stated in the data sheet does not account for current sourced through the op-amp output stage.

True micropower op amps make an increasingly viable option. Micropower means less than 100 μa per op amp; <50 μa is not unusual. This option demands careful reading of the data sheet, since many micropower amps suffer very high noise, suiting them only to control paths; or very slow speed, meaning insufficient gain for some applications.

Output voltage swing bears directly on headroom. Some high-performance types will not swing within 2V of either rail; others will swing within one diode
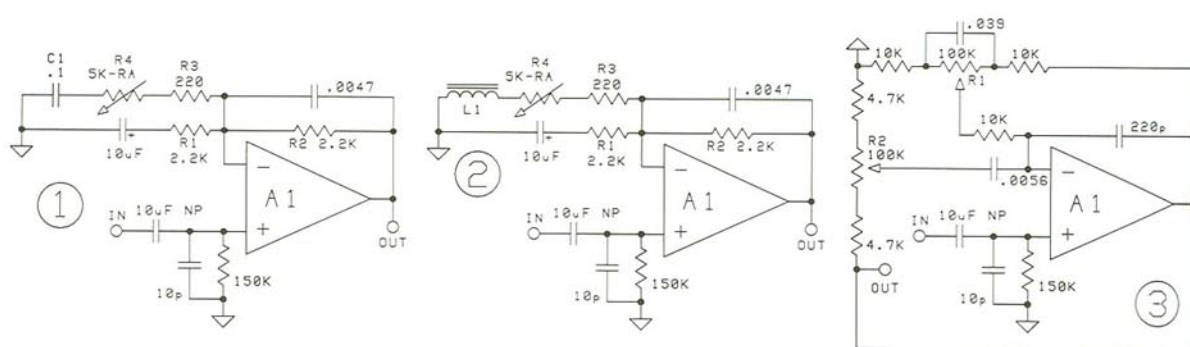
Fig. A26. Circuits that blur the line between preamp and tone control. 1—Noninverting preamp has gain of ~2 with R4 at maximum. As R4 is reduced, up to 20 dB of treble boost kicks in. 2—Identical to circuit #1, but an inductor, or a simulated inductance, has been substituted for capacitor. In this case, up to 20 dB of bass boost is available; f_{-3} depends on value of inductance. 3—By grounding the 'input' of standard active tone circuit, and using noninverting input of op amp as preamp input, tone control integrates with preamp. With R1 and R2 fully CCW, gain approximates 1. Gain rises as bass and treble controls are rotated clockwise. Resultant tone curve resembles that of the Fender 'ladder' circuit. Advantages of integrating tone control into preamp include simplicity, reduced cost, and less noise.

drop. Many CMOS types swing rail to rail.

Output drive matters if the output sees an impedance less than 1000 ohms; 5532/34 and LM837 amps can drive 600-ohm lines.

A preamp is usually a flat gain stage; tone controls act as line-level frequency shapers. The differences turn out to be purely semantic, for nothing prevents the builder from configuring the preamp to vary gain according to frequency (Fig. A26).

Alternatives to Op-Amp Preamps

As a pure gain block, a bipolar transistor is hard to beat, being simple, versatile, and cheap. Except in premium matched transistor pairs, current gain and noise vary greatly from sample to sample. Yet, bad samples excepted, the user probably won't hear a difference at gain levels below 20 dB.

The common-emitter amp (Fig. A25–1) makes a practical approach to an axe preamp. The high collector load needed to achieve I_C in the vicinity of 100 μa means high output impedance, often requiring a buffer, such as an emitter follower.

A discrete-FET preamp is even simpler and allows direct coupling to the axe, but requires many times the current for comparably low noise (Fig. A25–2). When overdriven, this stage tends to squash the signal before it clips.

Both single-transistor amps give significantly less headroom than an op-amp preamp at the same voltage. This disadvantage wanes as supply voltage rises.

Hybrid preamps couple a transistor to an op amp, combining the simplicity of a single transistor with the output drive of an op amp (Fig. A25–3). They also enable very low-power designs, the transistor preamp drawing ~100 μa; each low-power op amp, <200 μa. The higher noise of low-power op amps is less noticeable because the transistor supplies the initial gain.

Low-noise preamp chips, such as SSM2016 and SSM2017; and super-low-noise op amps, such as LT1028, best suit source impedances less than 1000 ohms. They obtain low E_n at the cost of high I_n, which drops significant noise voltage across the high axe impedance. These chips excel with 150-ohm microphones, but combine with typical pickups to yield noise well in excess of the possible minimum.

Preamps for Piezo Pickups

Piezoelectric pickups not only work differently from magnetic pickups, they possess significantly higher impedance. Avoidance of loading demands a very high-impedance input stage. By the noise factors previously discussed, discrete-FET preamps and FET-input op amps give least total noise with source impedances higher than ~25K ohms. FET stages also lend themselves to high-impedance configurations. The tone-robbing effect of axe cable capacitance becomes significant at these impedances, making active electronics an attractive option.

Piezo pickups do not present a DC path, allowing them to couple directly to the preamp.

Tone Control

The broad sense of tone involves factors beyond frequency response, notably, the character of an amp's distortion and the sound of the pickups and strings. This chapter limits the scope of tone to frequency response. Taking control of tone is easy once the builder understands frequency-selective gain. That, in turn, hinges upon a property called *reactance*, usefully modeled as frequency-dependent resistance. Capacitors exhibit reactance, defined by:

$$X_C = 1 \div (6.28fC)$$

where X_C is reactance in ohms, f is frequency in Hz, C is capacitance in farads. Capacitive reactance falls as frequency rises. A $1\mu F$ cap acts as a 7962-ohm resistor at 20 Hz; a 79.6-ohm resistor at 2000 Hz.

Inductive reactance is given by:

$$X_L = 6.28fL$$

where X_L is reactance in ohms, f is frequency in Hz, L is inductance in henries. Inductive reactance rises with frequency; in a sense the opposite of capacitive reactance. A 1-henry inductor acts as a 126-ohm resistor at 20 Hz; a 12,600-ohm resistor at 2000 Hz.

Resistive voltage dividers affect all frequencies equally. Dividers containing caps or inductors subject signals to frequency-dependent division. The simplest example of this is the tone control found in most guitars, a variable treble-cut circuit consisting of a capacitor and a pot (Fig. A27–1). Though rarely used, inductor-based bass cut is also possible.

Combinations of inductors and capacitors tap *resonance*, given by:

$$f = 1 \div (6.28 \times [LC]^{1/2})$$

where f is the resonant frequency in Hz, L is inductance in henries, C is capacitance in farads. Rearranging to solve for C and L:

$$C \text{ (in farads)} = (1 \div [39.4 \times f^2 L])$$

$$L \text{ (in henries)} = (1 \div [39.4 \times f^2 C])$$

In series resonance, the impedance of the network falls theoretically to zero. In parallel resonance, impedance rises theoretically to infinity. Imperfections in

Tone Control Options

Passive
  treble cut
  bass cut
  notch
  relative peak
  passive bass & treble (PBT)
    conventional (cut + quasi-boost)
    Fender 'ladder'
    Supro (12 dB shoulder)
  bright switch
  deep switch
  phase-cancellation notch
Active Bass & Treble
  true boost/cut based on PBT combined
    with transistors, op amps, or triodes
Equalization
  gain block type
    single transistor
    single triode
    discrete-transistor op amp
    discrete-triode op amp
    IC op amp
  bandpass filter type
    graphic equalizer
      inductor-capacitor
      simulated inductance + capacitor
      active filter
    parametric equalizer
      state variable filter

Fig. A27. 1—Simplest voltage divider tone circuit. Because R2 is large relative to R1, cap has no significant effect when R2 is maximum. When R2 is 0, frequency-dependent divider action causes treble to roll off above the frequency at which C1's reactance equals R1; ~1592 Hz in this case. 2—Inductor-based circuit acts similarly, but cuts bass. 3—Parallel LC network gives high impedance at resonance, low impedance above and below resonance. Divider action gives a relative peak. 4—Series LC network gives low impedance at resonance; divider action results in a notch.

115
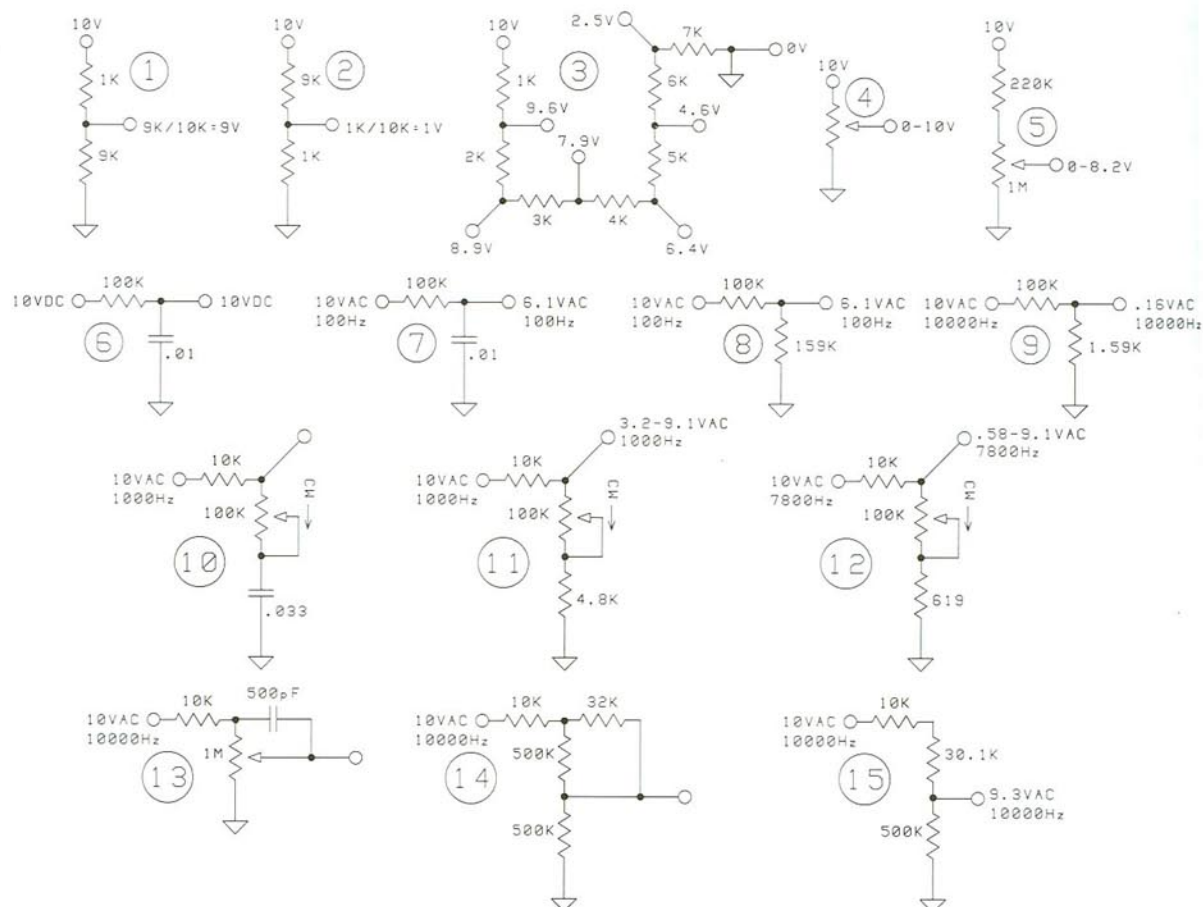
Fig A28. Understanding voltage dividers. 1—Two resistors in series, one end tied to 10V, the other to ground. Divider action dictates voltage at their junction. Sum all resistances; sum all resistances between the measurement point and ground; divide the latter by the former, multiply the result by the supply voltage; so (9K÷10K) × 10V = 9V. 2—Resistors have swapped places, so (1K÷10K) × 10V = 1V. 3—Analyze any number of series resistors the same way. 4—One of the commonest voltage dividers, a potentiometer or 'pot.' The middle terminal is called the wiper; voltage available at that terminal varies from 0–10V, depending on setting of pot. 5—Two dividers, one fixed, one variable. 6—10VDC applied to one end of 100K resistor appears as 10VDC at the other end, because 0.01 μF cap does not conduct DC and thus has no effect. 7—If 10VAC at 100 Hz is applied to input of 100K resistor, only 6.1VAC appears at the output. 8—Explains why; capacitor does not conduct DC, but does conduct AC. Its reactance (X_C), changes with frequency according to $X_C = 1/6.28fC$, where X_C is capacitive reactance in ohms, f is frequency in Hz, and C is capacitance in farads. Substituting X_C of .01μF @ 100 Hz shows divider action. 9—If frequency rises to 10 KHz, capacitive reactance falls, changing divider action. At this frequency 10VAC is reduced to 0.16VAC by divider action. 10—A variable divider consisting of two resistances and one capacitance. To analyze behavior, substitute X_C at desired frequencies. 11—At 1 KHz, the 100K pot varies output over 3.2 to 9.1VAC. 12—At 7800 Hz, output varies 0.58 to 9.1VAC. The 10K resistor could represent a humbucker; the 100K pot and .033 cap could be a network typical of in-the-axe tone controls. 13—Variable divider shunted by 500pF capacitor. At 10 KHz, X_C = 32K (Figs. 14, 15). When 1M pot is fully CW, cap is shunted and has no effect; if pot is fully CCW, wiper is grounded and no signal gets through. But if wiper is centered, signal output at 10 KHz equals roughly 93%, because the capacitor gives a low-impedance path around the resistive divider. 500pF cap has ~3.2M of reactance at 100 Hz; network's output at 100 Hz is ~50%, as dictated by the purely resistive portion of the divider. This action characterizes bright switches found in many amps.

real-world parts prevent both extremes. Used in dividers, series resonance gives a variable notch; parallel resonance gives a relative peak (Fig. A27–3, 4). Caps, inductors, and resonant networks can be used countless ways to change tone. For example, a pickup is also an inductor, which resonates with series or parallel capacitance (Ref. 38).

These few principles underlie a wealth of tone circuits in axes, pedals, and amps.

Tone Controls in Guitar Amps

Limitations of the cut-only approach bred passive bass & treble (PBT). Symmetrical forms reduce the signal a set amount, say, 20 dB. In 'boost' positions, they let some of the cut bass or treble through (Fig. A29). The 'cut' positions further reduce bass or treble. PBT's great losses demand gain before and often after the tone block.

Perhaps the commonest alternative to symmetri-

TREBLE-20KHz

(1)
C1 .002
IN
R2 82K
R1 50K-A
.9
OUT
.1
R3 8.2K
C2 .02

(2)
IN
4K
82K
52K
OUT
8.2K
400

(3)
IN
4K
45K
5K
400
OUT: .1 IN (-20dB)

(4)
IN
4K
OUT: .93 IN (-.7dB)
45K
5K
400

(5)
IN
4K
45K
5K
400
OUT: .2 IN (-42dB)

BASS-20Hz

(6)
IN
R4 5.6K
C3 .056
R5 50K-A
.9
OUT
.1
C4 .56
R6 560

(7)
IN
5.6K
45K
5K
560
OUT: .1 IN (-22dB)

(8)
IN
5.6K
OUT: .9 IN (-2.9dB)
45K
5K
560

(9)
IN
5.6K
45K
5K
560
OUT: .2 IN (-40dB)

BOTH

(10)
IN
220K
.001
R7
1M-A
.01
OUT

(11)
IN
470p
220K
.001
1M-A
.22K
1M-A
22K
.01
OUT
.0047

Graphs

input to treble network

output @ maximum treble 'boost'

'treble control point' = f2

output w/'flat' treble setting

output @ maximum treble cut

input to bass network

output @ maximum bass 'boost'

'bass control point' = f1

output w/'flat' bass setting

output @ maximum bass cut

0 dB, -20 dB, -40 dB

Frequency (Hz): 20, 200, 2000, 20K

$$C1 = \frac{1}{6.28\ f2\ R2}$$

$$C2 = \frac{1}{6.28\ f2\ R3}$$

$$C3 = \frac{1}{6.28\ f1\ R4}$$

$$C4 = \frac{1}{6.28\ f1\ R6}$$

rig. A29. Understanding passive bass & treble. 1—Passive treble control; R1 must have an audio taper to give 'flat' response when centered. 2—Substitution of capacitive reactances at 20 KHz for cap values. 3—Because R2 & R3 are very large relative to X_C of C1 & C2, respectively, the analysis can ignore R2 & R3. 4—With treble control fully CW, 93% of signal at 20 KHz gets through. 5—With treble control fully CCW, only 1% of 20 KHz gets through, an attenuation of 40 dB. The network has negligible effect on bass due to caps' enormous reactances at low frequencies; e.g., X_C of C1 at 20 Hz is 4 megohms. 6—Typical passive bass control. 7—Capacitive reactances of C3 & C4 at 20 Hz are so high relative to R4 & R6, respectively, to ignore caps in divider analysis. Their reactances at treble frequencies are low, providing a shunt for treble around the network, leaving treble largely unaffected by the divider setting. 8—With bass control fully CW, 90% of signal gets through. 9—With bass control fully CCW, only 1% of signal gets through. 10—Combining bass and treble controls requires the addition of R7 to reduce interaction between the networks. Assuming both pots to be equal in value, make R7 ~⅕ pot resistance; in this case, 10K. With these resistances, bass & treble control points can be shifted by changing cap values, keeping ratio of the two at 10 to retain symmetrical response; or deviating to skew the response. To raise impedance of network, scale all resistors up by the same percentage, scale capacitors down by the same percentage; and vice versa to scale impedance down. 11—PBT circuit found in many Ampeg amps; treble circuit omits two resistors. Graphs illustrate typical passive tone control response curves. Output is –20 dB with controls centered; 'boost' merely lets through some of the reduced signal.

Fig. A30. Evolution of the Fender 'ladder' tone circuit. 1—Early version of ladder; all ladder circuits take output off the wiper of the treble control pot. 2—Circuit whose 350K treble pot is tapped at 70K, allowing greater treble cut. 3—Perhaps the commonest configuration in vintage Fender amps. 4 & 5—'Mid' control does not actually control midrange, but modifies divider action of bass & treble pots. 6—Ladder tone control from Hiwatt amp; different from Fender's, yet of patent lineage. 7—Graph shows I/O of ladder from Deluxe Reverb, relative to constant 10V$_{p-p}$ input. The Fender tone curve is characterized by a wide, lower-mid dip approaching 20 dB.

cal PBT is Fender's ladder-like circuit (Fig. A30). Variations on this design are found in Rickenbacker, Marshall, Hiwatt, and many other amps. The network produces a curve with a deep notch near 300 Hz. Many amps add a control labeled 'midrange' that actually varies the divider action of the bass & treble controls. The ladder circuit underwent many changes in RC values, a few in configuration. Fender briefly made

amps with a tapped treble pot, apparently abandoned because improved treble cut had few takers.

Beyond PBT and Leo's ladder lies a hodgepodge of dividers that never gained much popularity, but which give the builder interesting and useful alternatives (Fig. A34). Most passive tone curves slope away from bass and treble corners at 6 dB/octave. Supro amps and Gretsch's Chet Atkins amp incorporated passive

Fig. A30B. 1—From preamp stage of Ampeg GV22; amp bills S1 as 'ultra-lo' switch; in fact, it creates severe treble cut by divider action. S2 is the 'ultra-high' switch, known as 'bright switch' in Fender amps; gives high frequencies a low-impedance path around the 1M volume control pot. 2—R1 is the 'presence' control typical of several Fender amps; as value of R1 is reduced, C1 shunts high feedback frequencies to ground, preventing negative feedback of treble.

Fig. A31. Basics of active bass & treble. 1—Pure bass circuit varies the electrical location of parallel RC network R3-C1. When R3 is fully CW (2), network is in A1's feedback loop. DC gain is 110K÷10K, but AC signals see a progressively lower impedance as frequency rises, due to falling reactance of C1, such that impedance of 100K in parallel with C1 becomes negligible, giving gain for treble frequencies of 10K÷10K=1. 3—When R1 is fully CCW, the RC network appears as an input impedance to A1. DC gain is 10K÷110K, but AC signals see a progressively lower impedance as frequency rises, such that R3-C1 impedance becomes negligible at and above 1 KHz, and gain approaches unity. 4—Pure active treble control differs in several respects. First, A1 needs a DC path in the feedback loop for biasing; R1 provides this. Second, bass frequencies cannot get through the treble circuit due to C1's high reactance at bass frequencies. R1-R2 provide a unity-gain path for bass frequencies around the treble circuit. When R4 is CW, C1 acts mainly as the input impedance to A1. This shows treble frequencies a low impedance, bass frequencies high impedance. When R4 is CCW, C1 acts mainly as a feedback impedance, reducing impedance for treble frequencies but not bass, which flows through R2. C3 is usually needed because A1's gain reaches into the AM radio band, causing interference or instability. Circuit uses 4.7K resistors instead of 10K of the bass version due to halving effect of 100K resistors with each leg of circuit. 6—Combination active bass & treble using single op amp. Because bass network provides a bass path and DC bias for A1, the 100K resistors of the pure treble network are no longer needed. Bass control affects treble, especially above 10 KHz; this interaction is minimized by addition of R1, which has little effect on bass response. Important to keep in mind that circuits 1–6 invert. 7—An example of Baxandall-type bass & treble control network. Circuit inside dotted line is recognizable as PBT circuit, but uses linear-taper pots and symmetrical component values. Output of network is taken off juncture of pots; feeds amplifier Q2, which inverts; signal couples to a second amp, Q3, through the emitter. The still-inverted signal is buffered by AC emitter follower Q4. A portion of signal is taken at 2.2K/220 resistor junction; feeds back to point of network that normally ties to ground. Connection from 220/2.2K resistor junction to PBT network is a negative feedback path; if this path is broken, Q2-4 provide gain of ~3.5. When loop is closed, gain approximates 1 with bass & treble controls centered. When taken off center, they alter the amount of feedback for bass and treble frequencies, and thus the gain for those frequencies.

second-order networks, doubling the slope.

Early PBT networks adapted to gain blocks came to be known as Baxandall tone controls, but these rarely surfaced in guitar amps

Many amps provide for tone control beyond PBT.

Bright switches select a cap in parallel with the volume pot, to give high frequencies a low-impedance path around the pot. *Deep switches* emphasize bass by passively cutting treble, or by giving bass a low-impedance path around a high-impedance divider. Most
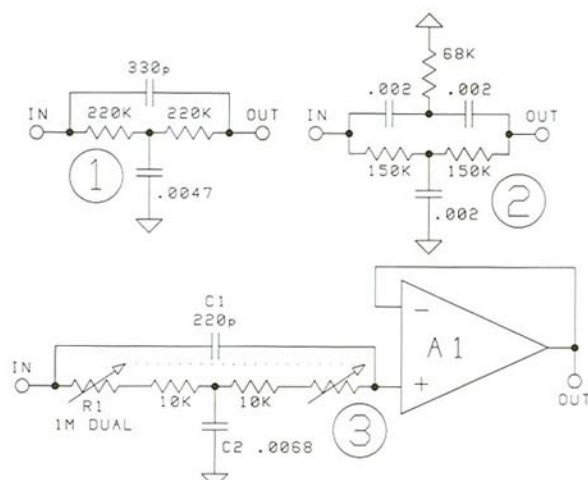
Fig. A32. 1—Phase-based 600-Hz RC notch found in Gibson GA-20RVT. 2—Network whose response is similar to that of the Gibson circuit, from the original Ampeg SVT. 3—Derivative circuit makes a simple tunable notch; component values shown vary range 140–14,000 Hz; change relative values of 1M dual pot and 10K series resistors to narrow the control range. Notch depth holds close to ~24 dB over the tuning range. To decrease notch depth, increase ratio of C1:C2; to increase depth, decrease the ratio. Changing C2 from 0.0068μF to 0.0015μF decreases notch depth to ~12 dB. Changing capacitor ratio also shifts frequency range. Scale frequency without changing notch depth and three-decade control range by changing C1 and C2 by the same percentage. For instance, to cover 70–7000 Hz, double values of both caps. This circuit is easily tuned by ear, by swapping out capacitors on the breadboard while auditioning changes in tone.
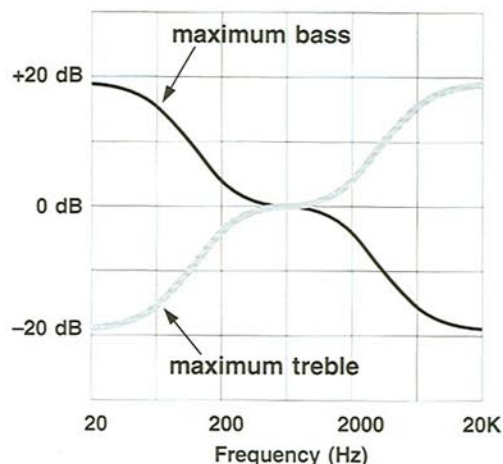
Fig. A33. Curves typical of ganged bass/treble wired such that bass boost also causes treble cut, and the reverse. Appearance of curves may account for the name 'tone-X.'

cause they manipulate the less obvious quantity of phase. Vintage Gibson and Ampeg amps used passive RC networks to produce a notch near 600 Hz (Fig. A32). Highpass and lowpass networks shift phase in opposite directions. Wired in parallel, the phase disparity at their output approaches 180° for one frequency, at which point partial cancellation occurs. The degree of cancellation depends on amplitude match at the 180-degree point. A 600-Hz notch heightens clarity when applied to guitar, or invokes an illusion of depth when applied to bass.

Active Tone Control

Where passive tone circuits harness reactance in simple dividers, active tone circuits use reactance to alter gain of op amps, transistors, or tubes. Boost/cut starts from unity gain, and allows use of symmetrical component values and linear-taper pots.

presence controls act in the phase splitter/power stage by reducing negative feedback above 5 KHz; treble emphasis at this stage sounds distinct from that occurring earlier in the gain chain (Fig. A30B).

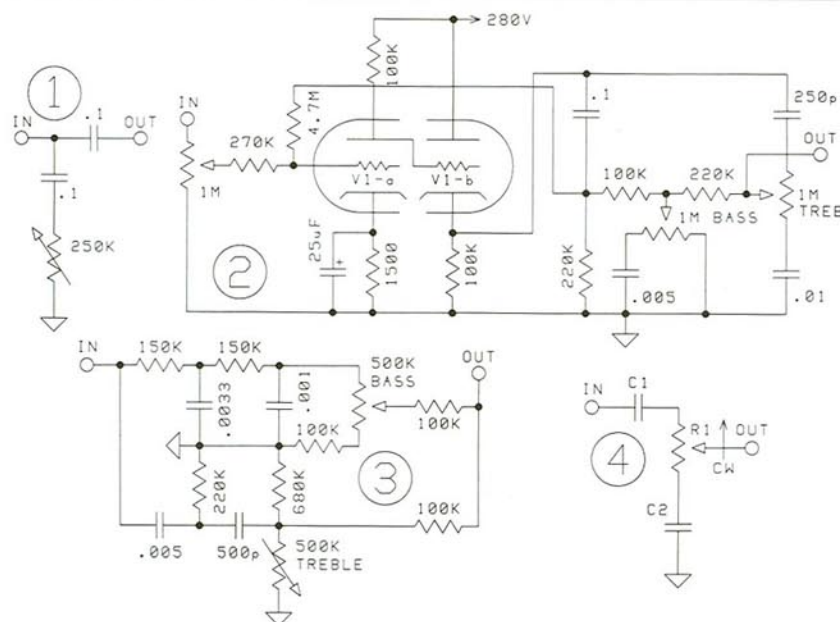Some tone networks resist divider analysis be-

Fig. A34. More guitar amp tone networks. 1—Identical to treble-cut circuit found in most axes. 2—Bass & treble circuit that appeared in Fender amps in the 1950s. Noteworthy mainly for negative feedback path through 4.7M resistor. 3—Bass & treble circuit from Supro S6699; consists of passive 12 dB/octave lowpass filter feeding bass pot; passive 12 dB/octave highpass network feeding treble divider. 4—Common configuration in older amps that provided a single tone control. When wiper of R1 is CW, treble frequencies see a low-impedance path through C1. When wiper is CCW, pot makes a treble cut circuit by divider action with C2.
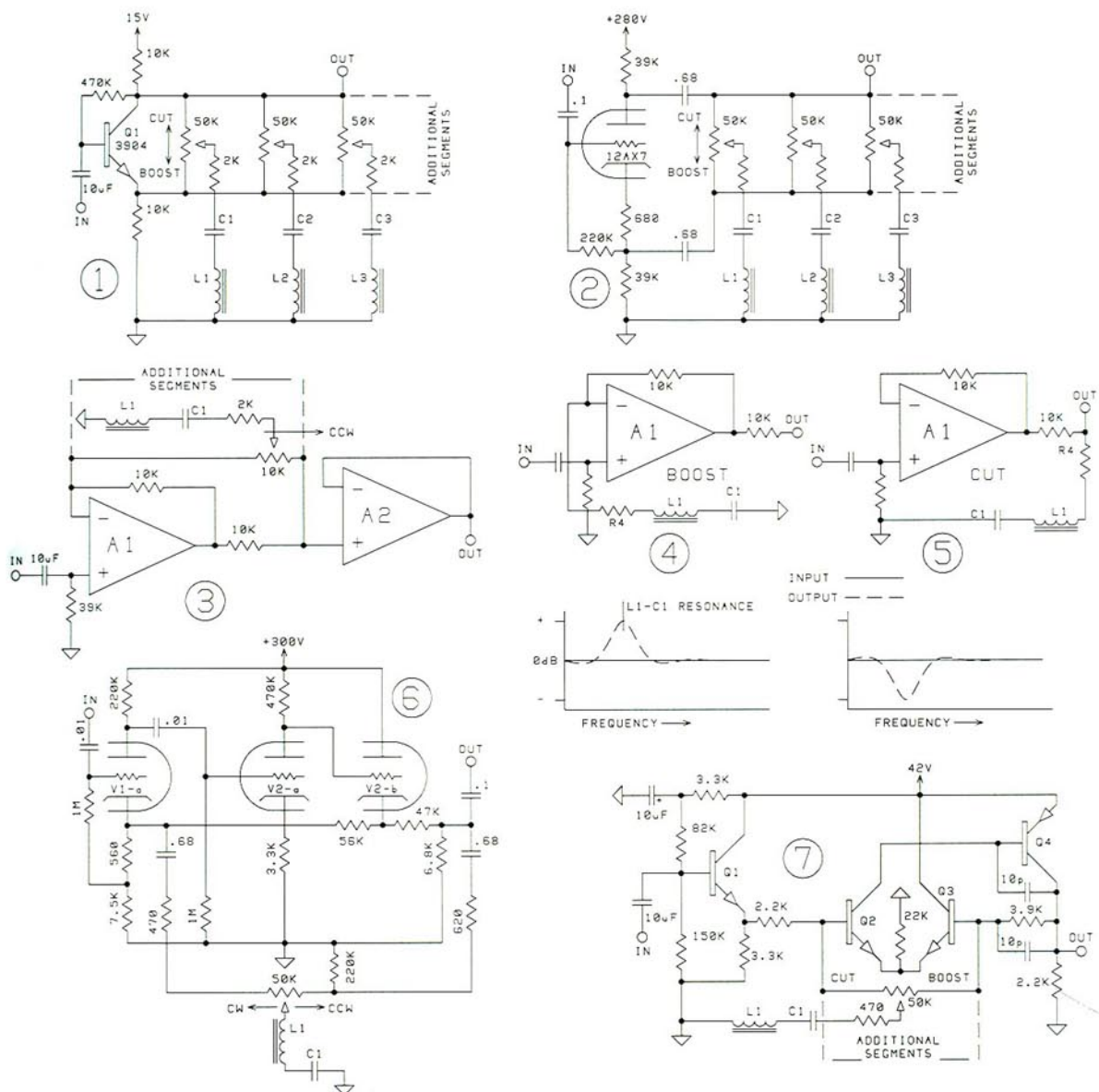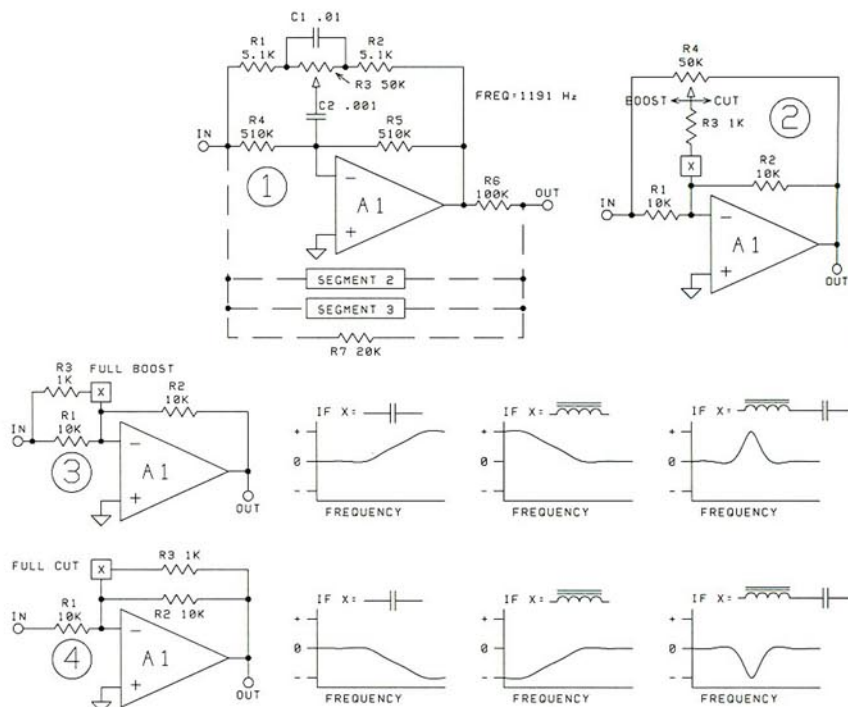
Fig. A35. Approaches to graphic EQ based on LC series resonance. 1—Single-transistor graphic EQ combines passive cut with active boost. When boost/cut pot is fully CCW, RLC network appears at Q1'c collector, whose impedance approximates 10K. Divider action gives dip at resonance; 2K resistance (or other value desired) controls depth of dip. When boost/cut pot is fully CW, RLC network appears in parallel with 10K emitter resistor. Because gain of this amp approximates (collector imped-ance ÷ emitter impedance), a peak occurs at resonance; 2K resistor limits gain. 2—Triode circuit works on same basis; parts require suitably higher voltage rating. Circuit is from an Acoustic® G100T amp, circa 1981. 3—Op-amp–based graphic EQ also achieves active boost/passive cut. When boost/cut pot is fully CW, series RLC network appears as an impedance be-tween inverting input and ground (4). Gain becomes 1 + (10K÷RLC impedance). When boost/cut pot is fully CCW, series RLC impedance makes a divider with 10K resistor tied to A1's output (5). A2 buffers the high-impedance output. 6—Active EQ built into the original Ampeg SVT. Signal coming off amp's PBT network is boosted and inverted in V1-a; boosted and invert-ed in V2-a, whose output is buffered by follower V2-b. A portion of the now-in–phase signal feeds back through 56K resistor to V1-a; but, because signal enters through the cathode instead of the grid, feedback is negative. The three tubes together make a feedback-dependent gain stage. When 50K pot is fully CW, series LC network appears predominantly at V1-a's cath-ode, coupled through 680-ohm resistor and 0.68µF cap. Divider action shunts resonant frequencies, preventing negative feedback and thereby boosting gain for those frequencies. When 50K pot is fully CCW, series LC network appears predomi-nantly at the output end of the 47K resistor, reducing gain by divider action. Circuit may require changes if duplicated using modern triodes, because the original tubes were 12DW7s, whose segments exhibited unequal transconductance. 7—Basic circuit from the Realistic® Model 31-1986 graphic equalizer, which debuted in 1976. Q2-4 form in essence a discrete op amp. When 50K pot is fully CCW, gain reduction occurs by divider action of RLC network with 2.2K resistor at Q2's base. When pot is fully CW, gain becomes proportional to ratio of 3.9K resistor to RLC impedance. Most common transistors work well in this circuit, e.g., 3904, 3906, etc. Relatively high supply voltage is necessary for adequate headroom with single-transistor stag-es. Variants of this circuit appeared in Acoustic and Mesa Engineering® amps of the late 1970s.

Fig. A36. More approaches to EQ. 1—Graphic equalizer typical of op-amp–based circuits of the late 1970s; up to 5 bands. To tune to different frequencies, scale caps by simple ratio, always keeping their relative ratio at 10:1. 2—One of the most versatile EQ circuits; the nature of X determines the response. Can use inductors & caps (3, 4), or active bandpass filters. If X is a state variable filter the resultant EQ gives independent control of boost/cut, center frequency, and bandwidth: parametric EQ. Utility is further enhanced by the fact that one op amp accommodates multiple boost/cut pots and separate X-networks.

Bass and treble pots wired as a dual such that treble boost equals bass cut (and the reverse) give a function known in guitar circles as tone-X, perhaps due to shape of curve at both extremes.

Equalization

Whether active or passive, classical tone controls focus at the outer frequency limits. Tweaking crowds their influence into the midrange, but a better approach falls under the partly artificial category of equalization (EQ), defined here as bandwise boost/cut between the bass and treble extremes. EQ demands segregating frequencies into bands, usually by means of circuits known as bandpass filters (BPFs). Classical BPFs harness LC resonance; modern circuits employ active bandpass filters. *Active filters* consist of RC networks combined with op amps to simulate functions realized with LC networks in the pre-op–amp era. A broad topic in itself, active filters have been treated extensively elsewhere (Ref. 16).

In light of the way gain stages work, several methods exist to achieve bandwise boost/cut (Fig. A35). A single transistor or triode uses a pot to vary the influence of a series resonance tied to ground. When the network appears at the tube's plate or the transistor's collector, a notch results by divider action. When the network appears at the tube's cathode or the transistor's emitter, a peak arises due to the low-impedance path to ground for the resonant frequencies.

Discrete op amps built from transistors or triodes achieve cut passively, but realize gain by placing resonant networks so as to reduce negative feedback for the resonant frequencies.

The equalizers just described use one or more bandpass filters, each of which has an end tied to ground. Floating bandpass filters enable active boost and cut using IC op amps combined with a pot to varying the action of the BPF. When the BPF appears in the
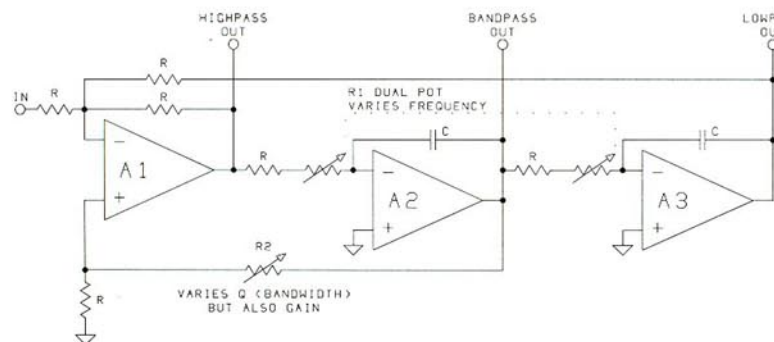

Fig. A37. Basic elements of op-amp–based state variable filter. Circuit naturally provides HP, BP, and LP outputs. Center frequency of BP output varies with setting of dual pot R1. R2 varies bandwidth, but also gain. More complicated circuits allow bandwidth to be varied without changing gain; Parametro-Matic achieves this with a dual pot; Parametro-Matic II injects the signal into the Q-control path, such that divider action of a single pot at A1's noninverting input alters bandwidth without changing gain. By giving the filter unity gain and making it an 'X' element in the circuit shown in Fig. A36–2, parametric EQ is realized.
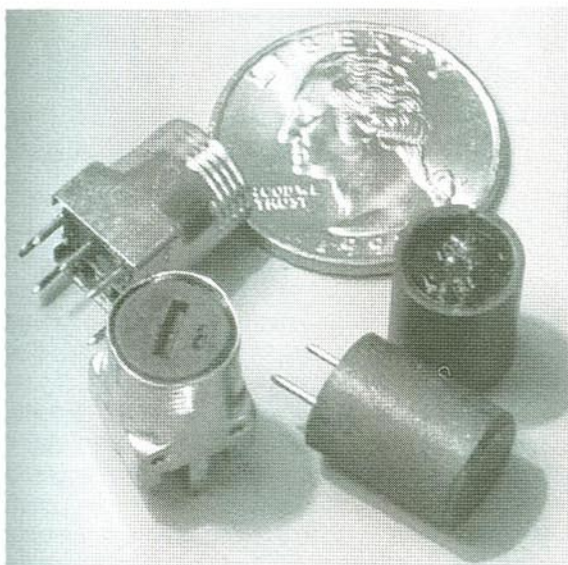
Fig. A38. Fixed and variable shielded inductors typically used at RF and ultrasonic frequencies. These coils adapt to the audio band when mated to low-Q caps. They saturate at relatively low voltage, but prove adequate for many 9V stomp boxes.

op amp's feedback loop, frequencies passed by the filter see a low-impedance path, reducing gain. When the BPF appears in parallel with the input resistor, frequencies passed by the filter see a low-impedance path, boosting gain.

The nature of the BPF defines two types of equaliz-

Fig. A39. Practical simulated inductance. Formulas demonstrate that the best way to change the Q of the resonant network without changing resonant frequency f_R is to alter the ratio of R2:R1 while keeping their product constant. Component values in table are taken from data sheet on LMC835 digitally controlled graphic equalizer (Ref. 18); scale both caps by the same percentage to tune to different frequencies. These values give a Q that suits many audio applications. A1 should possess a high gain-bandwidth product; LM833, NE5532, and similar types work well. Illustration shows the simulated inductance in series with C1, but it also resonates with caps in parallel.

$$L = (C2 \times R1 \times R2)$$

$$f_R = \frac{1}{6.28\sqrt{L \times C1}}$$

$$Q = \sqrt{\frac{C2 \times R2}{C1 \times R1}}$$

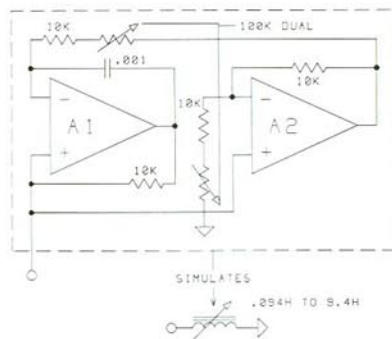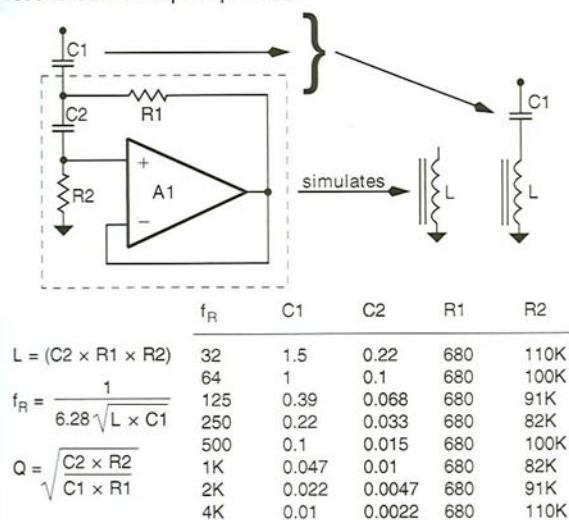| f_R	C1	C2	R1	R2
32	1.5	0.22	680	110K
64	1	0.1	680	100K
125	0.39	0.068	680	91K
250	0.22	0.033	680	82K
500	0.1	0.015	680	100K
1K	0.047	0.01	680	82K
2K	0.022	0.0047	680	91K
4K	0.01	0.0022	680	110K

Fig. A40. Two op amps simulate variable inductor tied to ground. Like fixed simulated inductance of Fig. A39, this one resonates in parallel or series circuits. Scale inductance by raising or lowering value of 0.001 µF cap; circuit simulates enormous inductances. Use polystyrene or similar high-quality cap.

er. Multiple BPFs, each having a fixed center frequency and bandwidth, combine with variable boost/cut to produce a *graphic equalizer*. One or more BPFs, each having variable center frequency and bandwidth, combine with boost/cut to give a *parametric equalizer*. The most common BPF which allows variance of frequency and bandwidth is known as a state variable filter (Fig. A37). The "sweepable mid" now in vogue describes quasi-parametric EQ, combining variable boost/cut and frequency with fixed bandwidth. Fig. A36 illustrates circuits that let the builder realize graphic, quasi-parametric, and true parametric EQ.

Resurrecting Inductors

Inductance lives like a pearl in the oyster of tone, for warmth commands a price. Tone-shaping inductors tend to be large and costly, and prone to saturation in the deep bass. They remain practical when the sound of an inductor flows only from inductance. Cap-based tone circuits predominate these days to the point that some players have never heard how warm inductor-based circuits sound.

While no inherent obstacles keep the builder from designing tone circuits around inductors, suitable parts aren't easy to find. Scarcity of 0.1–10H inductors prompted builders to adapt inexpensive audio transformers to the task. This compromise is practical, but in many cases involves a low Q.

Shielded inductors built for use in RF and ultrasonic range possess impractically high Qs, suitably diminished by pairing them with high-value/low-Q caps. These inductors can be had in values up to 0.15H. They saturate at a few volts$_{p-p}$, but this leaves enough room for many tone circuits.

Simulated inductance makes a powerful alternative. Besides resisting magnetically coupled hum, the simulated inductance can be configured for a high or low Q. This makes a simpler option than a state variable filter for a sweepable mid, and is readily adapted to voltage control by replacing resistors with LDR-based optocouplers (Figs. A39, 40).

Fig. A41. One means to implement a complex tone curve of fixed shape, variable extent. Curve resolves into three discrete segments: fairly broad 12-dB peak at ~85 Hz; narrow 20-dB dip at about 600 Hz; and slow 6-dB rise above 1 KHz. Observing the general principle to implement boost in parallel segments, A2 is an active bandpass filter tuned for gain and Q to give the 12-dB peak near 85 Hz. Unity-gain inverting buffer A3 is given a gentle rise above 1 KHz by R4-C3. The two boost curves combine in summing amp A4, whose gain can be altered, if desired, by changing R9; A4 also reverses the inversions that took place in A2 & A3. Dip at 600 Hz is implemented as a series element by a phase-cancellation network placed between output of A4 and input of A5. Pot R12 fades between full tone curve and unmodified signal. Input buffer A1 is not needed if the preceding stage has low impedance. Net signal path is noninverting.
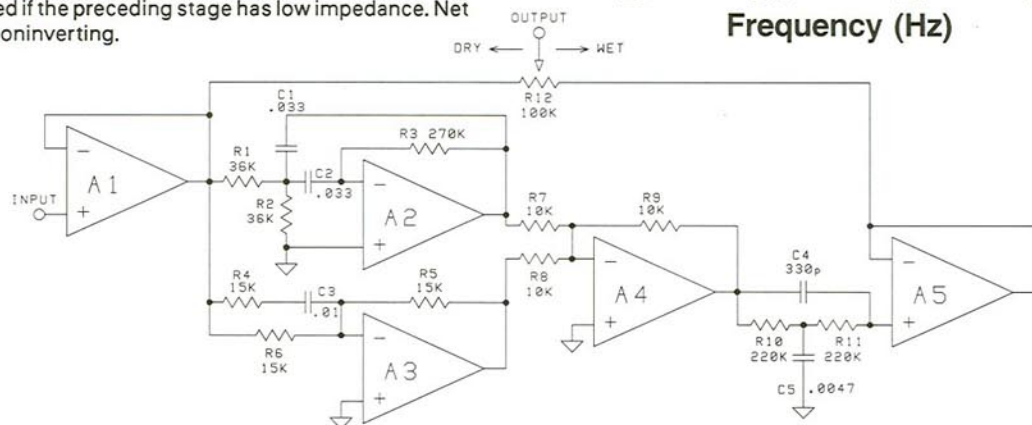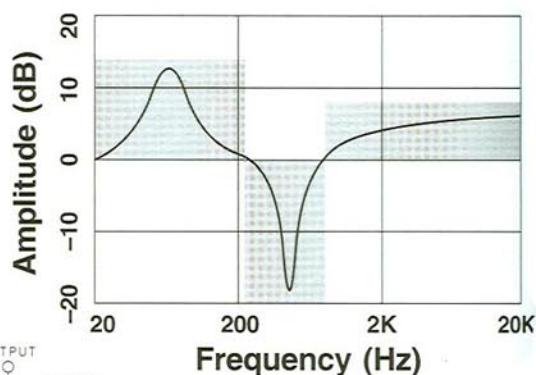
Cooking Up Custom Tone Blocks

It isn't unusual for a player's sound to consist of bass and treble, and sometimes midrange control in the amp; treble cut in the axe; plus outboard EQ. These combinations can be as maddening to specify as they are to reproduce. Players who find themselves fiddling with three sets of knobs to get the right tone may prefer to build their custom EQ into a dedicated box.

First, establish the curve. The fastidious approach demands a tone sweep, through outboard EQ and the amp, plotting the curve with enough accuracy to guide emulation. Measure output at the speaker terminals. The curve must account for treble cut in the axe, if the axe tone pot is not wide open. If a tone sweep is not an option, estimate the curve based on the amp's tone control settings and outboard EQ.

Whether measured or approximated, the curve gives a starting point. Fine tuning takes place by ear, at the breadboard. Divide the curve into peaks and dips. Perform estimated curve fitting using the active or passive tone blocks discussed in this Appendix. Generally, place boost-elements in parallel, cut-elements in series; and place boost before cut. Combine boosted signals in a summing amp that, optionally, cancels the gain of summed signals. Remember to re-invert after inverting stages, to avoid cancellation with noninverted feeds.

As a last touch, make the tone curve variable. Fig. A41 gives an example of a complex tone curve realized by this process. Subjectively, the curve lends electric guitar the timbre of an acoustic.

Active Electronics

Active, in guitar parlance, implies powered circuitry in the axe, usually a preamp and often a tone block. This shortens the span from pickup to first gain stage, making a system less prone to hum by that path. The preamp's low-impedance, line-level output resists pollution in transit a lot better than the pickup's high-impedance, instrument-level output. Low impedance also frees tone from dependence on length and grade of the axe cord. A balanced feed is easily achieved, allowing the player to plug into a rack without going through a direct box. Active tone circuits broaden the control range over passive. In fact, they enable any type of tone control up to parametric EQ. While most active setups stop with preamp and tone, nothing but imagination stands between the builder and his own version of Lester Polfus' stage rig.

Working up an active setup involves a procedure not unlike that used to cook up pedal effects:

1. Define the active functions as specifically as possible.

2. Assess #1's objectives in light of general practicality. For example, you can't boost the output of a hot pickup by 30 dB if the system runs on nine volts.

3. Prepare a block diagram of the active setup.

4. If the block looks feasible, prepare a schematic. Breadboard and test the setup. Listening is the key phase; measurements take a back seat to sound.

5. If the result fails to please, return to Step 3 and try an alternative approach.

6. Once you have "the sound," lay out the circuit board, build and install the setup.

Options for preamp and tone number too many to mention more than a few, shown in Fig. A43. These simple circuits make good choices for the first-time builder.

Power Options

Like stomp boxes, active systems usually run off one or two 9V batteries. Housing and changing batteries forms a big part of the setup. At least one manufacturer offers an alternative to unscrewing a cavity cover every time, a hinged-door battery compartment that mounts in a routed opening.

Repeated flexing of the battery cable eventually leads to internal separation. Veteran 9V users under-
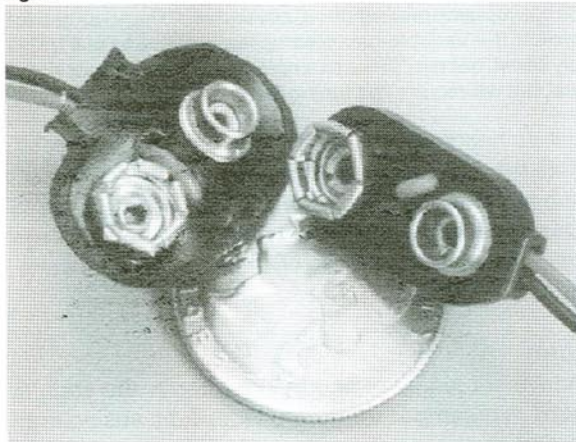
Active Setup Options

Functions
  preamp
  tone
  effect
    compression, distortion, etc.
Subcircuits
  transistors
  op amps
  specialized ICs
Output
  mono
    unbalanced
    balanced
  stereo
Power
  one battery
  two batteries
  phantom

stand the weak link to favor the point the wire crimps to a snap inside the vinyl housing. The player can prolong life of the snap by securing the battery, and by using molded-head connectors whose durability justifies their higher cost (Fig. A42).

The power switch integrates conveniently into the output jack; the power is off if the axe is unplugged. This and other switching options are discussed in a separate Appendix.

Phantom power is the art of sending current through a conductor that also bears an audio signal. Studio condenser microphones charge their diaphragms with 48V borne through the standard balanced cable. An extra wire carrying power is remote but not phantom, since this wire bears no audio. Phantom power eliminates batteries in the axe and ex-

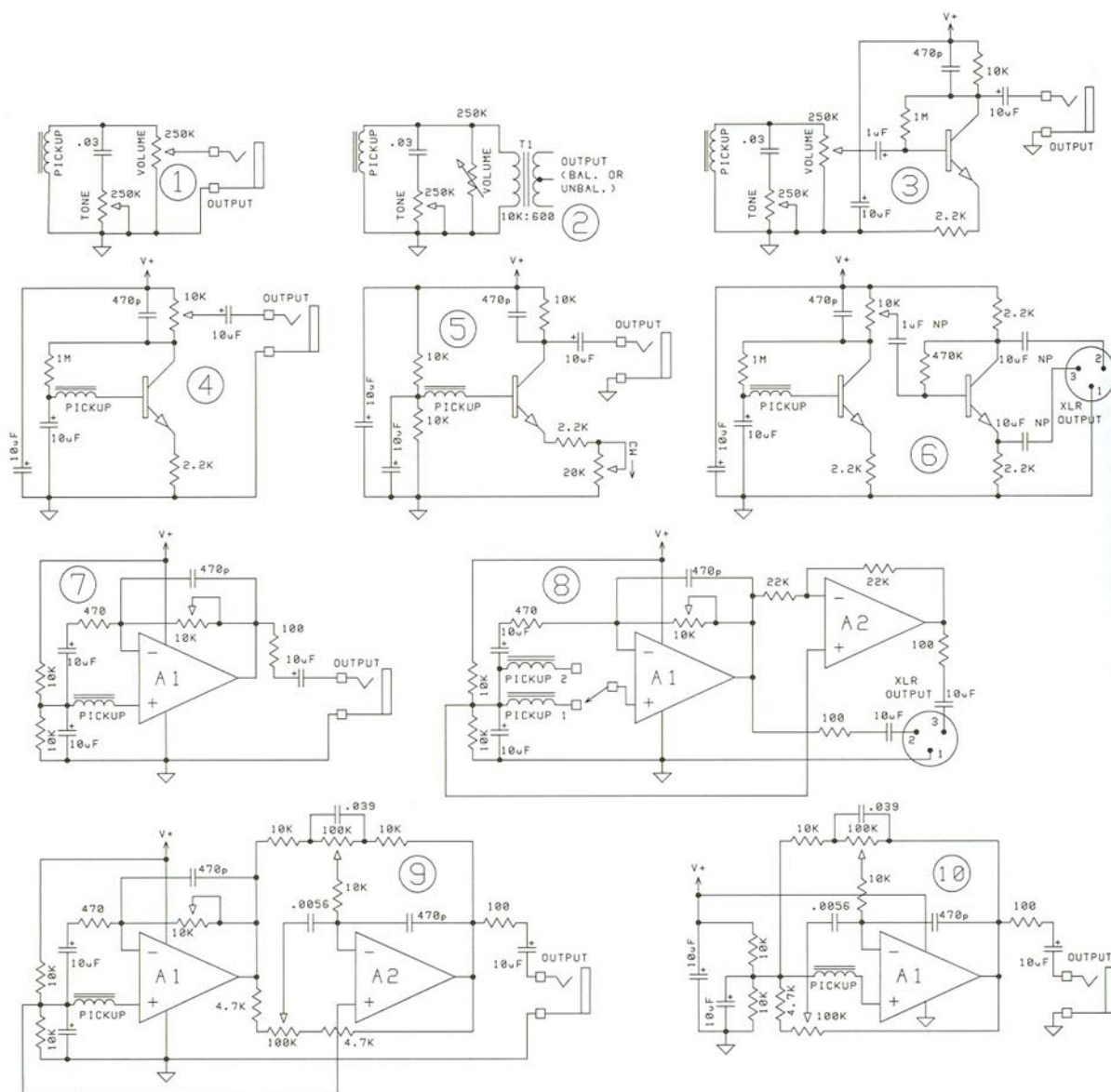Fig. A42. Light-duty 9V snap left, heavy-duty molded snap right.

Fig. A43. 1—Typical passive guitar wiring. 2—passive option for super-hot pickups: balance the feed and lower the imped-ance using a transformer in the axe. Transformer loads the pickup; how this affects the sound can only be determined by lis-tening. Output can be treated as balanced or unbalanced. 3—Gain stage appended to stock network gives line-level, low-im-pedance output. 4 & 5—Two straight gain methods and approaches to volume control. 6—Output is easily balanced. 7—Simple op-amp preamp. 8—Conversion of single-ended to balanced output using op amps; also, shown with pickup selector switch. 9—Variable-gain preamp feeds active bass & treble control. 10—Simpler tone circuit; noninverting, but can only ap-ply boost, not cut. These examples illustrate a few possible active setups for guitar & bass. Component values shown are typ-ical, but will require fine tuning for desired sound, supply voltage, etc.

pands headroom to any practical voltage available from the phantom source. The builder weighing an ac-tive setup might at least consider this option; Fig. A45 shows several possibilities. The most versatile ap-proach demands a balanced feed and a pair of trans-formers.

Hum

Active setups open another chapter in the saga of hum. A preamp in the axe reduces hum arising from exposure of a high-impedance path. It eases none of

single-coils' bent to pluck hum out of the ether, and may actually breed new avenues for hum, for phan-tom-power transformers might as well be pickups.

Many issues that arise in working up an active set-up have to do with hum, and often require empiric res-olution; for instance, how grounding the strings af-fects hum; whether to run the package off a single supply with an artificial ground or a dual supply with a true ground; etc. The builder can reduce the general predilection to hum by choosing pots having the least resistance in keeping with needs of the circuit; 10K
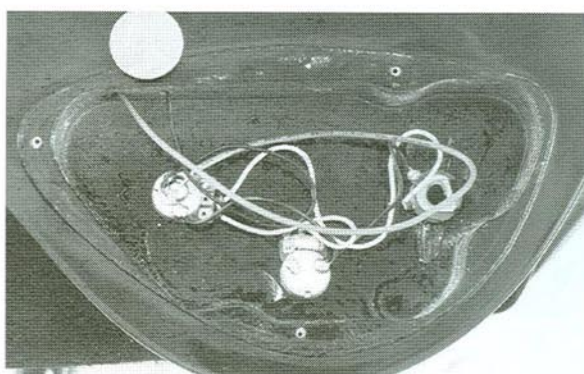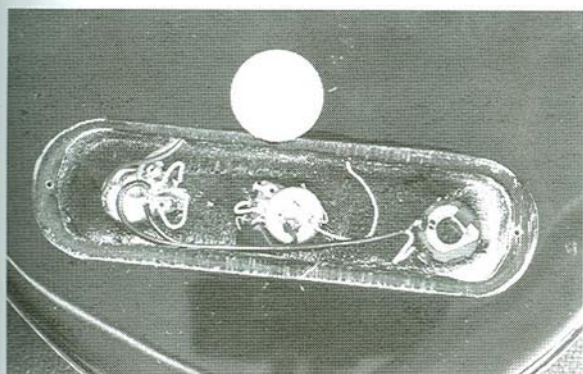
Fig. A44. Two control cavities, scaled relative to a quarter. Right cavity could house an elaborate system. Not clear from photos, but both cavities offer sufficient depth to accommodate the thickness of a 9V battery.

pots instead of the 250K and larger pots common to passive electronics. Some setups benefit from shielded cable for the few inches between electronics and pickups or switches.

Shielding the electronics cavity reduces hum in some setups, but only as a third or fourth priority. A grounded aluminum-foil lining makes an excellent voltage shield that does little to attenuate magnetically coupled hum. The utility of shielding rises with the impedances of the paths being shielded.

Construction
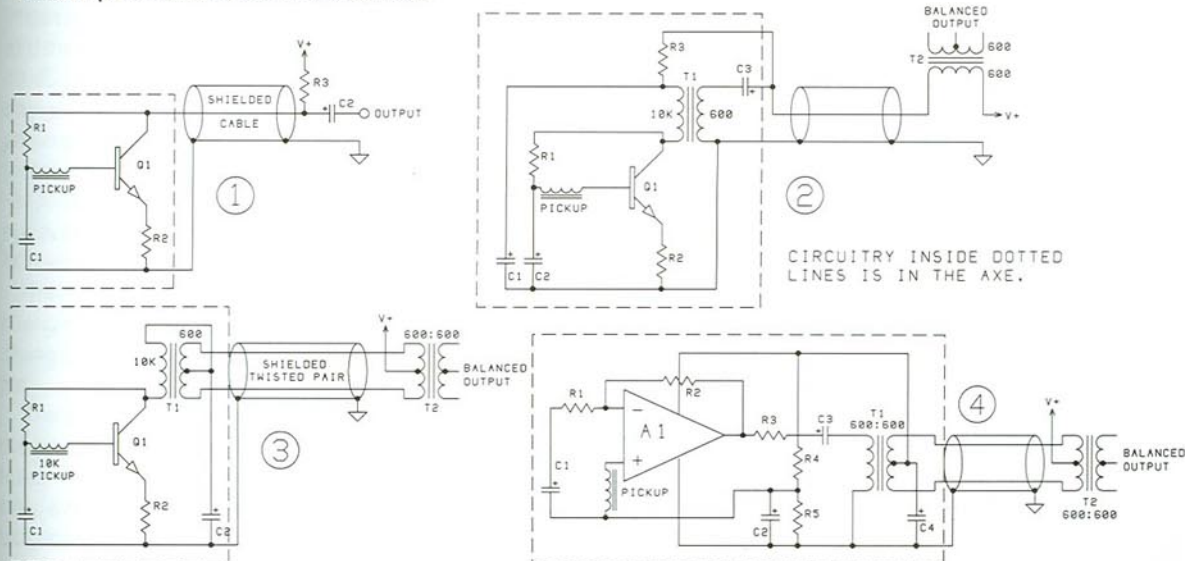
Axe-dwelling circuitry differs from outboard in having to fit cramped quarters. The average cavity gives room enough that the builder need not resort to surface-mount parts. Choosing small versions of standard parts saves a lot of space. Tantalum and mono-

lithic caps, and $1/8$W resistors aren't much bigger than their surface-mount counterparts, and can actually better them in horizontal density.

For vertically cramped cavities, lay out the board to flat-mount the parts. Radial-lead and disc caps lie flat, as do resistors and diodes. Use in-line holes for a transistor's leads so the body folds easily onto its flat side. Forgo sockets for IC's to save about $1/8$" vertically. Reverse these priorities for horizontally cramped cavities; vertically mount as many parts as possible. Stacked circuit boards call for a bit of ingenuity, but can solve otherwise difficult constrictions. Additional space savings flow from using pull-switch–equipped pots.

Perhaps the most valuable construction guide is to *plan the work*. Inspect the cavity, estimate how much room an active package can occupy and what space

Fig. A45. Phantom power schemes. 1—Simple; limited utility. 2—A more complicated scheme uses unbalanced cable but yields a balanced output. 3—Three-conductor cable opens greater possibilities. Single-transistor amp shown. 4—Phantom-powering an op amp; could power complex op-amp circuits, including parametric EQ, compressor, etc. In all cases, circuitry shown inside dotted lines is in the axe. Voltage supply in remote box allows headroom well beyond that practical with axe-mounted batteries. The phantom system must account for the coils' resistance. In the case of Mouser 42TX016 600:600-ohm transformers, this measures less than 200 ohms per winding. Current limit for an 8V system: 8000 mv ÷ 200 ohms = 40 ma. Most setups run on a fraction of that current.
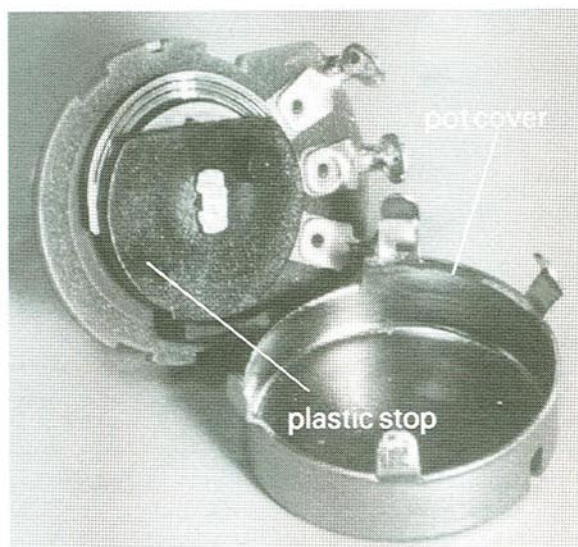
Fig. A46. Some pots contain a plastic stop likely to melt if pot cover is tinned in place. A preferable approach is to remove cover; tin, clean, and replace.

Fig. A47. Pot equipped with DPDT pull switch, a handy switching option for axe-dwelling circuitry.

the batteries will need. Prepare a cardboard mock-up of the circuit board, try it for fit. Diagram the existing wiring, to facilitate a return to stock; and diagram prospective routing for the active setup's wiring. Poor planning could lead to an installation well underway, at which point the builder realizes that reaching the next connection requires removal of the setup.

Ground all pots' covers. When installing a virgin pot, tin the cover before assembly. Tinning a cover can be difficult, even hazardous, while the pot is mounted in the axe. In some cases, tinning requires enough heat to damage a plastic stop tied to the pot shaft (Fig. A46). One way around this problem is to remove the pot's metal cover for tinning. If the cover is lacquered, or plated with metal that resists tinning, file a spot down to the bare metal, tin the pot. This approach allows use of high heat available from a soldering gun. Once the cover has cooled, clean it of debris and replace it on the pot.

Use shielded cable for high-impedance paths; minimize the length of these paths. This holds especially for piezo setups.

Solder as many connections as possible prior to installing pots & boards. Tin everything you cannot presolder: pot terminals, wire ends, component leads, and especially pot cases. This makes soldering a matter of tacking.

First-time builders may find it prudent to work on an instrument they are willing to sacrifice, and, even so, to start with reversible mods. Don't drill, file, cut, or rout. Before changing anything, prepare a detailed diagram of the existing circuit. If the stock setup is particularly complex, attach small tape labels to the wiring. Fit the electronics to the available space; save all parts removed. The results will tell whether permanent mods are worthwhile.

Project No. 28
Axe-O-Matic

AOM is an axe-dwelling preamp with active bass & treble. Runs off 9V and up; power-on with plug insertion.

Circuit Description

A single pickup, or the pole of a pickup selector switch, ties through C3 to noninverting preamp IC1-a, whose gain varies by trimpot R2 from 1 to 22. Preamp output couples directly to an active bass & treble network made up of IC1-b and associated components. Tone control output couples through C8 to pot R11, and from R11's wiper through R12 to the output jack.

Power switching is accomplished by a switch-activated transistor. When no plug is present, the base of Q1 is pulled HIGH by R13, turning Q1 OFF. Leakage current measures in microamps and is comparable to most batteries' self-discharge current. Insertion of a 2-conductor plug into the output jack grounds R14, turning Q1 ON, powering the circuit.

Use

Pots have these functions:

R2   preamp gain trim (1–22)
R5   treble
R9   bass
R11   output level

Use is self-evident.

Notes

Preamp gain must account for the pickups' output, and the fact that the bass and treble circuits add up to 20 dB of boost at the frequency extremes. A freshly charged nicad gives about 5V$_{p-p}$ of headroom. Raw preamp output should not exceed 500 mv$_{p-p}$ to allow for the tone controls' maximum boost. Preamp gain of ~3 (10 dB) proved suitable for a pickup whose raw output averaged 100 mv$_{p-p}$. The prototype used a fixed 1K resistor in place of R2.

The builder can alter the tone controls' ranges by changing the values of C4 & C6, or boost/cut by changing values of R8/R10 for bass, and R4/R6 for treble.

The prototype was tested in an inexpensive, single-pickup bass that had only two potentiometer holes. To accommodate three functions in two controls, a dual 100K pot was used for R5 and R9. The treble control was wired in stock fashion, but connections to the end-terminals of the bass pot were reversed, such that treble boost also gave bass cut, and the reverse. The setup proved versatile to the point of exuberance.

The 33172 dual op amp was chosen for its low current/long battery life/large headroom. Despite relatively high input noise voltage (32 nv), no audible reduction in noise was noted with TL072 (18 nv) and 5532 (5 nv) types. The prototype drew just under 1.8 ma when running off a freshly charged nicad.

The treble circuit makes a pretty fair amp for AM radio signals. If interference results, increase C1 to 220pF, increase C7 to 0.001μF, and wire a 100pF cap from IC1 pin 3 to ground.

Minimum hum required grounding the strings and both potentiometer cases. Since grounding schemes vary among axes, the builder should expect to experiment to find the wiring requirements for least hum.

AXE-O-MATIC PARTS LIST

Resistors
R1 470
R2 10K single-turn trimpot
R3 150K
R4 4.7K
R5, 9 100K pot
R6 2.2K
R7, 8, 10, 14, 15, 16 10K
R11 10K audio-taper pot
R12 100
R13 1M
R14 10K

Capacitors
C1 100pF
C2 4.7μF
C3, 8, 9 10μF aluminum electrolytic
C4 0.039μF polypropylene
C5 220μF aluminum electrolytic
C6 0.0056μF polypropylene
C7 470pF

Semiconductors
D1 1N4001
IC1 MC33172 dual op amp
Q1 2N3906 PNP transistor

Miscellaneous
J1 ¼" 3-conductor jack
wire, circuit board, knobs, solder, 9V battery snap, etc.

2.5" × 1.25" reference box
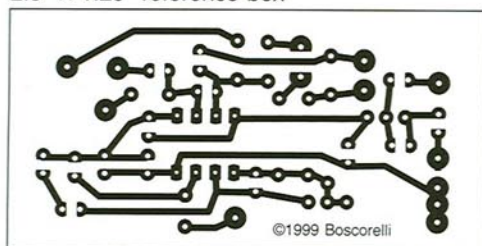

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Fig. 28–1. Axe-O-Matic circuit board.

Fig. 28–2. Axe-O-Matic layout & wiring diagram.
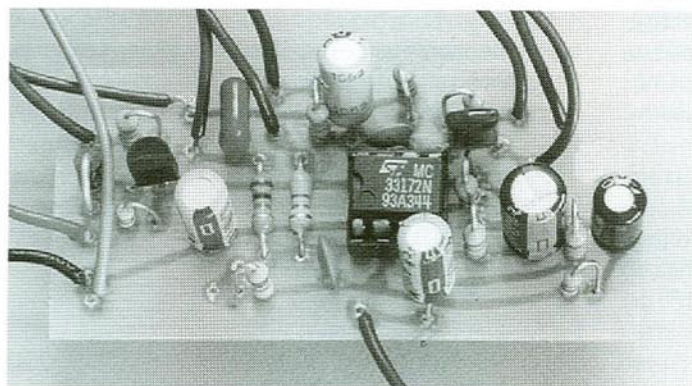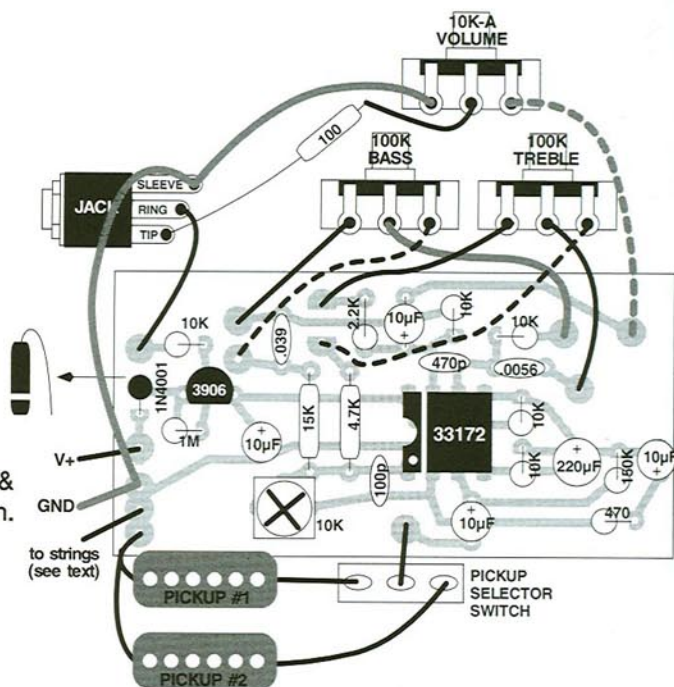

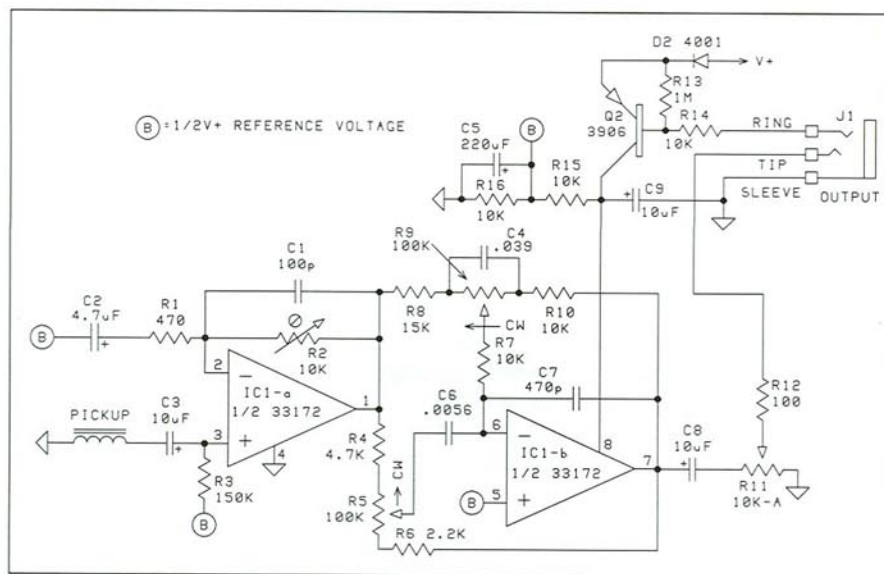Fig. 28–3. Axe-O-Matic prototype board.

Fig. 28–4. Axe-O-Matic schematic.

Tremolo

Tremolo means rhythmic intensity change; simple amplitude modulation, but also an effect of nuance. Tremolo suits the pedal scene because few amps allow control of anything but rate and depth, excommunicating the player from traits that truly change the sound. Taking charge of the process opens a vista of possibilities, simply achieved with a control voltage generator and a VCA.

The Control Voltage Generator
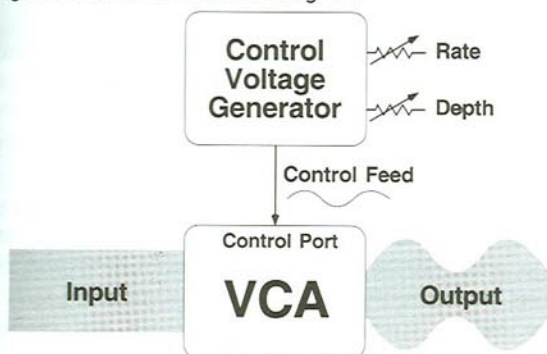
The sound of tremolo depends on the rate and shape of a subsonic control feed modulating an instrument feed. Most circuits use sinewave or triangle control feeds, the sonic difference being small. Function-generator chips make convenient, easily tuned sources, as do op-amp–based oscillators. Sources that generate incidental squarewaves have a way of polluting the signal path with clicks. The builder planning to stick with sine or triangle control might save a few headaches by using a source that does not also generate squarewaves. The modulating waveform may take any shape, with the caveat that sharp edges tend to bleed through when the instrument is silent. The control voltage typically couples to the VCA through an op amp (Fig. A49–2).

The most musical-sounding tremolo operates between 1 and 5 Hz. Rates outside this range are useful for special effects.

The VCA

Any VCA could make a competent tremolo if gain were all that mattered. In fact, feedthrough separates contenders from champs. Feedthrough manifests as *beating*, audible pulsations heard when the instrument is silent. VCAs differ significantly in trimmed feedthrough. The text chose an NE570-VCA for tremolo because this chip's feedthrough is so low as to obviate suppression beyond trimming, even when the control feed clips (Fig. A54).

Fig. A48. Basic tremolo block diagram.

Tremolo Options

Oscillator Complement
  single
  multiple
Rate
  fixed
  variable
    manually varied
    rhythmically varied by 2° oscillator
    level-dependent
Waveform Shape
  fixed
    sine
    triangle
    ramp
    negative ramp
    square
    diode clipped sinewave
    irregular
  rhythmically varied by 2° oscillator
  level-dependent
VCA Resting Gain
  fixed
    maximum
    medium
    minimum
  rhythmically varied by 2° oscillator
  level-dependent
Modulation Depth
  fixed
  rhythmically varied by 2° oscillator
  level-dependent
Outputs
  single
  mixed inverse (requires two signal feeds)
  dual inverse (pan-tremolo)

Feedthrough does not have to doom a tremolo that sounds exceptional otherwise. For example, an OTA driven to squashing might make a "tube sound" tremolo whose feedthrough could be suppressed by a highpass filter. The Vox AC-15 and the Ampeg B-12X contained stiff highpass networks to blunt subsonic feedthrough (Fig. A51–1). Highpass filtering cannot quell artifacts that accompany sharp control feeds, nor hiss modulated by the VCA. Gating and internal companding make viable solutions to severe feedthrough.

Tremolo: The Undiscovered Effect

Amp-dwelling tremolos provide control of rate and depth. This usually proves adequate, because the amp's manufacturer has tuned the other variables to give a pleasing sound. Moving tremolo to the pedal domain awakens sleeping possibilities. The first restriction to go is the number of oscillators. Two summed control feeds create the sound of "dueling
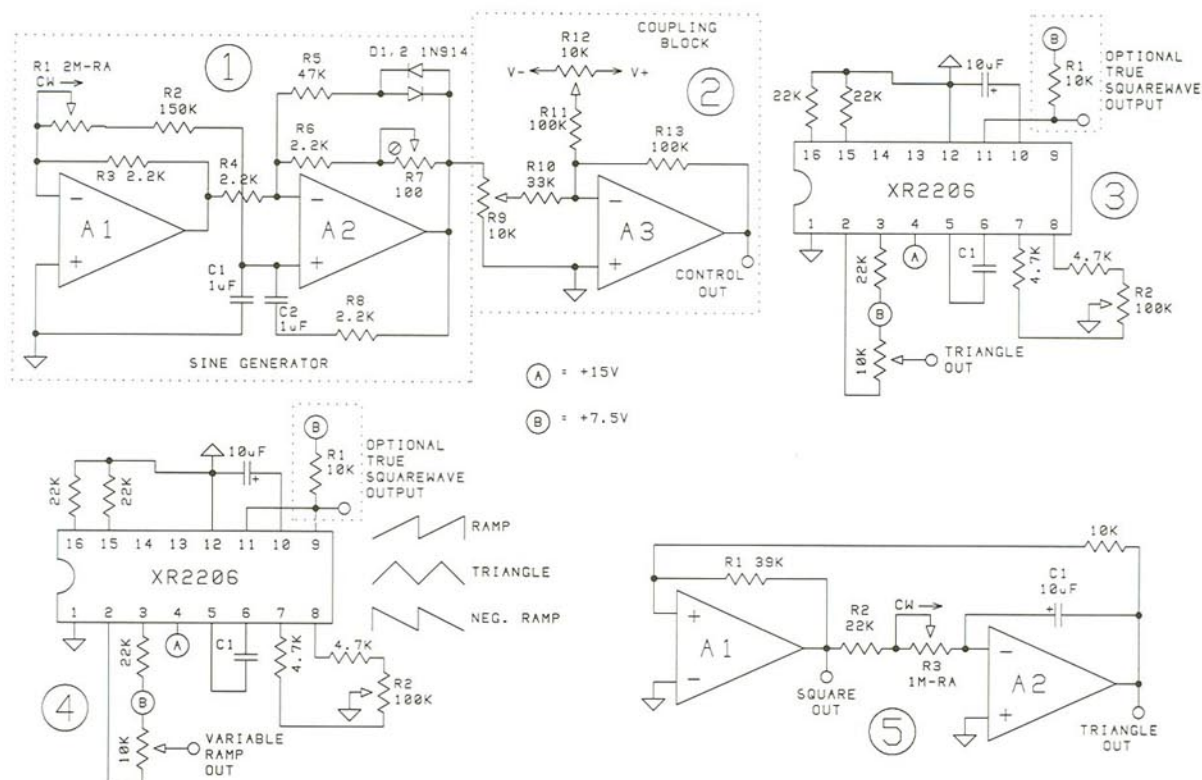
Fig. A49. Four tremolo-ready subsonic oscillators. 1—Sinewave oscillator is adapted from one shown in Ref. 8, p. 451. Original specs call for 1% resistors, but 5% types work fine. C1 & C2 must match reasonably well; 10% tantalum, 10% monolithic, and 5% polypropylene caps have given good results. R2 sets minimum frequency; can be reduced to 2.2K for much higher frequencies. Much lower frequencies can be had by raising value of capacitors, or replacing 2M pot with an LDR shunted by 10M. 2—Coupling block that can be used with all four oscillators. 3—XR2206 function generator configured as variable-rate triangle generator. A true squarewave output is available at pin 9; omit R1 if circuit does not use squarewaves. 4—XR2206 configured to vary slope of ramp output from ramp to triangle to negative ramp. Drawback is that changing frequency necessitates changing C1. Squarewave output gives varied duty cycle from ~10–90%. Omit R1 if squarewave output is not used. 5—One of the earliest op-amp-based function generators; creates simultaneous square and triangle outputs. Has advantages of simplicity and low cost. Value of R1 determines amplitude (which also varies with supply voltage), value of R3 sets rate, in this case, ~0.3–15 Hz; amplitude ~3V_{p-p} with a ±7.5V supply. Range can be changed by value of cap in feedback loop, or value of pot R3.

tremolos" as different rates interleave.

The next casualty, fixed rate, is easily yoked to instrument dynamics; the main instrument, or a secondary feed. Rate can also be made to vary under control of a second subsonic oscillator.

The modulating waveform's shape seems conspicuously neglected in the quest for new sound. Slow ramps create *whah*-like surges in held notes; convincing swells if the player picks in sync with the onset of the ramp. Negative ramps impart the *pong*-like sound of early analog synthesizers. Audibility of altered control-feed shape is greatest at slow modulation rates. Squarewave and quasi-squarewave control feeds yield sounds whose character varies with duty cycle.

Much of tremolo's impact hinges on the VCA's resting gain, and whether the modulating waveform boosts or cuts gain. The typical in-amp tremolo varies control voltage amplitude around some fixed point. Building from scratch puts this static point under the player's command, enabling three distinct dynamic classes. With moderate resting gain, the modulating waveform boosts gain on peaks, cuts gain on dips. With VCA gain at maximum, positive control peaks have no effect because volume cannot grow past max. Negative control peaks reduce volume, resulting in a subtler yet distinct effect. Finally, with zero resting gain, negative control peaks have no effect because volume cannot fall below zero. Positive peaks launch tones from a background of silence; a bold, almost percussive sound.

Once liberated from the amp, tremolo can go looking for partners, such as distortion, delay, or pitch vibrato.

No law says one-tremolo, one-output. A single input feeding dual VCAs wired for inverse gain, rhythmically pans the signal between channels. Two inputs feeding two VCAs synched to an inverse control voltage, and with each VCA feeding a summing amp, results in a rhythmically varied mix of main and secondary feeds. These effects and more are readily
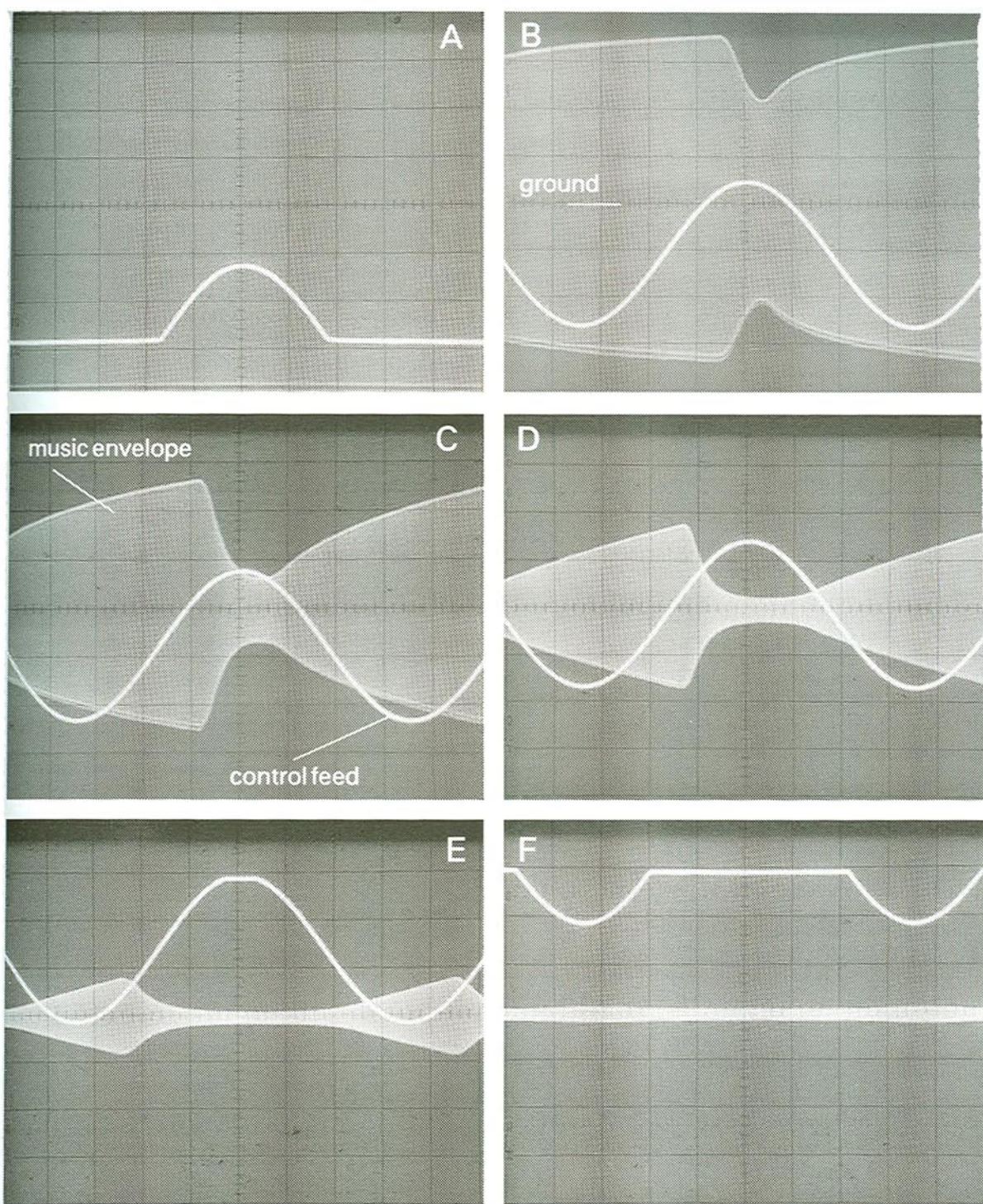
Fig. A50. Function of Fig. A53, LDR-based tremolo. Scope display shows 5-Hz control feed superimposed on signal envelope; both signals are DC-coupled to scope. A—Feed does not exceed ground; LED does not light, signal envelope is not affected. B—When control signal reaches 2V, LED turns on, LDR resistance changes, envelope changes. C & D—resting DC bias of control feed continues to rise, envelope shows greater modulation. The LDR's slow decay distorts the envelope; the resultant sound suits some players' tastes; slower control frequency results in lower distortion and greater tremolo depth. E & F—DC bias applied to control waveform increases; LED stays on, LDR does not turn off. Compare these traces with those of the several Tremolo-Matics, which use chip VCAs. Control feed scale 5V/div., tremolo output scale 1V/div. R8 has been trimmed to give a symmetrical output envelope.
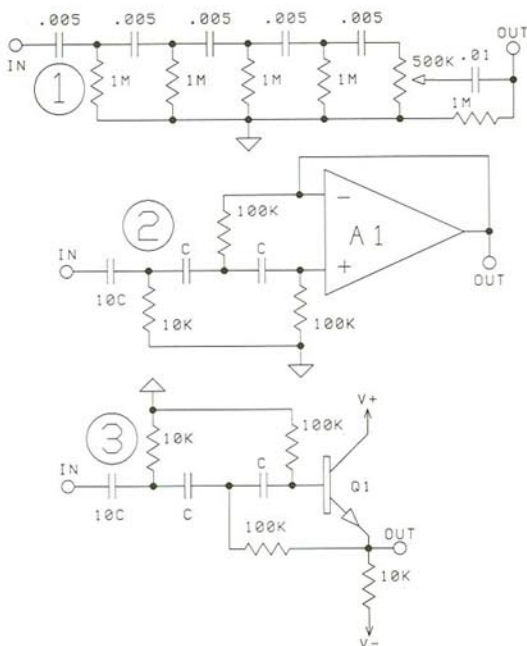
Fig. A51. 1—Passive highpass network found in vibrato channel of the Vox AC-15. Signal must traverse six poles, f_{-3} of each around 32 Hz; lossy. 2—Op-amp-based quasi-18 dB/octave highpass filter. 3—Similar to #2, but based on emitter follower. C = 0.027μF for 60-Hz cutoff.
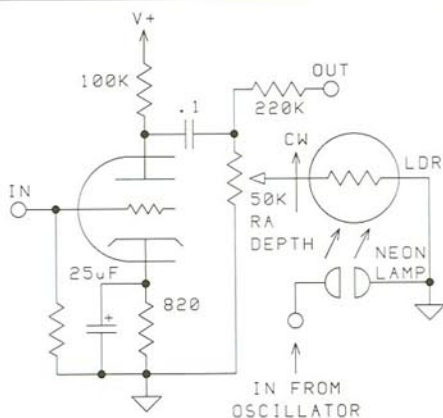
Fig. A52. Schematic representative of Fender 'shunt' tremolo. What is not apparent is that varying the volume at this point takes subsequent stages into and out of tube-sound distortion/compression modes.

Fig. A53. LDR-based tremolo whose I/O is shown in Fig. A50. R1 varies the resting DC bias applied to the LDR; trim R8 for best symmetry of output waveform. A1 must be able to supply enough current to light the LED in the optocoupler; TL07X and similar work fine.
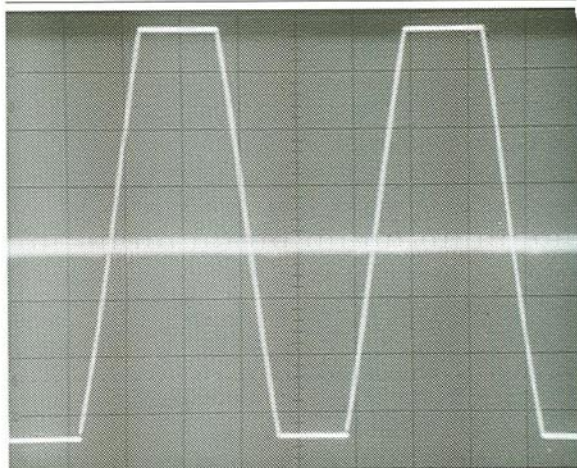

Fig. A54. Trimmed feedthrough of NE570-based VCA is extremely low. Despite control feed clipping at both extremes (output of 3317X op amp feeding 47K resistor tied to pin 1 of 570; 15V supply), feedthrough is lost in noise. Control feed scale 2V/div., VCA output scale 10 mv/div.; sweep 20 ms/div.

achieved by coupling a subsonic oscillator to various voltage controlled blocks.

Tube-Sound Tremolo

Tube amps' on-board tremolos vary gain through several methods, including an optocoupler-based divider, and bias modulation in the preamp. In both mechanisms, each cycle takes subsequent stages into and out of the compression and distortion modes that characterize tube sound. Outboard tremolo allows elaborate modulation control, but cannot replicate the tube amp transfer function through a VCA alone. Combining a sweet tube amp with outboard tremolo lends both sounds an added dimension.

Project No. 29
Tremolo-Matic IV

Claims of tube sound from solid state usually need plenty of salt. Not so TM4. Read on.

Circuit Function

Signal path: Instrument feed couples through C14 to noninverting preamp IC1-d, whose gain is fixed at 11. Preamp output couples through C12-R14 to IC2, an LM3080 configured for gain of ~1. IC2 output feeds a four-pole highpass filter made up of IC3 and associated components. IC3 output couples through C9-R25 to output level pot R26. The signal path is noninverting.

Control path: IC1-b/-c and associated components form a sinewave oscillator whose amplitude is trimmed by R5 and whose rate is varied by R1 over the approximate range 1–10 Hz. IC1-b output couples through R8-a & R9 to inverting buffer IC1-a. R8-b applies a variable positive DC potential through R10 to the inverting input of IC1-a. The control mechanism works thus: When depth control pot R8 is fully CCW, the potential at IC1-a's output is near ground; the audio signal passes from input to output at roughly unity gain. As R8 is turned CW, R8-a increases the sinewave amplitude, up to the trimmed limit of ~7V_{p-p}; the DC voltage applied through R8-b & R10 results in an offset at the output of IC1-a that peaks at about –3.75V. This offset is reflected as a positive offset at the output of IC2, because the control voltage feeds IC2's inverting input. The sinewave impressed on this DC potential alternates the offset between ground and IC2's positive output limit. Rhythmic alteration of headroom effectively modulates IC2's gain.

Use

Pots & switch have these functions:

R1	tremolo rate
R5	sine trim
R8	depth
R26	output level
S1	effect/bypass

First, trim sinewave amplitude. Connect scope probe to IC1 pin 7, trim R5 for ~7V_{p-p}.

Initial settings: R1, R26 fully CW; R8 fully CCW, S1 tremolo. Connect unit to axe and amp, establish desired listening level. In this state the box acts as a preamp/highpass filter with gain of ~1. Turn R8 CW and note the sound with various combinations of tremolo depth and rate.

Notes

TM4 is optimized to run off a pair of 9V batteries. Significant departure from ±6.5 to ±10V is not recommended. The prototype's fixed preamp gain of 11 may have to be altered to suit various guitars. The prototype photo shows R5 as a single-turn trimpot; a multiturn pot is recommended.

The fact that TM4 sounds different from VCA-based tremolos should not surprise the player, since it contains no VCA. Gain modulation by unipolar squashing is the same mechanism found in certain

TREMOLO-MATIC IV PARTS LIST

Resistors
R1 2M reverse-audio pot
R2 180K
R3, 4, 12 2.2K
R5 1K multiturn trimpot
R6 1.5K
R7, 20, 22 47K
R8 10K dual pot
R9, 11, 23 100K
R10 200K
R13 4.7K
R14, 25, 29 10K
R15, 16 200
R17 7.5K
R18, 28 1K
R19, 21, 24 39K
R26 1K audio-taper pot
R27 150K

Capacitors
C1, 2 1μF 10% tantalum
C3, 11 200pF
C4 0.001μF
C5, 6, 7, 8 0.1μF
C9, 14 10μF nonpolar electrolytic
C10 10pF
C12, 13, 15, 16 10μF aluminum electrolytic

Semiconductors
D1, 2 1N914
D3, 4 1N4001
IC1 TL074 quad op amp
IC2 LM3080 OTA
IC3 TL071 op amp

Miscellaneous
S1 DPDT switch
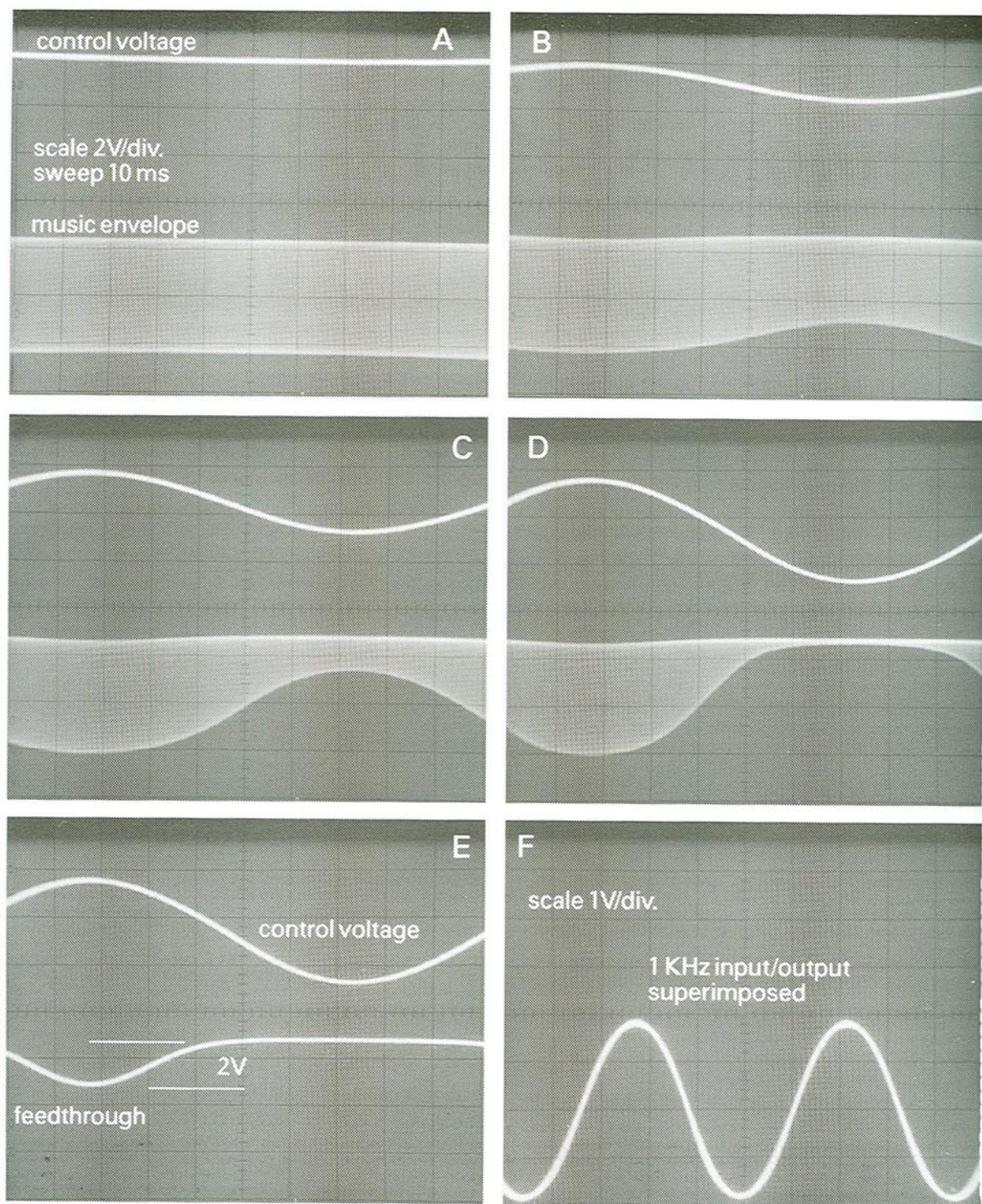1/4" jacks, wire, solder, circuit board, 9V battery snaps, etc.

Fig. 29–1 (this and facing page). A—Tremolo-Matic IV control voltage top trace, music envelope bottom trace. B–D—Control voltage amplitude increases, music envelope shrinks unidirectionally due to loss of headroom. E—Removal of music feed reveals substantial feedthrough, about $2V_{p-p}$. F—LM3080 input/output of 1 KHz sinewave; scope DC coupled. With no control voltage applied, input approximates output. G (facing page)—Static DC voltage applied to control port first causes unipolar squashing, then complete loss of headroom as IC2's output is driven to its positive limit (H).
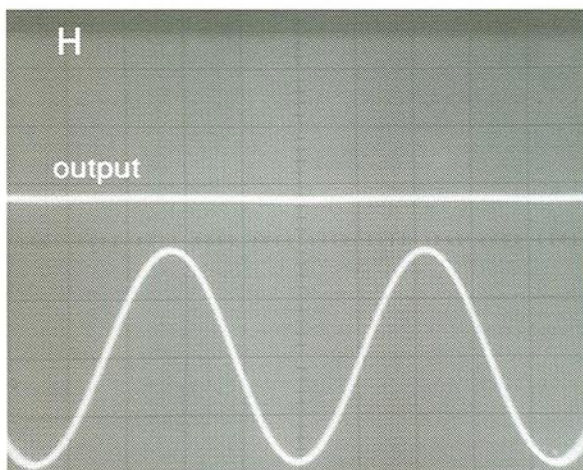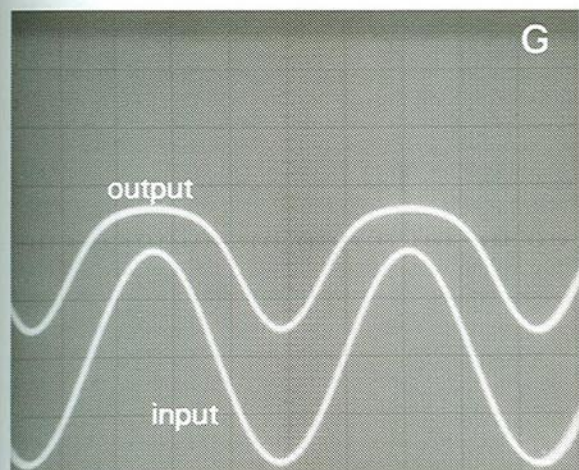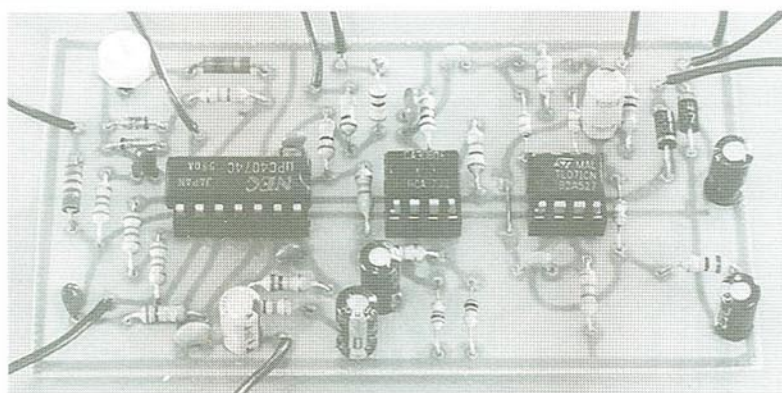
Fig. 29–2. Tremolo-Matic IV prototype board.
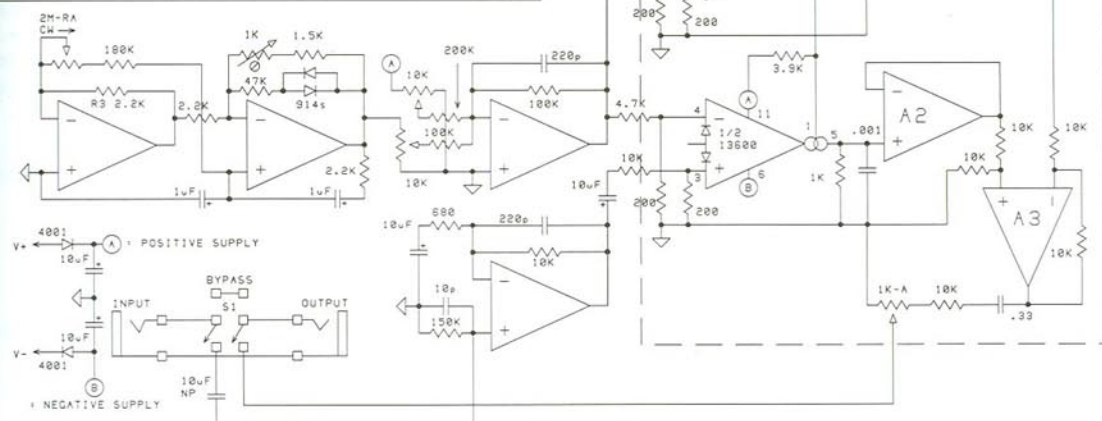
classic tube amp tremolos. This box walked away with an easy tie in a head-to-head shootout with a vinyl copy of "Harlem Nocturne," whose nasty rhythm guitar is an archetype of tube amp tremolo. VCA-based tremolos don't duplicate this sound because they don't distort.

The unorthodox approach saddles the system with heavy feedthrough, ~2V_{p-p} coming off IC2 at maximum depth. Four-pole highpass filtering and 20 dB of quasi-companding make the box as quiet as the average tremolo.

The nature of the modulator makes an alternative means to suppress feedthrough. An identical OTA running in parallel with the main OTA—but receiving only the control voltage—puts out a pure feedthrough signal; that voltage subtracted from the main OTA output leaves pure audio. Fig. 29–3 shows one incarnation of this approach; 100-ohm pot trims feedthrough.

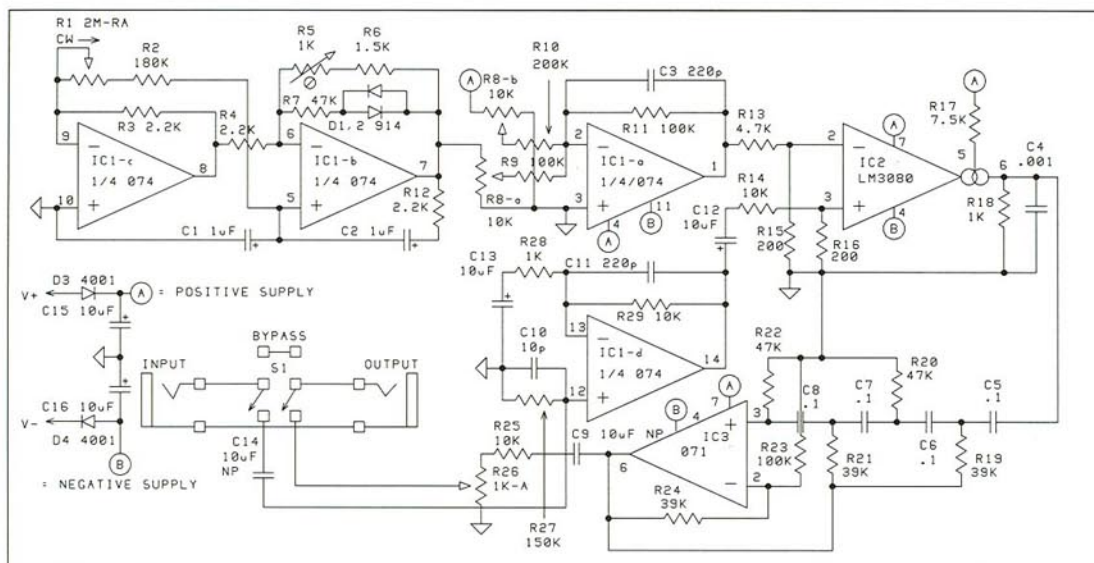Fig. 29–3. Circuitry inside dashed line shows a means to cancel feedthrough.

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Fig. 29–4. Tremolo-Matic IV schematic.

4.25" × 2.5" reference box
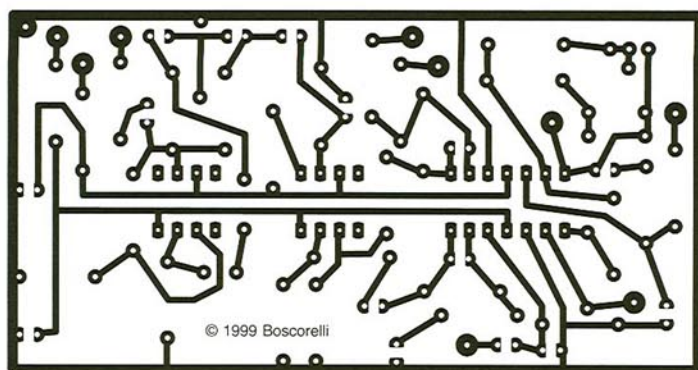
© 1999 Boscorelli

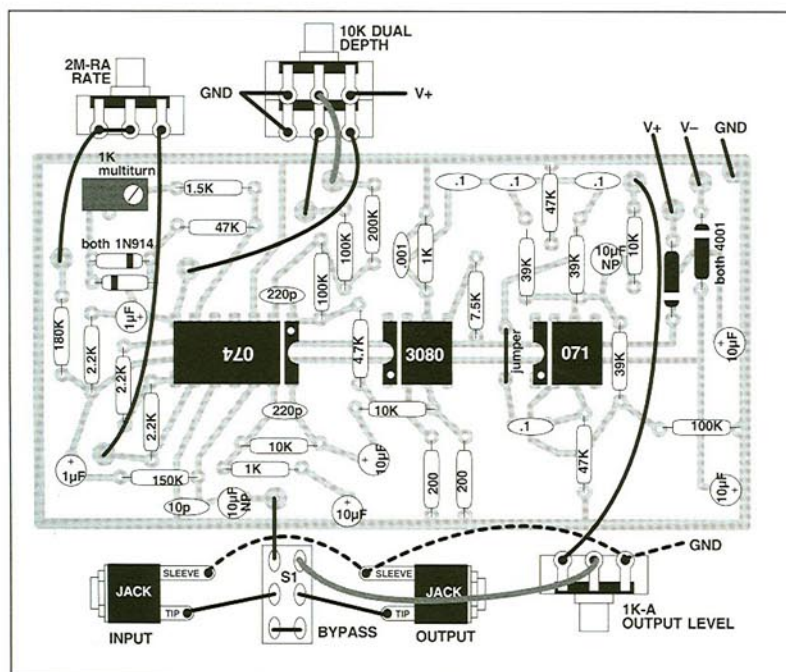Fig. 29–5 Tremolo-Matic IV circuit board.

Fig. 29–6. Tremolo-Matic IV layout & wiring diagram.

Phase Effects

Phase underlies several stomp-box effects. The commonest phase-altering blocks are tone controls, which shift phase, besides altering amplitude. Blocks that shift phase without altering amplitude are known as allpass filters (APFs), such as those shown in Fig. A55. APFs come in *lead* versions that push output phase ahead of input phase, and *lag* versions that shift output phase behind input phase. A single APF achieves nearly 180° of phase shift. A limited band of frequencies is affected by a single APF for a given value of C1, such that musically suitable circuits cascade stages.

Statically shifted phase gives a non-effect in mono, but owns the distinction of having underlaid the two major quadraphonic encoding systems in the seventies, and currently sees use in certain types of stereo synthesizers. Stereo phase shift keys on one way hearing localizes sound. A listener centered between two small speakers eight feet apart and eight feet away hears a single sound source placed between the speakers when a balanced, monophonic signal is played (Fig. A57). This is known as a centered *image*. As phase between channels shifts, the image first broadens, then becomes diffuse and impossible to pinpoint.

Rhythmically shifted phase manifests as a move in pitch apparent only while phase is changing. The mechanism compares to an electronic emulation of the Doppler effect. The result sounds distinct from tremolo and is often called *pitch vibrato*. Guitar figures in "Wheels" (the String-A-Longs, 1961) and the Beatles' "It's Only Love" typify the form. *Stereo vibrato* connotes a system in which each channel's phase moves in a direction opposite the other's. Both

Fig. A55 (right). APFs have unity gain, but alter phase of output, depending on setting of 250K pot. 1—Op-amp–based lead version moves phase of output ahead of input. 2—Lag version causes output phase to trail input phase. Each stage approaches 180° of shift; 0.01 µF cap value is a good compromise; smaller values favor higher frequencies, larger values favor lower tones. Practical embodiments cascade stages and place the effect under voltage control. APFs can use inductors in place of caps. 3 & 4—An alternative phase-shift method starts with transistors configured as phase inverters; signal at collector/drain is antiphase with signal at emitter/source. The juncture of 250K pot and 0.01 µF cap is a simple summing junction; as pot value is reduced from 250K to 0K, the percentage of noninverted signal rises. Due to the way matched sinewaves sum, this manifests as smooth phase shift from 180° to 0°.
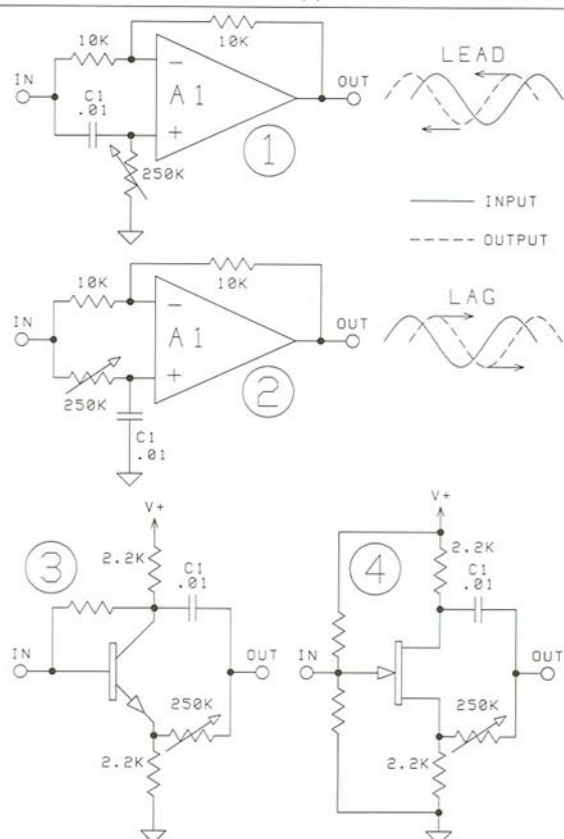
Fig. A56. Block diagrams show how phase shift can be used to alter frequency response. Given a single APF, only one output frequency approaches 180° of phase shift relative to input. Subtraction of a negative number equals addition, creating a peak at that frequency. Addition creates a notch because equal and opposite quantities cancel when added. Phase shifts accumulate in cascaded APFs, placing several frequencies 180° out of phase, giving multiple notches or peaks.
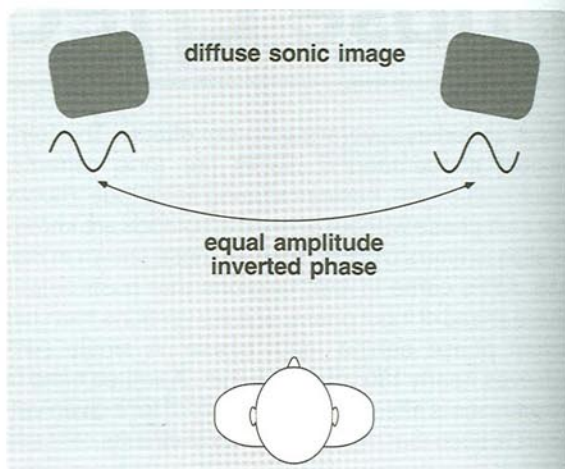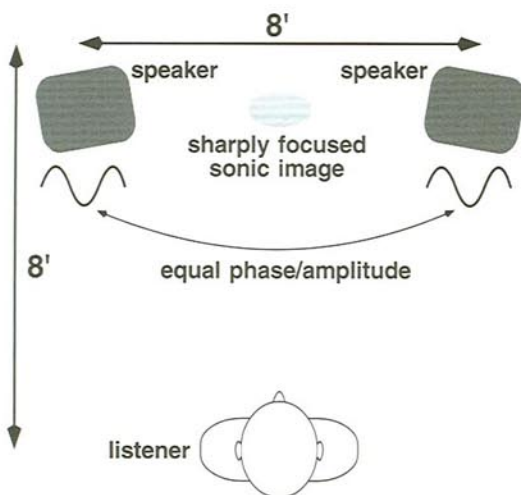

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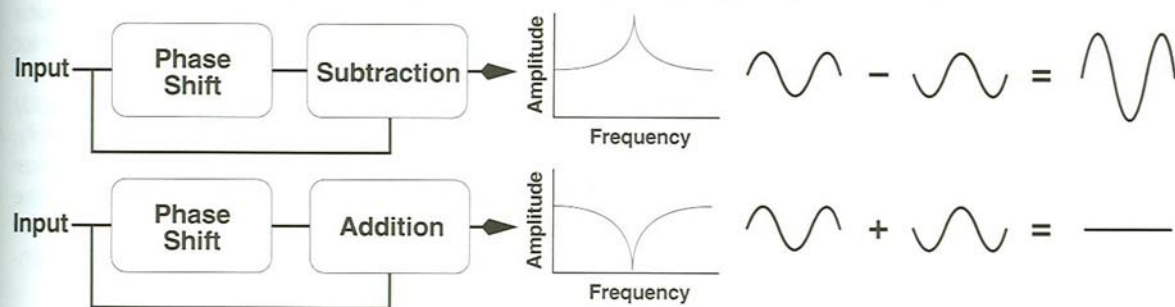
Fig. A57. Effect of phase on sonic image. If listener is centered between stereo pair in-phase mono, sound seems to originate from a point halfway between them. If inverted, sonic image becomes diffuse and impossible to localize; 'phasey'.
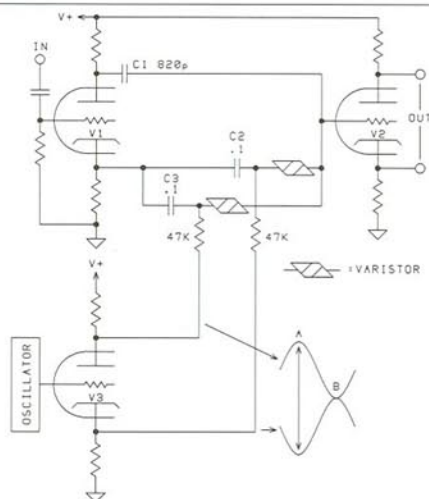

Fig. A58. Basics of the phase-shift block found in the Magnatone vibrato circuit, including stereo version of the legendary Model 280-A amp. V1 splits input signal into inverted and noninverted feeds. Inverted signal couples to V2 through C1; noninverted version couples to C2 & C3, each of which ties through a varistor to V2's grid. The capacitor-end of each varistor ties through a 47K resistor to one output of phase splitter V3, driven by the vibrato oscillator. When a large voltage difference exists between V3's plate and cathode (point A), the resistance of both varistors drops; phase at V2's input moves toward 0°. When a small voltage difference exists between V3's plate and cathode (point B), both varistors act as very large resistances; phase at V2's input moves toward 180°. The result of this process is smooth phase shift with relatively little amplitude change. The Model 280-A amp used two of these networks in series for each channel, but one channel's varistors received a version of the oscillator signal inverted compared to the other channel's. One channel's pitch rose as the other's fell. But for the varistor mechanism, the phase shift technique is identical to that of Fig. A55, circuits 3 and 4. Other Magnatone amps varied the specifics of phase shift, but most used varistors.

pitch and sonic image shift with each cycle. Magnatone® amps gained renown for the sound of their stereo vibrato (Fig. A58).

Like tremolo, vibrato holds hidden potential. The typical vibrato circuit cascades phase shift networks, each using a $0.01\mu F$ cap. This cap defines the band in which phase shift is most prominent, corresponding to low frequencies in guitar. Treble, and especially treble harmonics, get the short shrift. Lowering the cap values raises the frequencies emphasized (cap values should not be less than $0.0015\mu F$). Choosing different cap values for each phase shifter gives vibrato a shimmering quality, less forceful but more musical than the standard $0.01\mu F$ cascade.

Pedals with *phase* in the name distinguish themselves from vibrato by summing or subtracting straight and shifted signals. Addition of signals 180° out of phase results in a notch whose depth depends on amplitude match between the clean and shifted feeds. Subtraction of antiphase signals equals addition, producing a peak. These measures gain greater impact in cascaded APFs due to cumulative phase shift, resulting in multiple notches or peaks. Commercial phase pedals sport four or more APFs in series. Feeding back an in-phase portion of the shifted signal is known as regeneration. This introduces an active-filter aspect to the system, leading to oscillation if excessive.

Phase shift finds use in distortion boxes by virtue of shifting the transition points in the conversion of sine to square; and in sonic 'maximizers' and 'exciters' for as-yet obscure psychoacoustic reasons.

Phase cancellation and reinforcement accompany the physical interaction of sound sources, notably, that which occurs in front of stacked 4×12 speaker cabinets. Resultant comb-filtering contributes to the distinctive stack sound. Stack simulators reproduce this using phase-shift networks, or through conventional frequency manipulation.

Project No. 30
Vibrato-Matic

True stereo vibrato, à la Magnatone, plus a couple of twists to give the box greater warmth and versatility.

Circuit Function

Signal path: Signal input couples through C7 to IC1, a noninverting preamp with gain of 23. Preamp output couples to input of phase-shift block 1, and to one throw of 3PDT switch S2-a, whose pole ties to signal input of phase-shift block 2, and whose other throw ties to the output of phase-shift block 1. Each phase-shift block feeds an identical quasi-18 dB/octave high-pass filter that cuts off below ~60 Hz. Output of filter 1 couples through C3 to one throw of S2-c, whose pole couples to the output path through R21-R22 and whose other throw ties to the juncture of C4-R23. Output of filter 2 couples through C4-R23 to output level control pot R22.

Control path: IC3-c/-d and associated components form a sinewave oscillator whose rate varies over the approximate range 1–10 Hz by pot R1. IC3-d output couples to pot R8, which controls the sinewave level feeding inverting buffers IC3-a and IC2-a/-b. R15 controls the variable DC offset to the outputs of IC2-a/-b. The sinewave at IC2-a is inverted compared to IC2-b. IC2-a output couples through R19 to all three phase shifters in block 1; IC2-b output couples to one throw of S2-a, whose other throw ties to output of IC2-a and whose pole ties to R20, which feeds the control voltage to phase shift block 2.

When S2 selects stereo mode, the signal passes independently through phase shift block 1 and phase shift block 2; each block receives a control voltage with identical DC offset, but with sinewave polarity inverted relative to the other. This causes phase to move in opposite directions between the phase-shift blocks. When S2 selects mono mode, three changes occur: (1) both phase shift blocks receive an identical control voltage; (2) the signal passes first through block 1 then through block 2, a total of six phase shift networks; (3) both outputs get the same signal.

Use

Pots & switches have these functions:

R1	rate
R7	sine trim
R8	depth
R15	static point
R22	output level (both channels)
S1	effect/bypass
S2	mode (mono/stereo)

First, trim the sine generator. Connect scope probe to IC3 pin 14; set scope for DC coupling; power up circuit, trim R7 to give a sinewave of ~10V_{p-p}.
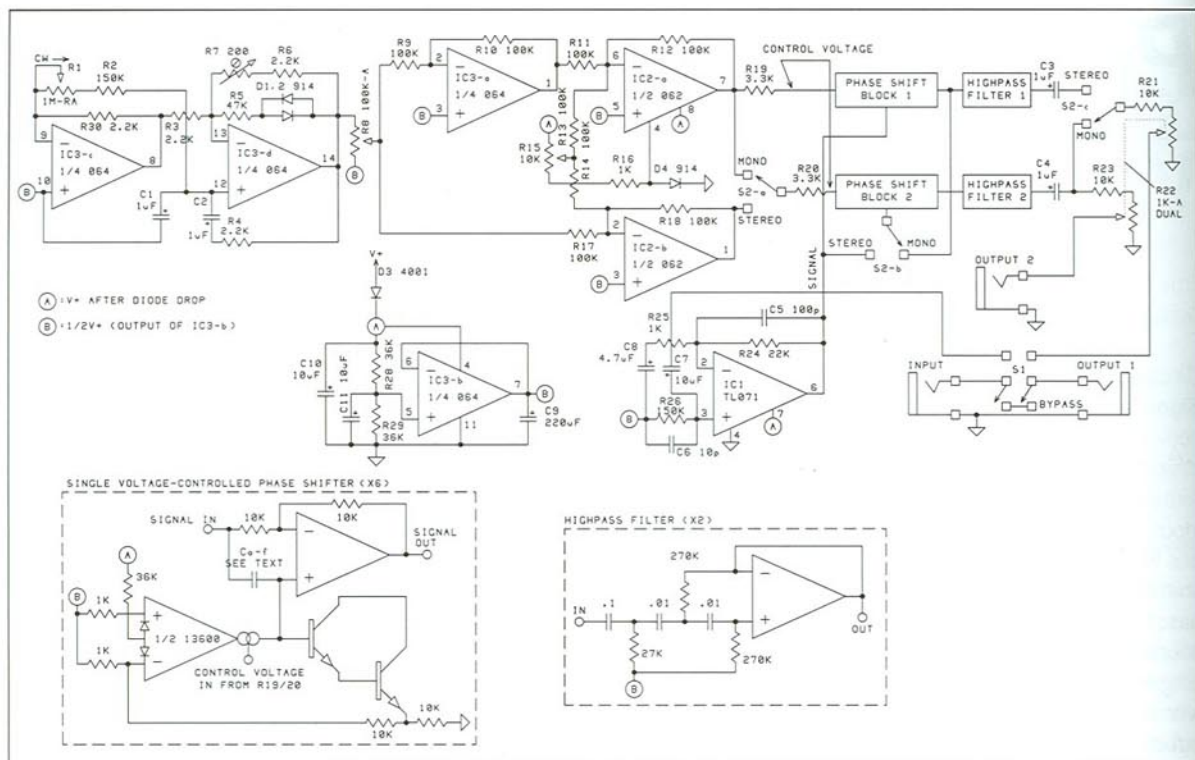
Fig. 30–1. Vibrato-Matic schematic.

Initial settings: R8, straight up, R1 fully CW, R15 9 o'clock, output level max, effect in, mode stereo. Connect unit to axe and two amps (a home stereo may be easier to manage at this point); set volume of each amp for equal level. Obvious vibrato should be heard; if it isn't, tune R15 until you hear vibrato. Vary rate, depth, and static point, while noting changes in sound. Switch S2 to mono mode; note greater vibrato depth, at a given setting, compared to stereo mode.

Notes

Vibrato-Matic is optimized for a 15V supply; significant departure from this voltage is not recommended. The prototype ran fine off a pair of alkaline 9V batteries.

In stereo mode, Vibrato-Matic acts as a pair of three-stage phase shifters, each receiving a control feed inverted relative to the other, and each affecting a different band of frequencies, due to variegated values of phase-shift caps. Switching to mono places all six phase shifters in series and feeds the same signal to both outputs.

Capacitors Cabc make up phase-shift block 1; caps Cdef make up block 2. The prototype used these values:

Ca	0.01μF
Cb	0.0068μF
Cc	0.0056μF
Cd	0.0047μF
Ce	0.0033μF
Cf	0.0022μF

This box delivers a vast range of vibrato, from subliminal "Day Tripper" levels to sci-fi sound effect at maximum depth. Variegated caps create multiple overlapping bands of phase-shift that treat different frequencies to different levels of vibrato. The sound changes as the player moves up and down the fretboard. Wide speaker separation accentuates the stereo mode.

The box blunts feedthrough with highpass filtering, ~20 dB of quasi-companding, and limiting the negative swing of the control feed (the purpose of D4; compare Phase-O-Matic's control path, which omits this diode). Also to this end, IC2 should be a TL062 or other dual op amp having the same negative output limit.

The prototype configured the preamp for gain of 23 to suit a specific guitar. Other axes may need more or less gain. Preamp output should average at least 3Vp–p, to leave room for peaks while accommodating the circuit's quasi-companding.

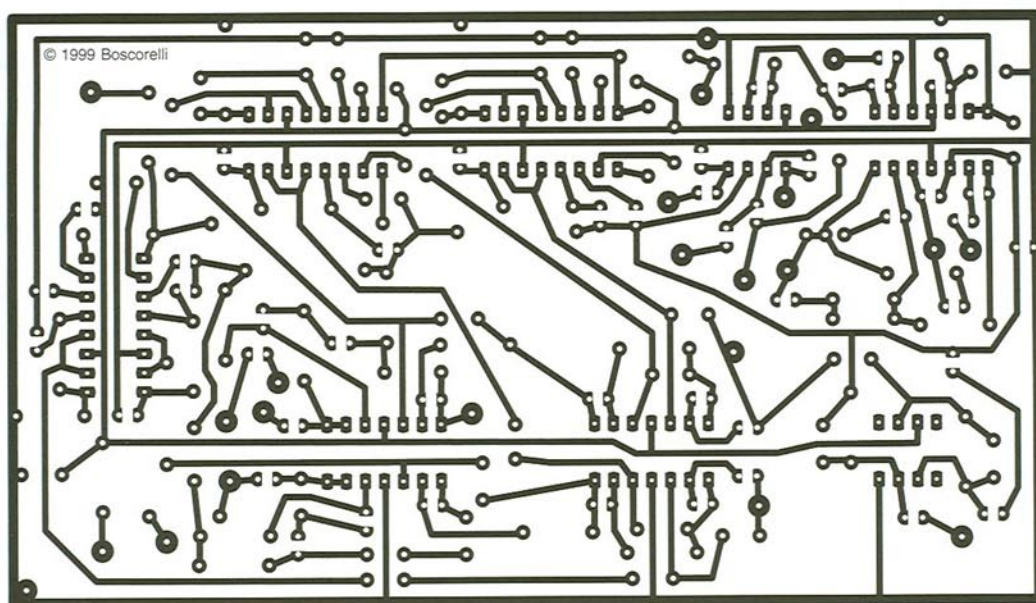Fig. 30–2. Vibrato-Matic prototype board.
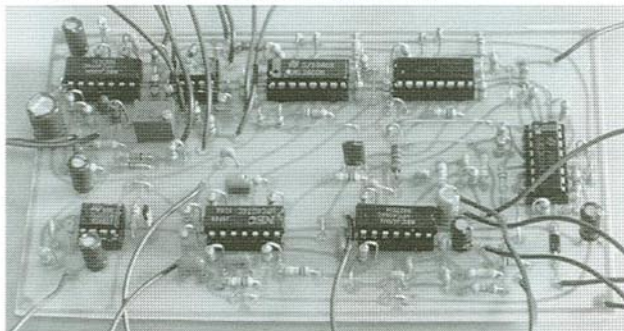
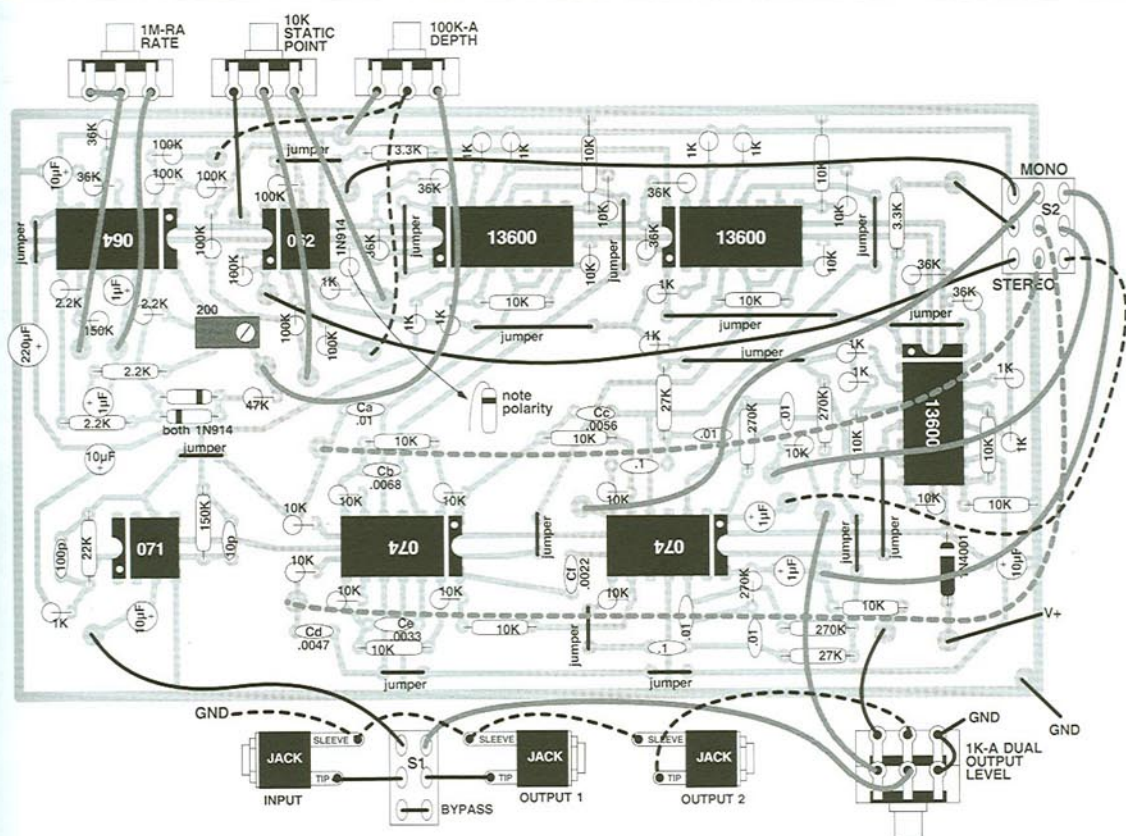6.25" × 3.75" reference box

Fig. 30–3. Vibrato-Matic circuit board.

Fig. 30–4. Vibrato-Matic layout & wiring diagram.

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Delay Effects

The ability to delay an audio signal enables several stomp-box effects. *Echo* defines discrete, individually perceptible repeats. *Reverberation* describes a confluence of echoes, early and late. *Chorus* is meant to simulate the slight asynchrony of two instruments playing at the same time. *Flanging* attempts to duplicate the effect of two identical analog tapes played simultaneously, but with a slight, changing delay imparted by briefly dragging one reel's flange. The resultant shifting delay creates a distinctive moving *whoosh* (check out your grandparents' 45 single of Miss Toni Fisher singing "The Big Hurt"). Delay modes available to the stomp-box builder include bucket brigade devices, springs, and analog tape.

Bucket Brigade Devices

BBDs delay audio signals by handing them off between capacitors linked by MOSFET transistors (Fig. A59–1). Transfer takes place stepwise, under direction of a *clock*, a tunable squarewave source covering ~10–200 KHz. The lower the frequency, the longer the delay. Total delay depends on clock rate and number

Fig. A59. 1—A bucket brigade device consists of a chain of capacitors linked by MOSFET transistors. Clock 1 and Clock 2 are logical inverses; when one is HIGH the other is LOW. When Clock 2 is HIGH, Q1 turns ON, and signal present at the input flows into C1. It cannot get through Q2 because Clock 1 is LOW, keeping Q2 turned OFF. When Clock 2 goes LOW, Clock 1 goes HIGH; signal stored in C1 flows into C2. Process repeats for all stages. Time between clock cycles delays transfer. Odd-numbered transistors' gates all see +1V. Signal is recovered by summing network in a way that partially cancels clock impulses. The 5.6K resistors can be replaced with a 10K pot to optimize feedthrough trim. Figure adapted from manufacturer's diagram, and applies to the MN3009; other chips in the MN series work on the same principle, but have slightly different internal configurations. 2—Schematic illustrates general measures required to implement most of Panasonic's MN series of BBDs. Signal couples through buffer A1 to anti-aliasing lowpass filter A2. A1 and A2 are biased off A3. Important to note that filters and buffers run off ½V+; MN3009 is biased off 50K trimpot. BBD's bias voltage is critical to least distortion and may not fall exactly on ½V+. Clock generator IC2 supplies two inversely related squarewave trains @ ~10–200 KHz; R2 determines rate, and thus delay. BBD output is taken at juncture of 5.6K resistors, feeds lowpass filter A4 to remove clock artifacts. The MN3009 is a 256-stage device capable of supplying delay between 0.64 and 12.8 milliseconds. Choose other chips for longer delays. Specific applications may demand sharper lowpass filtering ahead of and/or after delay to realize desired performance.
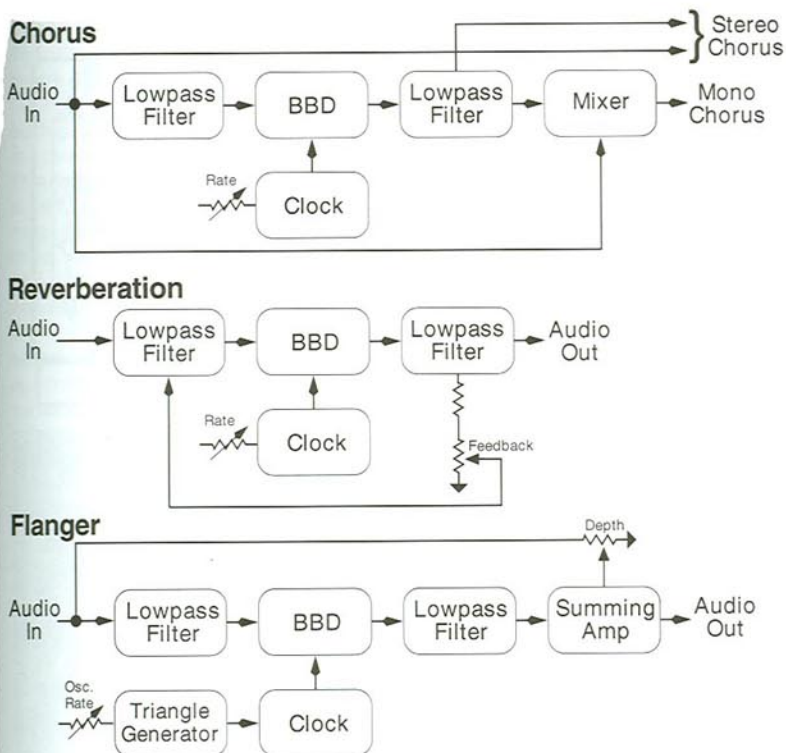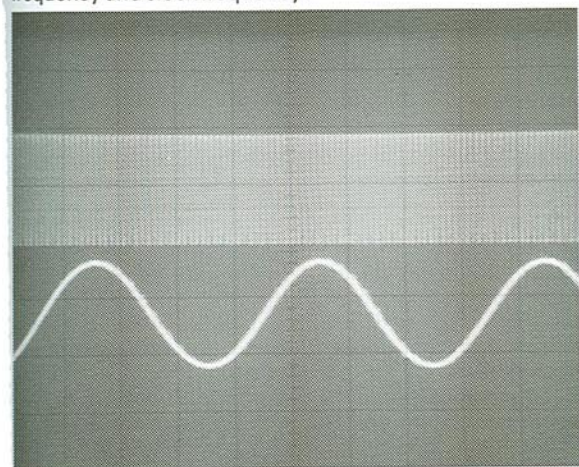
Chorus

Reverberation

Flanger
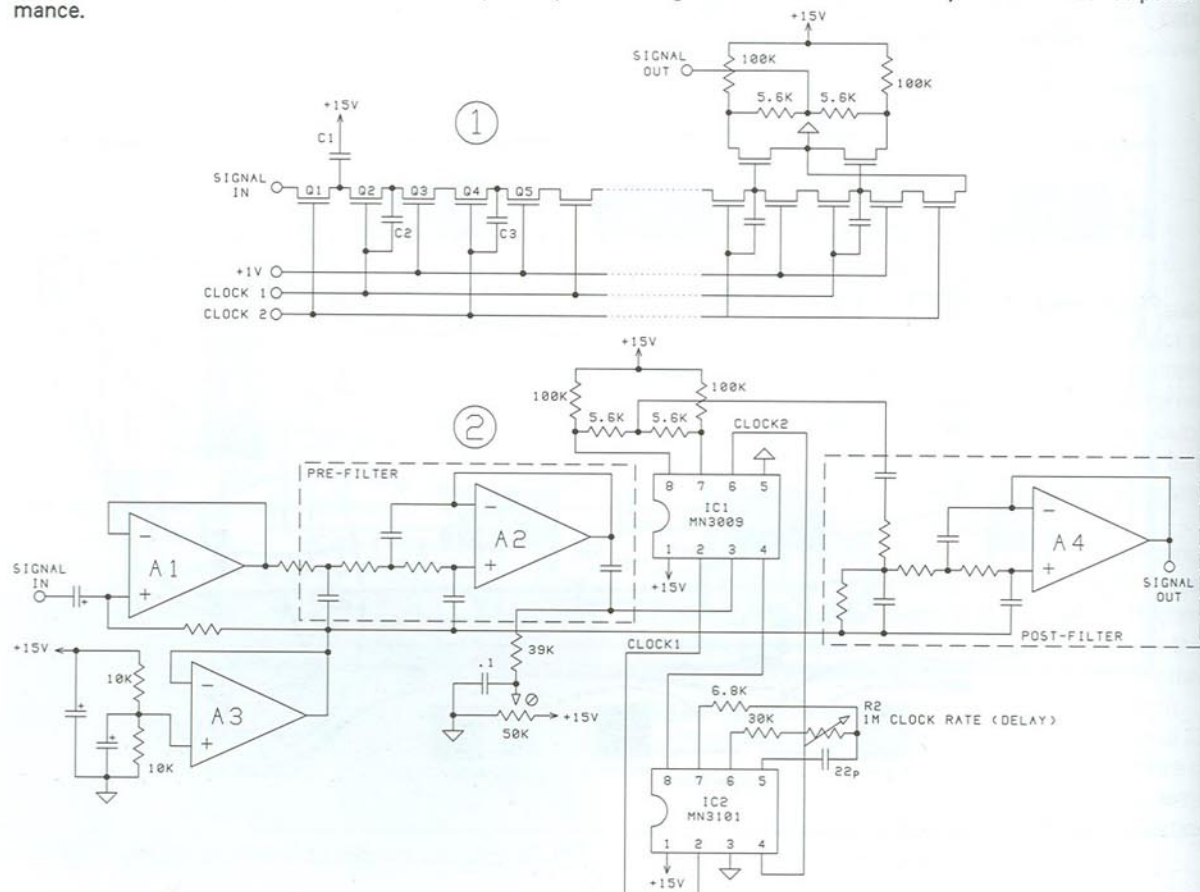
Fig. A60. Basic BBD block consists of lowpass filter ahead of BBD to prevent aliasing; BBD, whose delay depends upon clock rate; and lowpass filter after BBD to remove clock feedthrough artifacts. Chorus is achieved by producing a delayed version of the signal. Mixing straight with delayed yields mono chorus; separate clean and delayed feeds give stereo chorus. Reverberation is produced by feeding back a variable amount of delayed signal to the input, creating a decaying loop. Flanging results from rhythmically varying phase and adding the resultant wet/dry signals. Phase shift due to delay causes cancellation at some frequencies; functional overlap with effects based on allpass filters.

of stages. BBD chips come with as few as 64 or as many as 4096 stages, giving <1 ms to >200 ms delay. The chip's output contains clock feedthrough artifacts. While not audible above 20 KHz, maximum delay clocks near 10 KHz. Stiff lowpass filtering is needed to blunt audible residue. In addition, clocked transfer makes a digital stigma that leaves BBDs subject to *aliasing*, the generation of spuriae when the system meets frequencies higher than half the clock frequency (Fig. A61). Lowpass filtering ahead of the input

Fig. A61. An example of aliasing. Circuit similar to Fig. A59–2, but using MN3007; clock frequency and input frequency are near 12 KHz. Top trace shows input frequency, bottom trace shows output frequency, a spurious tone generated by aliasing; frequency equals the difference between input frequency and clock frequency.

keeps out tones that might cause aliasing. Performance is further restricted by a rapid rise in distortion above an input level of ~1.5V_{p-p}. These idiosyncrasies make BBDs great candidates for internal companding.

BBD effects include chorus, in which straight and delayed feeds simulate two instruments playing slightly out of sync. The effect gains realism by the application a slight, random shift to the delayed feed, because humans picking in tandem cannot keep perfect sync. Stereo chorus attains greater impact by feeding wet and dry outputs to separate speakers.

BBDs recreate reverberation by feeding part of the delayed signal back to the BBD input to make a decaying loop. Character of the 'verb changes with duration of delay and percentage of feedback. More realistic reverb is achieved by BBDs that provide taps at different delay times, to simulate early and late returns.

Finally, delay alters phase, allowing BBDs to produce the crawling notches characteristic of phasing, and to give a facsimile of the flanging effect.

A rich hardware selection puts BBD effects within the builder's reach. Chips come in 5–10V and 9–15V families; 64 to 4096 stages. The 64- to 1096-stage chips can be had for less than $10; long-delay chips currently cost ~$40. Clocks come in dedicated chips, but are easily built from standard CMOS parts. Fig. A59–2 illustrates general considerations for using BBDs.

Spring 'Verb

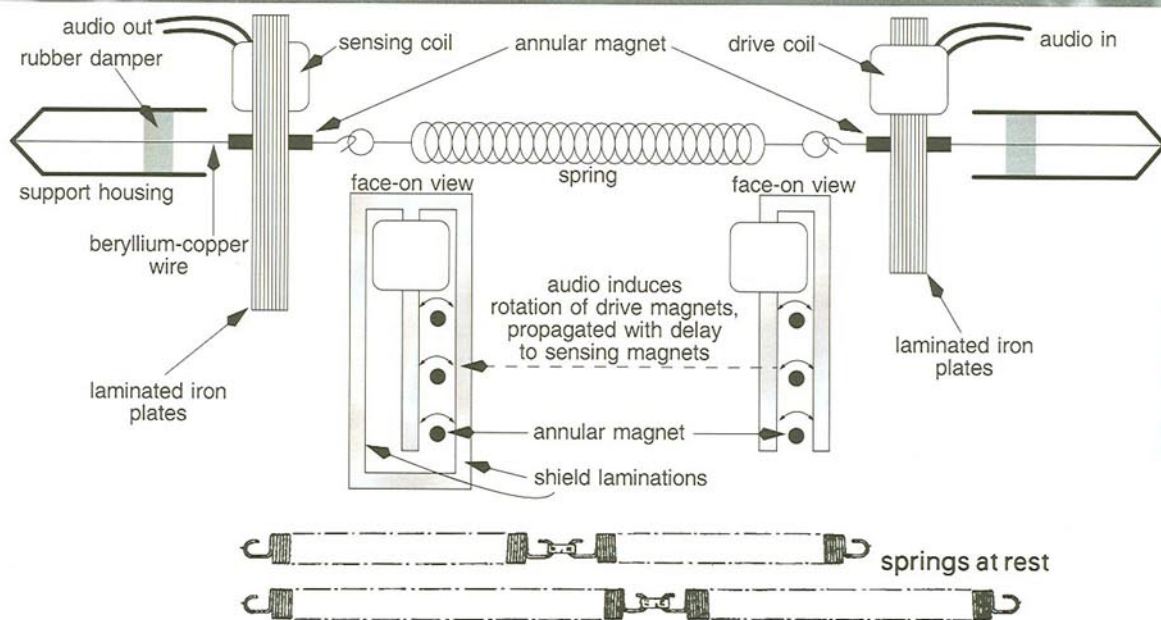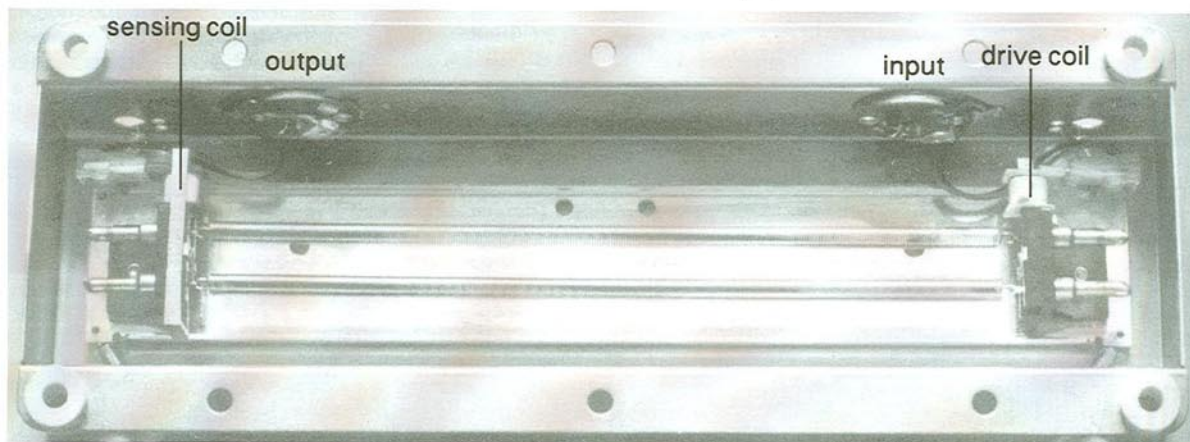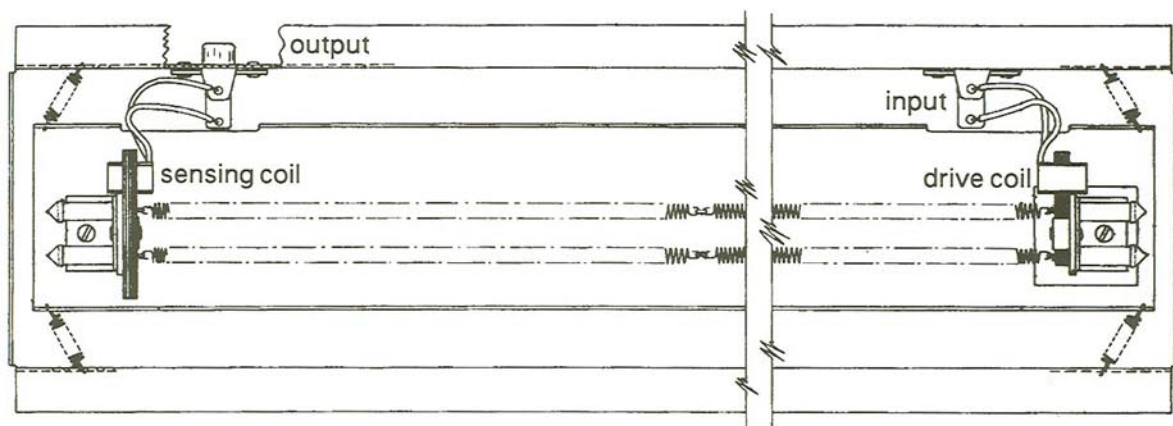Surf music is hard to imagine without spring reverb,

Fig. A62. Top—Diagram copied from Ref. 35, the patent whose design underlies most spring 'verb systems in guitar amps. Accutronics' short two-spring pan (photo) closely resembles patent diagram, but has unitary springs. System works thus: Audio feeds drive coil wound on U-shaped ferrous laminations. Field imparts rotary motion to magnets; torsion takes some tens of milliseconds to propagate to other end of spring; some energy rotates magnets near output coil, but most of the energy bounces back and forth through this complex system in a decaying smear that simulates reverberation. The magnets are supported at their outer ends by beryllium-copper wire, which offers very high compliance. Each wire passes through a rubber damper, to minimize lateral movement and absorb energy. The magnets, maybe $1/16$" diameter, are ceramic. If desired, the tube/holder block can be assembled, then subjected to a strong field that magnetizes the ceramic in situ, eliminating manufacturing hassles associated with small magnets. The springs, though stretched to equal length in the pan, differ in resting length. In pans that use segmented springs, one spring is wound CW, the other CCW, to cancel compressional mode that otherwise accompanies the torsional mode (bottom diagram copied from Ref. 35).

Fig. A63. Spring reverb designs prior to 1961. A & B— Springs suspended in a catenary arc. This design was significant for magnetic torsional drive, and coupling a CW-wound spring to one wound CCW. C—Shows springs suspended vertically, connected by a series of levers; drive mechanism can be magnetic or piezoelectric; bottom ends of springs may be immersed in damping fluid. Diagrams copied from Refs. 33 & 34.
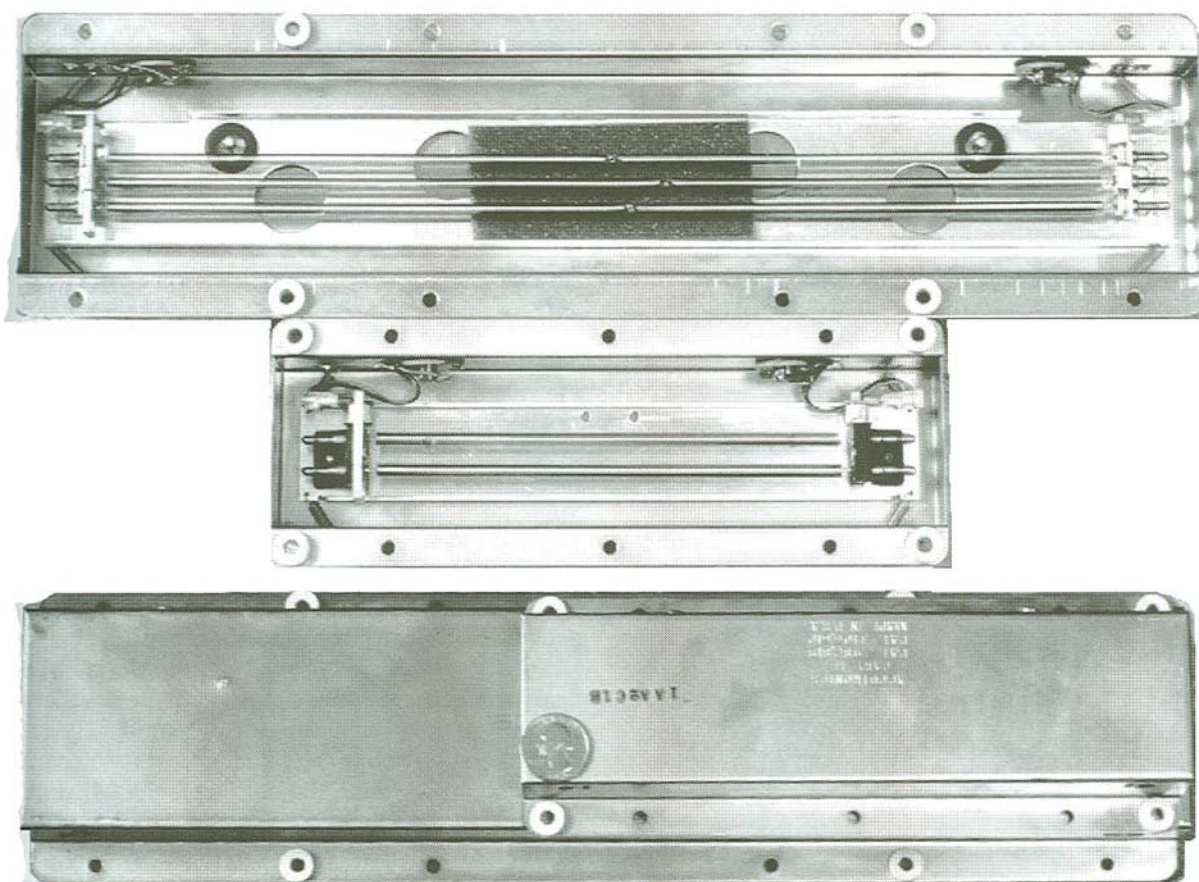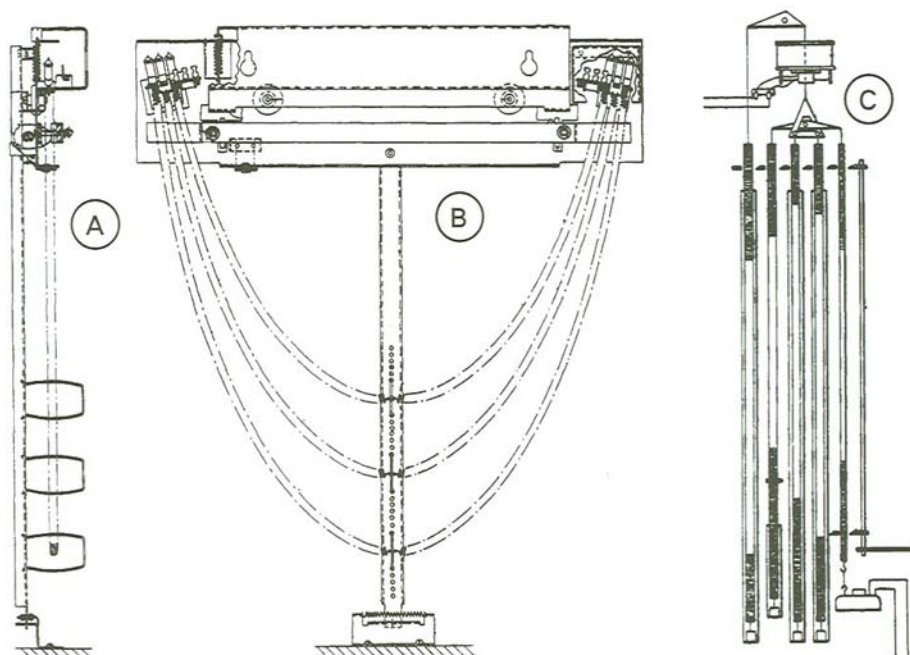

Fig. A64. Photos show short two-spring pan relative to long three-spring pan. Longer pan uses coupled springs and gives wider frequency response and greater delay, but at nearly 17" does not fit all enclosures.

Fig. A65. Close-ups of three-spring reverb sensing assembly (top) and drive assembly (bottom).

one of guitar's most alluring and enduring sounds. Per the name, the effect involves a system of springs tied to transducers, an idea that hearkens to the first half of the twentieth century. Early designs mounted springs vertically, either straight or in a catenary loop, and drove them with magnetic or piezoelectric transducers. Evolution of this approach culminated, in 1961, in the design that defined spring reverb in the form still used today. Indeed, modern hardware could have been blueprinted from the patent diagram (Fig. A62).

Like tube amplification, spring 'verb is outwardly simple, functionally complex. Audio energizes an electromagnet whose field moves an annular ceramic magnet tied to one end of a helical spring. The magnet is polarized such that the field imparts rotary motion, propagated torsionally through the spring to a second magnet at the other end. Rotation of the second magnet induces the output voltage in a second coil. This spring-magnet assembly is suspended horizontally, tied to supports at each end by high-compliance beryllium-copper wire. Mechanical impedance of the system is such that only part of the energy in the spring transfers to the output magnet. Much of it reflects within the spring, losing power with each bounce. Energy dissipates in a smear lasting up to several seconds, simulating reverberation. The setup mounts on a shallow metal pan, suspended by four springs from a larger pan. The entire assembly is sometimes called a tank.

Single-spring systems deliver bland reverb and add tonal emphasis. Two- and three-spring designs sound livelier and impart less tonal emphasis, especially when the springs offer different primary delay times (37 and 29 milliseconds are typical). Close inspection of springs shows them to differ in wire diameter, or coil diameter, or number of turns per inch. In pans in which the springs are not glued to the magnets, decoupling shows them to differ in resting length. All these differences represent means to achieve unequal propagation time. Also, in designs that couple two springs, one spring is wound clockwise, one counterclockwise. The purpose of coupling anti-wound springs is to cancel the compressional propagation that results when a single spring is wound or unwound.

A complete reverb system consists of a driver to couple audio to the springs; an amplifier to boost the output of the pan; and a mixer to vary the strength of the reverb signal (Fig. A66).

The sound of spring 'verb depends on the number of springs, their relative delay; the rigidity of their coupling to the magnets and, in series springs, to each other; the compliance of the support wire; the 'deadness' of the rubber dampers around the wire; and countless other minor factors. These variables are pretty much fixed in a given pan, though adventurous experimenters can tinker with drive coupling.

Spring reverb represents yet another of guitar's undiscovered effects. Amp-based 'verb offers a single control that mixes wet with dry. An outboard unit allows alteration of variables that change the sound. Drive level ('dwell' on the old Fender 6G15) varies strength of the signal sent to the springs. Muted strings played through hard-driven springs creates an artifact reminiscent of a pebble plopping in a well. The surf-punk school prizes this as much as the 'verb. The sound stands out on the Ventures' studio cut of "Journey to the Stars," and scores of unfamed surf gems

Fig. A66. Basics of Leo's main reverb drive & mixing circuit. Signal feeding spring driver V2 rolls off below 320 Hz through action of 500 pF cap. Divider action of 3.3M with 220K attenuates dry signal; 100K pot varies percentage of wet. Output feeds directly to next triode stage. Circuit redrawn after Fender "Pro Reverb" amp, type AA165.
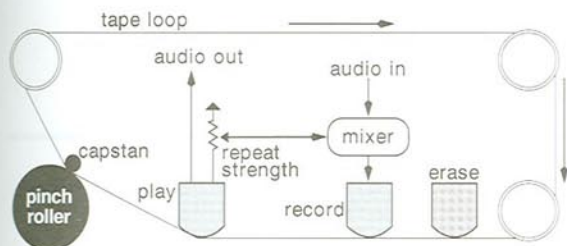
Fig. A67. Functional basis for true echo devices, such as Echoplex. Device is a modified analog tape recorder that stores a tape loop in a cartridge. Tape speed and/or distance between playback and recording heads can be changed to shift repeat rate; strength of signal sampled off play head and fed back to recording head determines echo strength and number of repeats.

Spring 'verb also responds smartly to altered drive tone, and to tone shaping applied to the output of the pan.

Reverb pans come in two-and three-spring models, short- and long-spring designs, and with low- or medium-impedance inputs. Stout op amps drive the medium impedance, but the 8-ohm types present a load better served by audio power chips (LM380, LM386, etc.). The classic circuit found in Fender amps drives the pan off a step-down transformer. Under average drive conditions a pan's output measures <10 mv, requiring ~40 dB of boost before mixing.

A tone sweep of a typical spring pan reveals numerous resonances, with response rolling off below 80 Hz and above 3 KHz. Soundwise, the long three-spring pan gives longer decay and somewhat brighter tone, but the short two-spring model acquits itself well; a perfectly good option where space is tight.

Echo

Echo, in the limited sense, means a single repeat. Analog echoes are simulated by a tape loop passing over a recording head and at least one playback head. The system feeds a variable portion of the output back to the recording head, making a decaying repeat whose rate varies with tape speed, or with distance between recording and playback heads. The genesis of this effect traces to the Echoplex, a product that became synonymous with its sound. The builder with access to a three-head tape deck can achieve true echo by supplying an external path to feed a variable amount of output signal back to the recording head.

Project No. 31

'Verb-O-Matic

Entry-level spring 'verb on nine volts.

Circuit Function

Instrument feed couples through C11 to noninverting preamp IC1-a, whose gain is fixed at 11. IC1-a output couples to R18, which varies signal level feeding IC2, an LM386 audio power driver whose output drives the low-impedance input of a reverb pan, whose output couples through C5 to IC1-b, a noninverting amp whose gain is fixed at 101. IC1-b output couples through R11 to divider/mixer R5-R6-R7. Signal couples through C9 to the output path.

Use

Pots & switch have these functions:

R11	reverb mix
R18	spring drive level ('dwell')
S1	reverb/bypass

Initial settings: S1 reverb; R11 fully CCW, R18 straight up. Connect unit to spring pan, axe, and amp. Establish desired listening level, then turn R11 CW and note the increasing percentage of reverb. Test various combinations of dwell and reverb mix.

Notes

VOM is designed to drive low-impedance pans. It gave excellent results with Accutronics' short two-spring pan and long three-spring pan, each having an input impedance of ~8 ohms. A pan with a medium-impedance input gave acceptable performance, but this improved significantly with the interposition of an 8:500-ohm step-up transformer (e.g., Mouser p/n 42TU400) between VOM output and pan input.

VERB-O-MATIC PARTS LIST

Resistors
R1, 2, 4, 8, 14 10K
R3 1K
R5 100K
R6 470K
R7 220K
R9 100
R10 47K
R11 100K pot
R12 150K
R13, 16, 17 10 ohms
R15 470
R18 10K audio-taper pot

Capacitors
C1 220pF
C2, 3, 13 220μF aluminum electrolytic
C4, 15 0.1μF
C5, 8, 9, 11, 12, 16 10μF aluminum electrolytic
C6 0.001μF
C7, 10 100pF
C14 10pF

Semiconductors
D1 1N4001
IC1 MC33172 dual op amp
IC2 LM386 audio power driver

Miscellaneous
reverb spring pan (see text)
S1 DPDT switch
circuit board, solder, 9V battery snap, jacks, etc.

3" × 2" reference box

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Fig. 31–1. 'Verb-O-Matic circuit board.
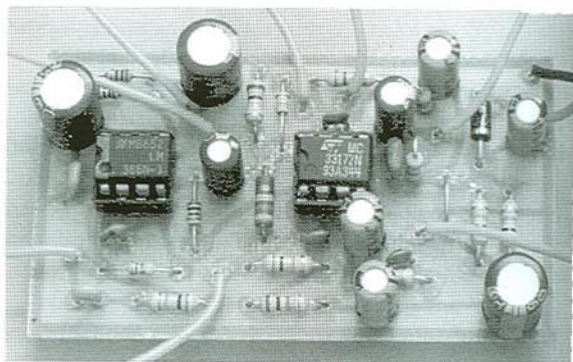
Fig. 31–2. 'Verb-O-Matic prototype board.
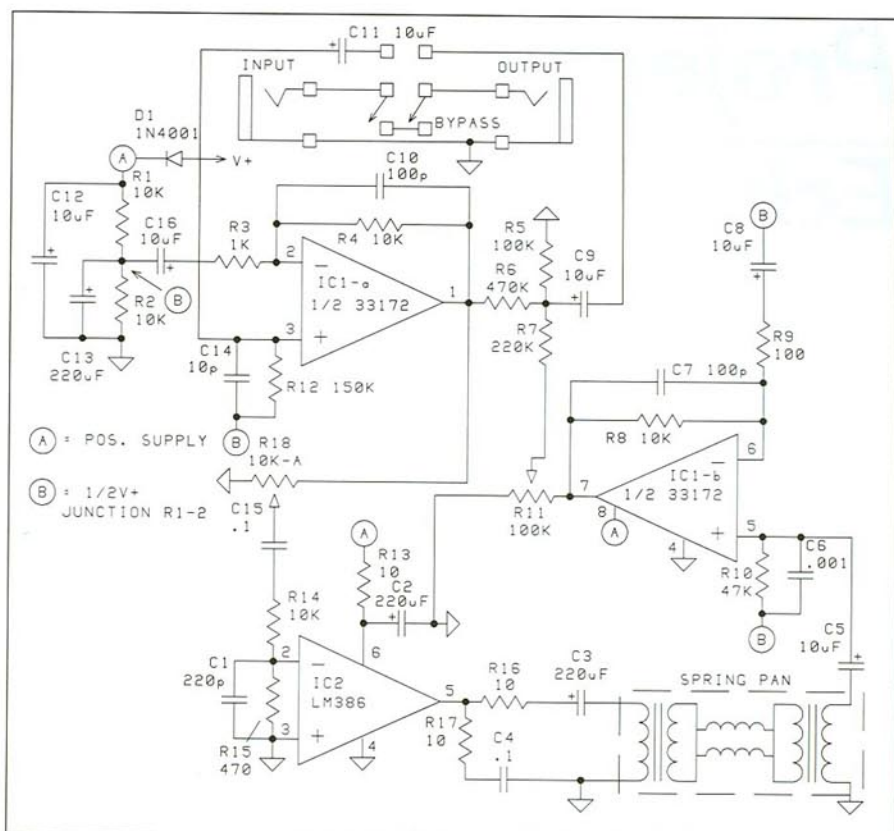
Fig. 31–3. 'Verb-O-Matic schematic.
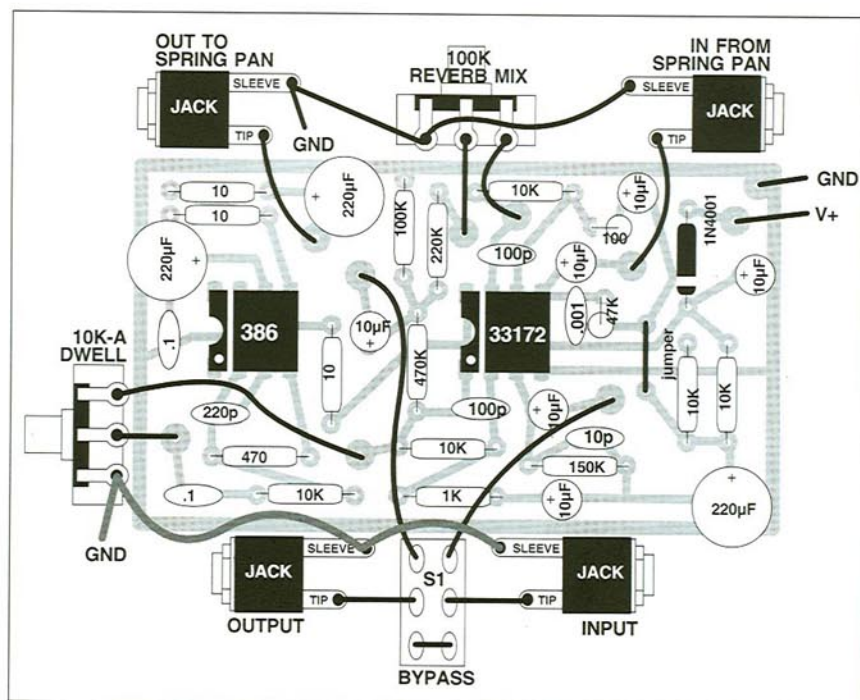
Fig. 31–4. 'Verb-O-Matic layout & wiring diagram.

Project No. 32
Echo-Matic

Can't afford an Echoplex®? Echo-Matic and a three-head tape deck get you into action for less than $10 in parts.

Circuit Function

Instrument feed couples through C1 to inverting pre-amp IC1-a, whose gain is fixed at 10. IC1-a output feeds unity-gain inverting buffer/mixer IC1-b, whose output couples through R5-C4 to the input of a three-head tape deck. IC1-a output also couples through R14 to unity-gain inverting buffer/mixer IC1-d. Tape deck monitor output couples through C5 to voltage follower IC1-c, whose output feeds pot R10, whose setting determines the number of repeats by controlling the amount of feedback to summing amp IC1-b. IC1-c output also feed pot R11, whose setting determines the strength of the wet (echo) signal feeding output mixer IC1-d. IC1-d output couples through C12 to pot R16, which varies the output level. Divider R7-R8 provides a ½V+ reference.

Use

Pots & switch have these functions:

R10	repeats
R11	straight/echo mix
R16	output level
S1	echo/bypass

Initial settings: S1 echo; R10, R11 fully CCW; R16 centered. In this state the box acts as a preamp with gain of ~2. Connect unit to three-head tape deck and amp as shown in Fig. A32–3 (left diagram). Switch the tape deck to record while monitoring the playback head. Roll tape; use the deck's meter to establish a recording level that does not exceed 0 dB. After establishing desired listening level, slowly turn R11 clockwise and note the increasingly strong single echo. Next, turn R10 clockwise and note the greater number of repeats. Depending on the recording level, these may reach the point of feedback and saturation.

The prototype was tested with a two-speed open-reel deck (3.75 & 7.5 ips). The results embraced everything from "Be-Bop-A-Lula" to outtakes from *(The) Ventures in Space*. Using the deck's built-in dbx Type I noise reduction gave a much quieter system, but lost some of the classic tape echo sound. To lengthen the repeat time, wire both channels in series (Fig. A32–3, right diagram).

Echo-Matic functions over the range 3–36V. The prototype drew ~1.7 ma running off a 9V nicad.

ECHO-MATIC PARTS LIST

Resistors
R1, 7, 8 10K
R2, 3, 4, 13 100K
R5 100
R6 47K
R9 1.5K
R10 20K pot
R11, 16 10K audio-taper pot
R12, 14, 15 22K

Capacitors
C1, 4, 5, 7, 9, 10, 12 10µF aluminum electroltyic
C2, 3, 6, 11 100pF
C8 220µF aluminum electrolytic

Semiconductors
D1 1N4001
IC1 MC33174 quad op amp

Miscellaneous
S1 DPDT switch
circuit board, solder, 9V battery snap, jacks, etc.
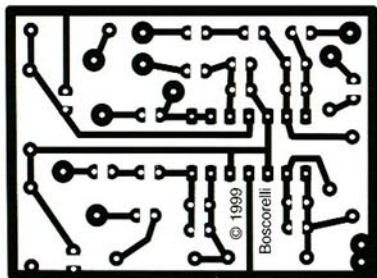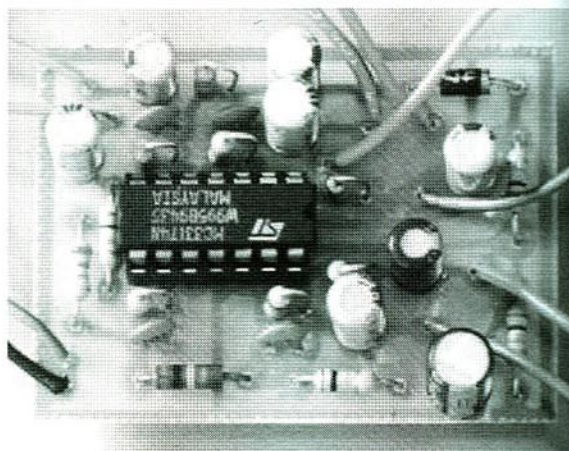

Fig. 32–1. Echo-Matic circuit board.

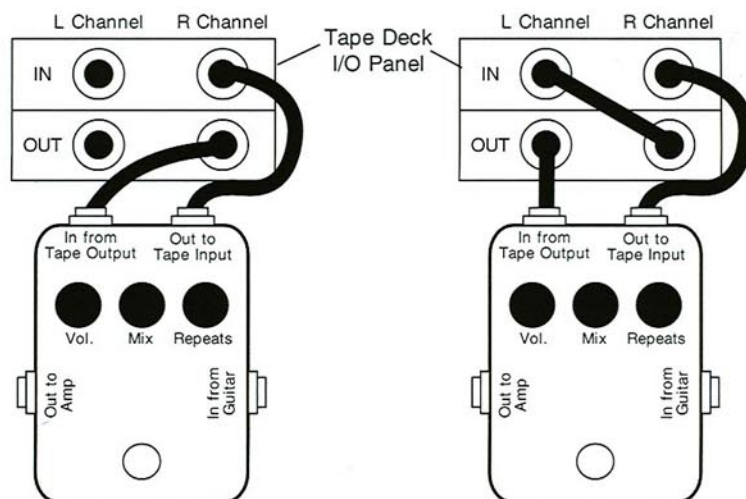Fig. 32–2. Echo-Matic prototype board.

Fig. 32–3. Two wiring options for Echo-Matic. Left diagram shows simplest; feed EOM's tape output to input of one channel; feed deck's output back into EOM. Right diagram shows sequential wiring of channels, giving longer delay.
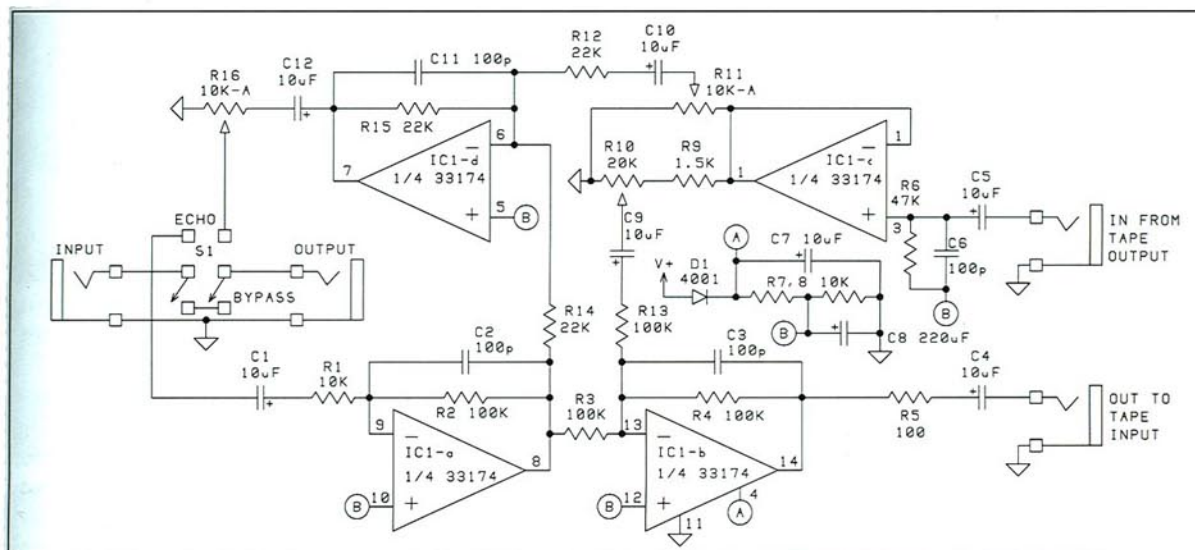

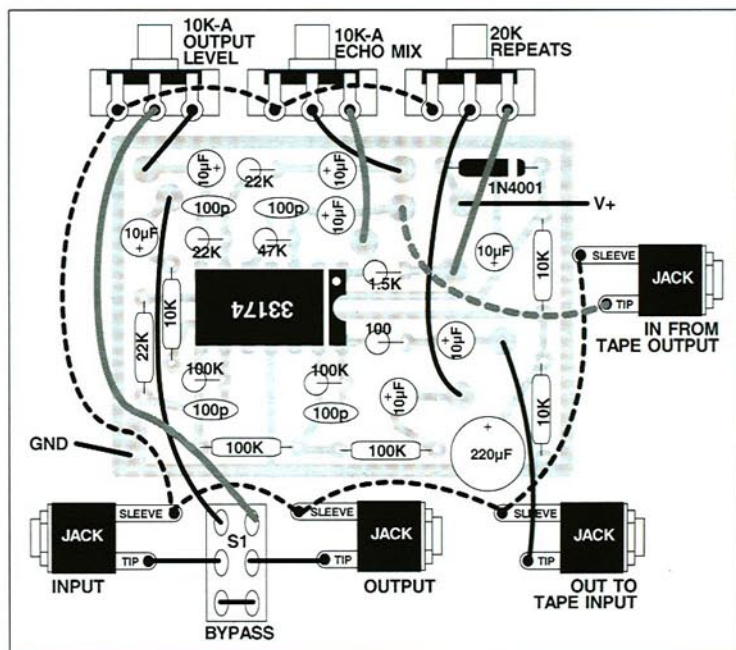Fig. 32–4. Echo-Matic schematic.

Fig. 32–5. Echo-Matic layout & wiring diagram.
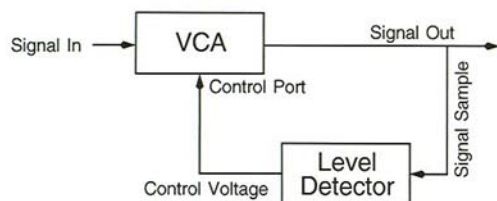
Dynamic Effects

Dynamic effects react to or track the changing level of a musical signal. The category includes compressors, expander-based and filter-based noise reducers, de-essers, and envelope-driven tone modifiers. Dynamic effects consist of the voltage controlled block, usually an amplifier or filter; and circuitry to derive a DC control voltage from audio, known as a level detector. While both segments affect the sound, the level detector dictates dynamic personality. Accordingly, this treatment highlights ingredients that let the builder realize several types of level detectors.

Compression

Compressors are amplifiers built to diminish dynamic range by changing gain on the fly. Compression serves to prevent clipping, to avoid saturation of recording/transmission media, or simply to achieve a

Fig. A68. Compression block diagrams.
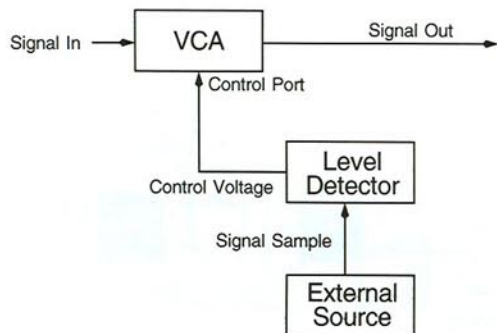
Feedback Control Loop

Parallel Control Loop
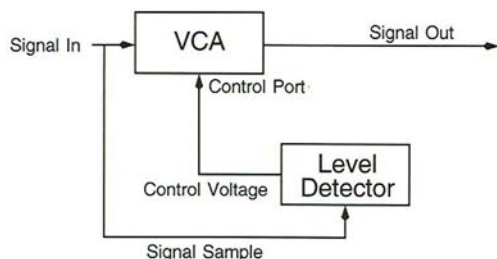
Side Chain

Dynamic Effects & Processors

Compressors
  upward
  downward
  dual-mode
Limiters
  reactive
  clipping
  power-amp
Expanders
  upward
  downward
  dual-mode
Companding Systems
  dbx (wideband)
  Dolby (mainly treble)
Level-Dependent Filters
  envelope filters (auto-wah, etc.)
  sibilance reducers
  single-ended noise reduction (e.g., DNR)
Level-Dependent Panners & Mixers
Level-Dependent Distortion Generators
Level-Dependent Tremolo Effects (rate, depth)

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desired sound. Downward compressors reduce gain, upward compressors boost gain, and dual-mode compressors do both.

Compression is realized in a variable gain block known as a voltage controlled amplifier (VCA). VCAs come in dedicated chips, or can be built from op amps, multipliers, transistors, or tubes. A VCA has an input, an output, and at least one control port, fed by the level detector. The process is readily perceived from block diagrams (Fig. A68). Signal feeds the VCA and the level detector. As signal level rises, detector output also rises, lowering gain of the VCA. When signal level falls, detector output falls and VCA gain moves back toward the resting state.

Compressor behavior depends on where the level detector samples the signal. It could sample any point, but usually takes the output of the VCA (feedback compression) or the input of the VCA (parallel compression). Feedback compression strikes the ear as tight, controlled. Parallel or "feed-forward" compression leans toward exuberance.

When the control voltage originates outside the signal path, one signal moderates the volume of another, a practice called ducking. The control path in this case is known as a side chain. Ducking happens, for example, when the announcer's voice moderates background music in a radio spot.

Skewing the signal prior to sensing by the level detector affects the compressor's response to different frequencies. For instance, treble emphasis exagger-
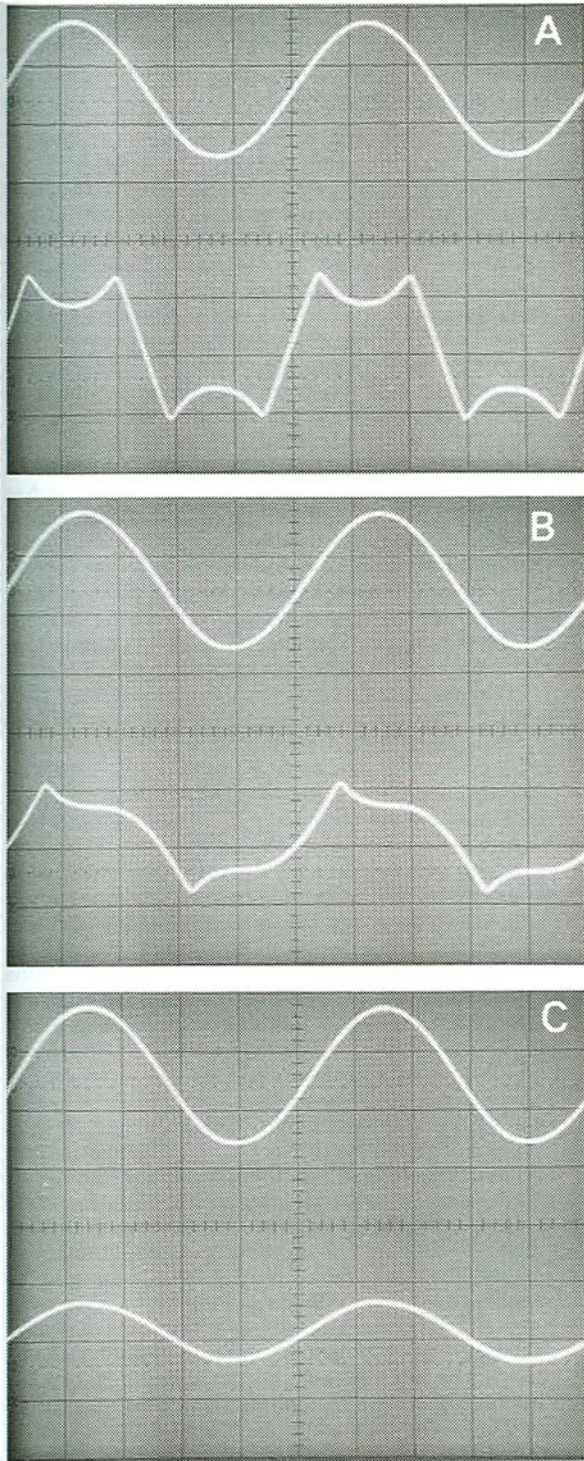
Fig. A69. Squeeze-O-Matic input/output curves illustrate effect of decay on distortion. Top trace 20-Hz input, 2V/div.; bottom trace SOM output, 1V/div.; parallel control path, ratio maximum, attack & decay minimum. A—Severe distortion results mainly from excessively fast decay. B—Decay pot is straight up; distortion decreases. C—Decay is at maximum; low distortion of 20-Hz sinewave. This demonstrates the most extreme musical example; higher frequencies do not require such lengthy decay to avoid high distortion. Attack also plays a role in distortion, but does not constitute the level detector's main integrating mechanism.
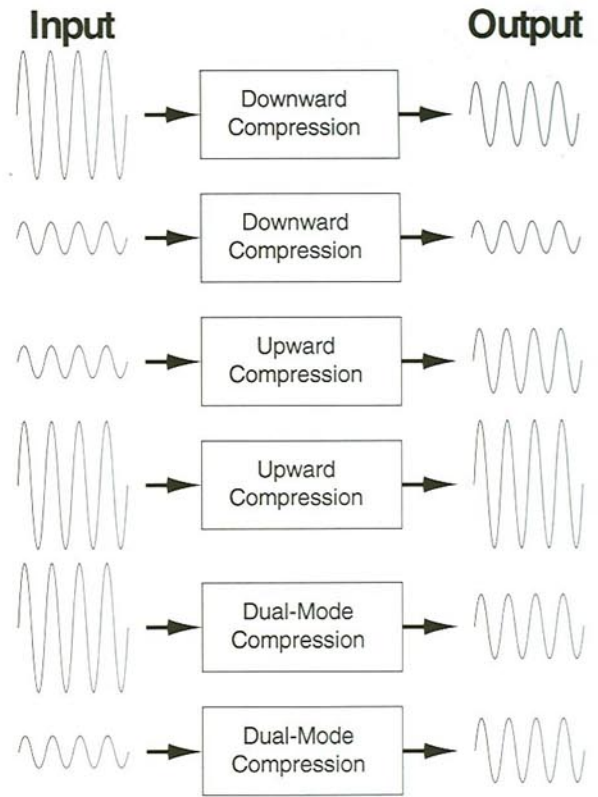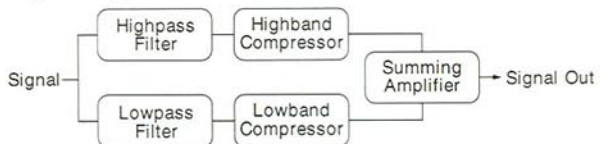
Fig. A71. In downward compression, signal above threshold is reduced, signal below threshold is unaffected. In upward compression, signal below threshold is boosted, signal above threshold is unchanged. In dual-mode compression, signal above threshold is reduced, signal below threshold is boosted.

ates compression of high frequencies, giving one means to blunt sibilance.

Compression does not happen instantly. Delay between the onset of a transient and the time compression commences is known as attack. Delay between subsidence of a transient and the time the compressor returns to resting gain is called release or decay. Attack and release affect several aspects of compression. The nature of audio dictates that attack happen quickly, usually in 20 ms or less. If attack takes too long, the waveform may clip. If attack happens too fast, the feed loses dynamic impact. Proper attack also depends on the size and shape of the transient. Explosive percussion demands attack in under a millisecond. Compression of an orchestral crescendo can kick in leisurely.

Release determines what frequencies the compressor can process without inducing undue distor-

Fig. A72. Split-band compressor block diagram.

Input   # Output

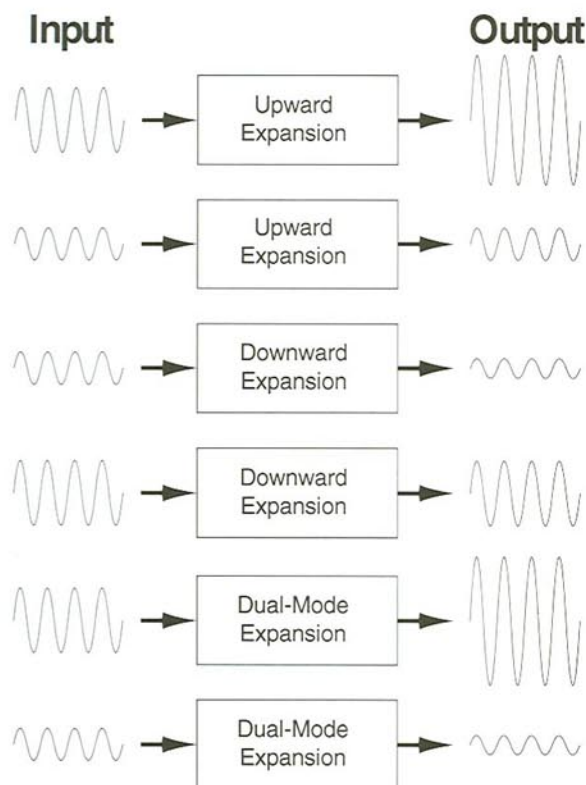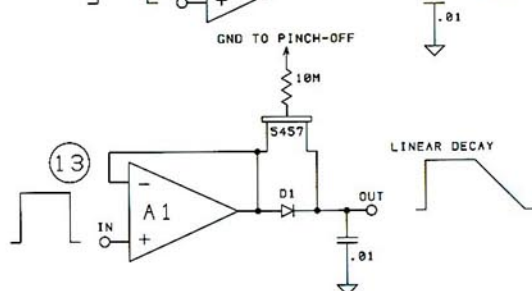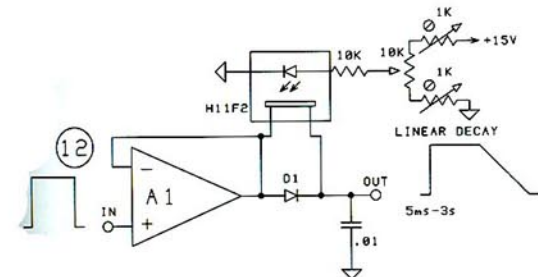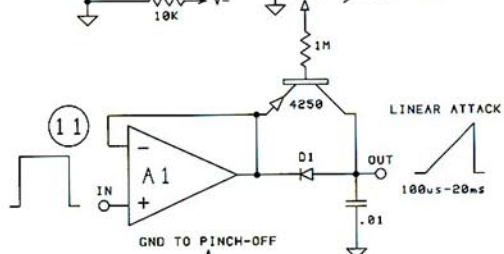|	
Upward Expansion	
Upward Expansion	
Downward Expansion	
Downward Expansion	
Dual-Mode Expansion	
Dual-Mode Expansion	

Fig. A73. In upward expansion, signal above threshold is boosted, signal below threshold is unaffected. In downward expansion, signal below threshold is reduced, signal above threshold is unaffected. In dual-mode expansion, signal above threshold is boosted, signal below threshold is reduced.
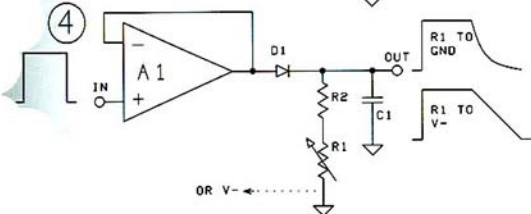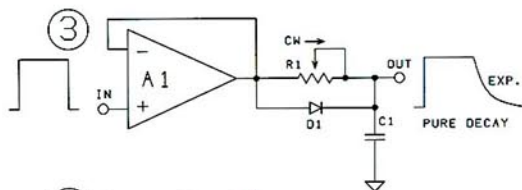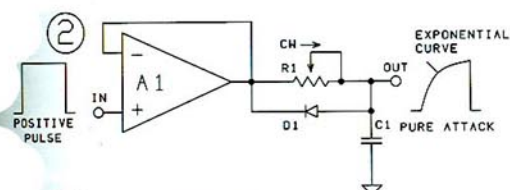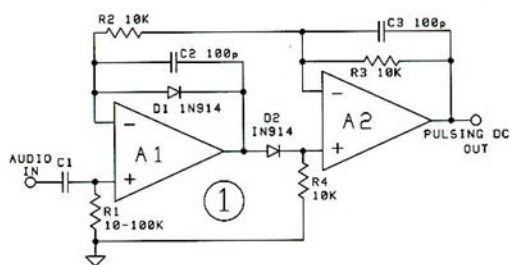
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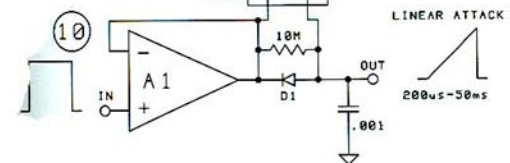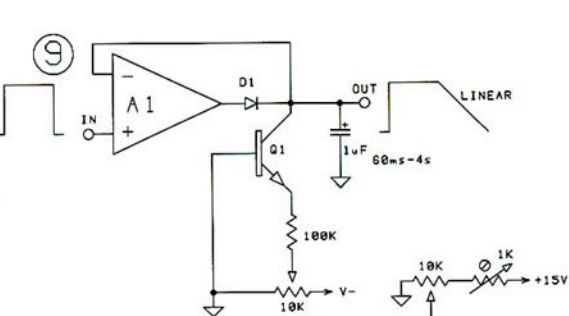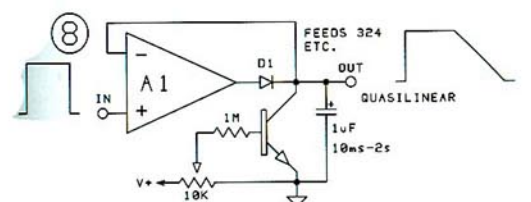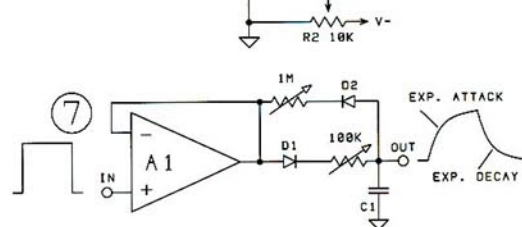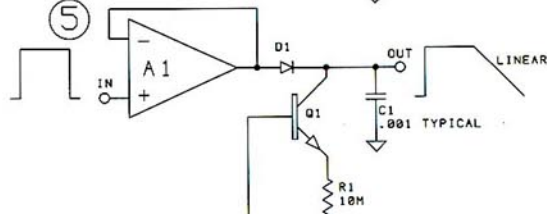
tion. If attack takes 1 ms and release takes 10 ms, the compressor should not distort signals down to ~100 Hz. Distortion increases below 100 Hz, because those frequencies' periods exceed 10 ms. The control voltage modulates during individual audio cycles, resulting in distortion. For this reason, compression acting on wideband signals should release no faster than 50 ms, the period of 20 Hz. (In practice, low distortion demands much longer decay; see Fig. A69.) If release takes too long, the compressor that has responded to a loud peak will still be running at reduced gain for soft passages that follow, which can be heard rising in volume as release occurs.

No one attack/release suits all feeds. Wideband audio demands compromise settings that abuse frequencies at the extremes. A split-band compressor solves this problem by dividing the program into two or more bands according to frequency. Each band feeds a separate compressor whose attack and release are optimized for the band. Compressed feeds recombine at the output. This technique has drawbacks, notably the phase shift due to filtering and the cost of a second compressor. Alternative solutions include program-dependent attack and decay.

The signal level at which compression kicks in is

known as threshold. In downward compression, threshold defines the point above which gain reduction occurs. In upward compression, threshold defines the point below which boost ensues. Threshold may define both points in dual-mode compression, or

Fig. A74 (facing page). 1—Precision fullwave rectifier used in level detector of most projects in this book; adapted from one shown in Ref. 14, p. 249. Positive half of AC input passes through augmented diode D2 to A2's output; negative half of AC input is blocked by D2, but conducted through augmented diode D1, thence through R2 to A2's inverting input; the now-positive output passes to A2's output. Reverse polarity of both diodes to get a negative output. C1 & R1 may be omitted if rectifier is driven off ground-biased output of op amp. Accuracy is adequate using 5% resistors; C2 & C3 are necessary for stability, and do not materially affect response in the audio band. A1 & A2 should be reasonably fast FET-input types, such as 08X/07X/06X. Figs. 2–13 show basic attack & decay networks. Variable attack networks vary the time needed to charge a capacitor; variable decay networks vary the time needed to discharge a capacitor. 2—Pure attack network. Pulse must pass through R1 to charge C1. When R1 is 0, this happens instantly, because A1's output has a low impedance; the time constant with C1 is small. Charging takes progressively longer as R1 increases. Resultant attack curve follows exponential contour of a cap charging through a resistor. When pulse subsides, C1 discharges through D1 into A1's output stage. This happens instantly because output stage has low impedance. 3—Pure decay network. Identical to #2, but D1 has been reversed. C1 charges instantly through D1 because A1's output stage has a low impedance and thus a short time constant with C1; but C1 can only discharge through R1 into A1's output stage. Decay follows exponential curve of cap discharging through a resistor. 4—Similar to #3, but now C1 discharges through R1-2 to ground; or, if the resistance ties to V–, R1-2 becomes a quasi-constant-current source that neutralizes the positive charge stored in C1. 5—Linear decay results from neutralizing the charge stored in C1 using a negative constant-current source, Q1. This circuit works fine when the output is buffered by FET-input op amps, but not when feeding certain bipolar-input types, such as LM324/358 or MC3317X. 6—Quasilinear decay circuit that can be buffered by bipolar-input amps. 7—Independently variable attack & decay in a single stage. 8—Similar to #6, but D1 is now in A1's feedback loop, eliminating the forward drop. Decay is not purely linear; can feed bipolar-input op amps. 9—Similar to #5, but diode is now inside the feedback loop. 10—Similar to #2, but R1 has been replaced by an FET in an optocoupler, providing a constant-current path, resulting in linear attack. The 10M resistance keeps the path from opening completely when the FET is pinched off. 11—Similar to #10, but FET has been replaced with a PNP BPT. 12—Similar to #3, but R1 has been replaced by an FET constant-current path for linear decay as cap discharges into A1's output stage. 13—Similar to #12, but a standard FET is used in place of optocoupler FET. Control range of this circuit is extremely narrow, such that FET is best used as a switch, fully ON or fully OFF. In most cases, output of attack/decay network feeds a noninverting op-amp to buffer the control voltage, provide gain, or impress a DC offset. Those buffers have been omitted from the illustrations to save space. All diodes are 1N914 or similar. In circuits #11 & #12, adjust trims to obtain desired control range.
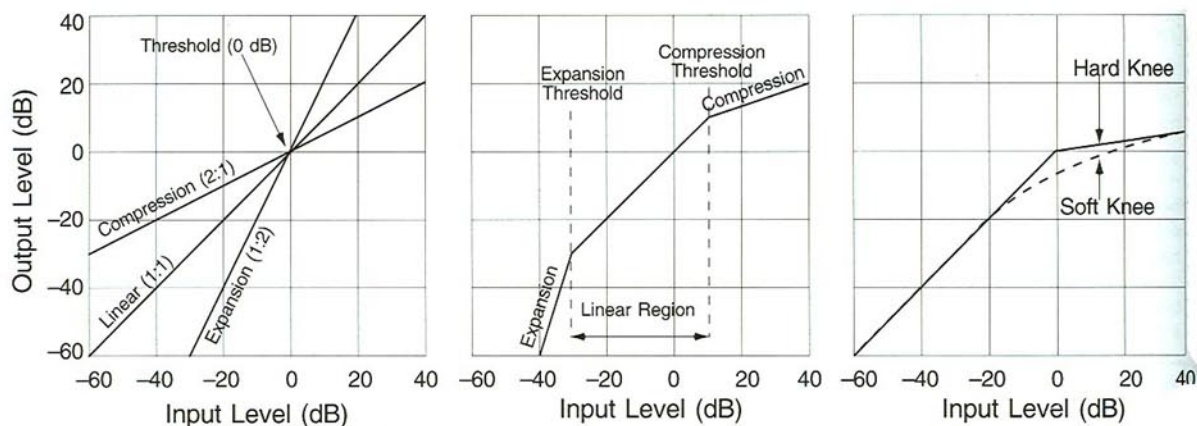
① R2 10K · C2 100p · D1 1N914 · A1 · AUDIO IN C1 · R1 10-100K · D2 1N914 · C3 100p · R3 10K · A2 · PULSING DC OUT · R4 10K

② IN · A1 · POSITIVE PULSE · CW → R1 · OUT · D1 · C1 PURE ATTACK · EXPONENTIAL CURVE

③ IN · A1 · CW → R1 · OUT · D1 · C1 · EXP. PURE DECAY

④ IN · A1 · D1 · OUT · R2 · C1 · R1 · OR V- · R1 TO GND · R1 TO V-

⑤ IN · A1 · D1 · OUT · Q1 · C1 .001 TYPICAL · LINEAR · R1 10M · R2 10K · V-

⑥ IN · A1 · FEEDS 324 ETC. · OUT · QUASILINEAR · 10M · .033 5ms-3s · V+ 10K · 470

⑦ IN · A1 · 1M · D2 · D1 · 100K · OUT · C1 · EXP. ATTACK · EXP. DECAY

⑧ IN · A1 · FEEDS 324 ETC. · D1 · OUT · QUASILINEAR · 1uF 10ms-2s · 1M · V+ 10K

⑨ IN · A1 · D1 · OUT · Q1 · LINEAR · 1uF 60ms-4s · 100K · 10K · V-

⑩ H11F2 · 10K · GND TO +15V · 10M · A1 · IN · D1 · OUT · .001 · LINEAR ATTACK 200us-50ms

⑪ 10K · 1K · +15V · 1M · 4250 · A1 · IN · D1 · OUT · .01 · LINEAR ATTACK 100us-20ms · GND TO PINCH-OFF · 10M · 5457

⑫ H11F2 · 1K · +15V · 10K · 10K · 1K · A1 · IN · D1 · OUT · .01 · LINEAR DECAY 5ms-3s

⑬ IN · A1 · D1 · OUT · .01 · LINEAR DECAY

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Fig. A75. Input vs. ouptut curves typically used to specify compressor/expander action. Left—Compression/expansion I/O curves expressed in dB. Center—Compression/expansion curves more typical of a compressor equipped with a downward expander. Right—Soft knee vs. hard knee I/O curves.

independent thresholds may exist.

The compressor follows a gain-change law known as a ratio, stated in linear terms or in dB. For instance, 4:1 downward linear compression means that the portion of a signal exceeding threshold will emerge from the compressor reduced to 25% of its original amplitude. In 2:1 dB compression, the dB output equals half the dB input. For a dual-mode compressor, −40 dB input generates −20 dB output; +10 dB input generates +5 dB output.

The manner in which the compressor switches from linear transfer to compression law is known loosely as its knee. An abrupt switch is called a hard knee. Soft-knee devices activate gradually, resulting in less audible compression. Soft knee takes place as a sliding variable ratio, or in several discrete steps.

Of the infinite possible compression settings, three broad tactics keep the signal within bounds of the medium; each confers a different sound. A low ratio/very low threshold allows practically no linear transfer. Medium ratio/medium threshold allows the feed a moderate linear transfer before compressing it. High ratio/high threshold allows the feed a linear transfer over most of its range, but subjects it to severe compression above threshold. This type of compression tends more to be audible than the first two.

Sustaining a note past its natural decay is one of compression's commonest uses. Traditional sustain squeezes the feed at low threshold and high ratio. During closely spaced notes or chords, the compressor applies heavy gain reduction. When the player lets a note hang, the compressor lessens the amount of gain reduction as the note decays, maintaining constant volume.

A more aggressive-sounding sustain uses upward compression. In this case the compressor runs at unity gain above threshold, but boosts signals that fall below threshold. The effect sounds different from downward sustain and is useful to preserve dynamic

peaks, while prolonging held notes and chords.

The two types can work together. Downward sustain is easier to implement, in part because dynamic controllers attenuate a lot more than they boost; the SSM2120 can apply up to 100 dB of attenuation but only 40 dB of boost. Attenuation is also inherently quieter than boost. Upward sustain meets a practical limit in the noise present when gain tops out.

Expansion

Expansion is the reverse of compression. Loud signals intensify, faint signals soften. Like compression, expansion comes in downward, upward, and dual modes. Threshold, ratio, attack, and release apply.

Downward expansion is a common noise-reduction technique in which a VCA reduces gain when the signal falls below a threshold. Hiss and hum fall by the amount of gain reduction. Upward expansion finds occasional use in circuits that exaggerate dynamics. Dual-mode expansion has been used in consumer audio gear to restore dynamic range to compressed feeds, but crops up more commonly as the mirror image of compression in compression-expansion ('companding') noise reduction systems, such as dbx.

Level Detector Basics

The soul of compression—of all dynamically responsive effects, in fact—lives in the level detector. Clean VCAs are a dime a dozen. Any VCA worth its salt does what the level detector tells it to do. Because the level detector forms the heart of dynamic effects, the stomp-box builder should be equipped to tune it for the needs of the effect. A full-featured circuit incorporates:

▸ an audio-to-DC converter
▸ an integrator
▸ a means to delay onset of the effect (variable attack)
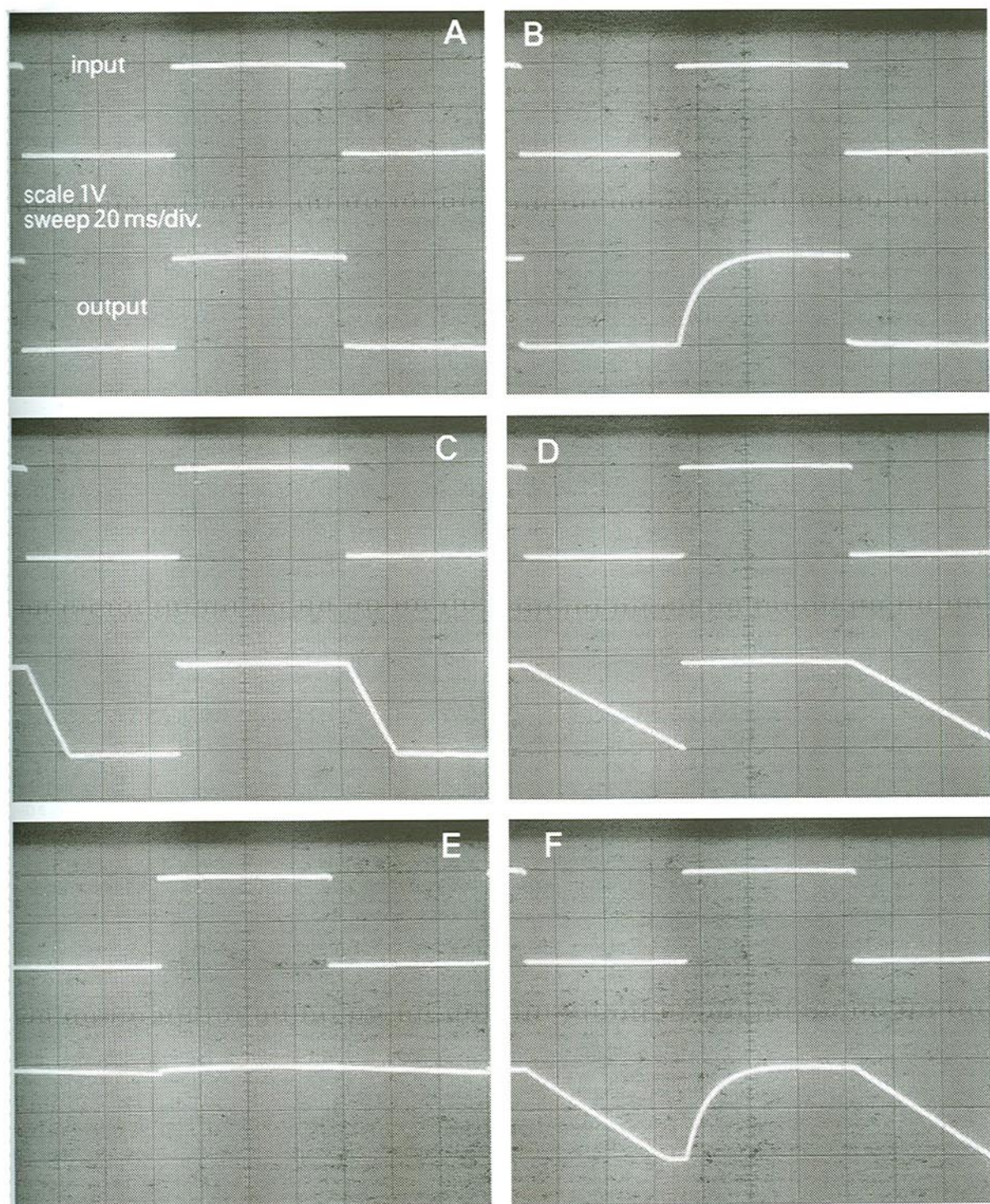▸ a means to delay subsidence of the effect

Fig. A76. A—Variable attack network (Fig. A74, schematic 2), with R1 at 0; instant attack & decay. B—R1 has been advanced to 100K. Now attack takes about 20 ms, decay is not affected. Attack curve has an exponential contour. C—(Fig. A74, schematic 5) Decay minimum; note linear decay curve. D—Decay increases. E—Decay is at maximum; voltage hardly falls between pulses, illustrating the integration function of decay networks. F—Result of cascading the two circuits (decay before attack), independently variable attack & decay.

159
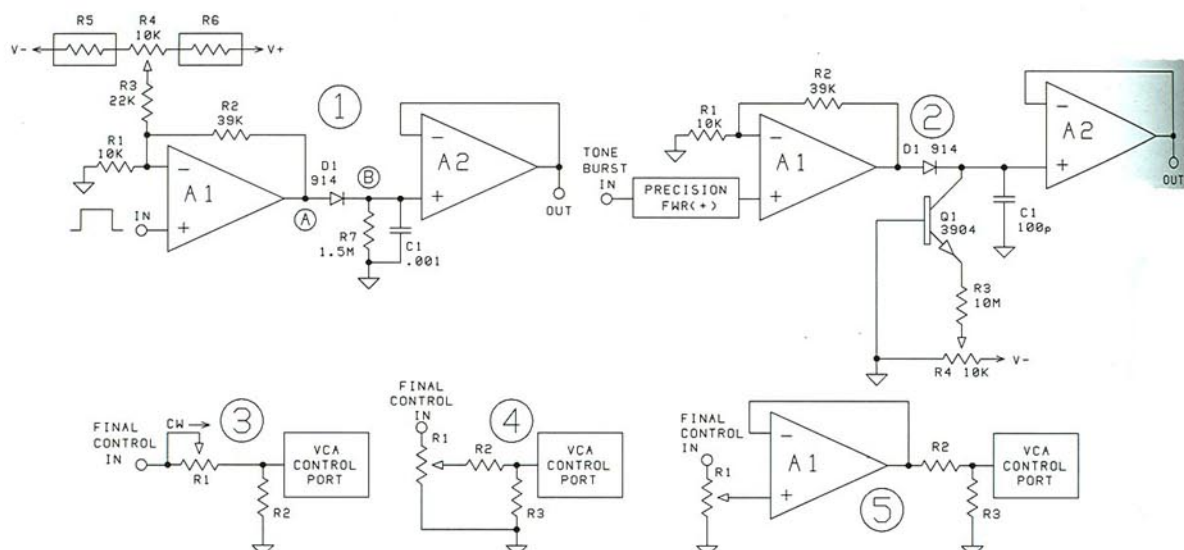
Fig. A77. 1—A common means to vary threshold. Assume that only positive voltage will be allowed to VCA. Positive voltage is boosted by A1, which also ties to network R3-R4, which applies a variable DC bias to A1's inverting input, resulting in a variable DC offset at A1's output. The potential at point B is determined by R7 and is very close to ground. For D1 to conduct, voltage seen by the anode will have to exceed 0.6V, the forward drop of a silicon diode. R4 varies the resting DC offset present at point A. If this were set at 0.59V, the signal at that point would have to have an amplitude of 10 mv to get D1 to conduct. Therefore, the threshold at point A is 10 mv, or −40 dBV. Because raw signal is boosted by a factor of 4.9 in A1, threshold at A1's input is (10 mv÷4.9) = 2 mv. If the DC bias were shifted to make the potential at point A −2V, then threshold has been raised to [(2V + 0.6V)÷4.9] = 0.531V. Choose values of R5 & R6 that limit the output offset that R4 can produce, to keep the threshold setting from altering VCA resting gain. Photos in Fig. A78 illustrate action of this network using a squarewave input. 2—Setup used for photos in Fig. A81, to illustrate the integrator action of a decay network. 3—To vary how strongly the effect responds to control—ratio, in the case of compression or expansion—simple resistive dividers can suffice. Divider action of R1 in series with R2 varies the percentage of control voltage that gets to the control port. However, a 1:1 ratio is not practical with this configuration. 4—Illustrates a better solution. 5—Interposition of a buffer may be needed, depending on drive level required by control port. Circuits 4 & 5 allow ratios down to 1:1.

(variable decay)
► a means to vary the point at which the effect kicks in (threshold)
► a means to vary how strongly the effect responds to control (ratio)
► ancillary circuits, such as feedthrough trims & I/O buffers

AC–DC Conversion

Conversion usually occurs in a precision fullwave rectifier that generates pulsing DC whose polarity the builder defines during design (see Ref. 14, pp 235–253). Derivation of a control voltage usually requires that the voltage be smooth enough not to distort the signal it controls. Smoothing happens by the process of integration, which can be considered a form of low-pass filtering. Integration occurs naturally in a variable-decay network (Figs. A80, A81).

Variable Attack

[*Note: Though attack and decay networks act upon rectified audio, it's customary to describe their effect on a squarewave, to isolate the traits under consideration. In this context,* instantaneous *means less than one millisecond.*]

A pure attack network delays the leading edge of a pulse without affecting fall. This is achieved using a diode, a cap, and a potentiometer, fed by an op amp (Fig. A74–2). C1 charges through A1's output stage in series with R1. Because the output stage has a very low impedance, charging happens instantly when R1 is zero. Charging takes progressively longer as R1 increases. Duration depends on the *time constant* of the resistor and the capacitor. Time constant in seconds is given by the product of resistance in ohms times capacitance in farads. A 10K resistor combined with a 10μF cap has a time constant of [10K × (10μF ÷ 1,000,000)] = 0.1 seconds. Roughly five time constants are required for the cap to charge fully. Once the pulse subsides, C1 discharges through D1 into A1's output stage. The low impedance of this path makes decay instantaneous so long as C1 has a relatively low value. Suitable attack times for compression range 0–20 ms.

Variable Decay

Intuitively, we see that reversing the diode of the variable-attack circuit results in variable decay (Fig. A74–3). C1 charges instantly through D1, but C1 can only discharge through R1 into A1's output stage. An equally practical approach discharges C1 to ground through R1, kept from shorting the signal by R2, whose value sets minimum decay (Fig. A74–4). In
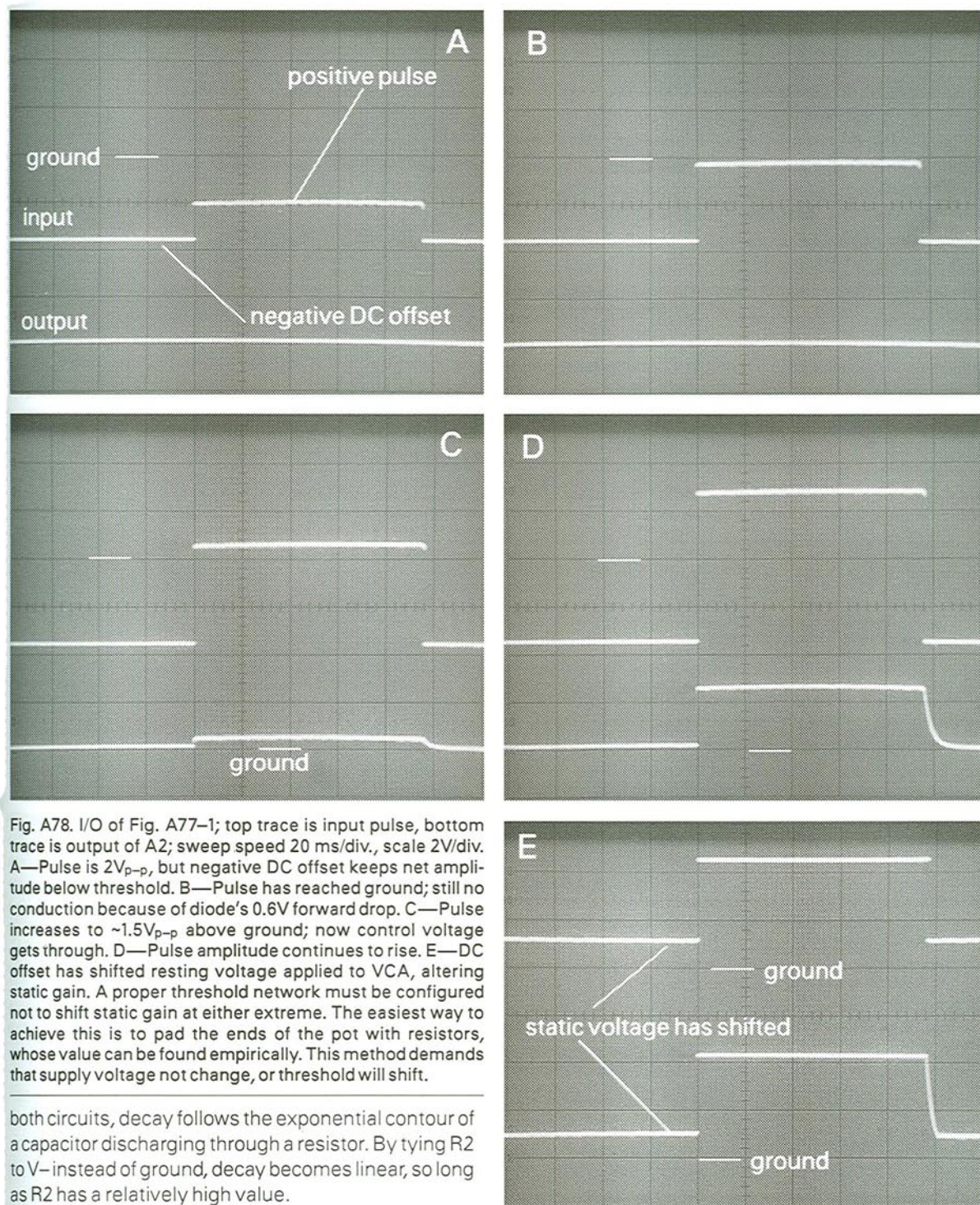
Fig. A78. I/O of Fig. A77–1; top trace is input pulse, bottom trace is output of A2; sweep speed 20 ms/div., scale 2V/div. A—Pulse is 2V_{p-p}, but negative DC offset keeps net amplitude below threshold. B—Pulse has reached ground; still no conduction because of diode's 0.6V forward drop. C—Pulse increases to ~1.5V_{p-p} above ground; now control voltage gets through. D—Pulse amplitude continues to rise. E—DC offset has shifted resting voltage applied to VCA, altering static gain. A proper threshold network must be configured not to shift static gain at either extreme. The easiest way to achieve this is to pad the ends of the pot with resistors, whose value can be found empirically. This method demands that supply voltage not change, or threshold will shift.

both circuits, decay follows the exponential contour of a capacitor discharging through a resistor. By tying R2 to V– instead of ground, decay becomes linear, so long as R2 has a relatively high value.

Another linear solution uses a transistor as a variable constant-current source (Fig. A74–5). This circuit varies decay from ~5 ms to several seconds. Capacitor values on the order of 0.001 μF preserve instantaneous attack. The circuit works great buffered by FET-input op amps (TL06X, 07X, 08X, LF44X, etc.), but not when buffered by certain bipolar-input types, including 324 and 3317X.

Variable attack combines with variable decay in single stages like Fig. A74–7, or in cascaded networks used in several projects in this book. Placing the decay network ahead of the attack network has given the author best results.

The circuits under consideration work with positive inputs. They adapt to negative pulses by reversing diode polarity and capacitor polarity, if applicable. The transistor-decay networks switch to a PNP transistor
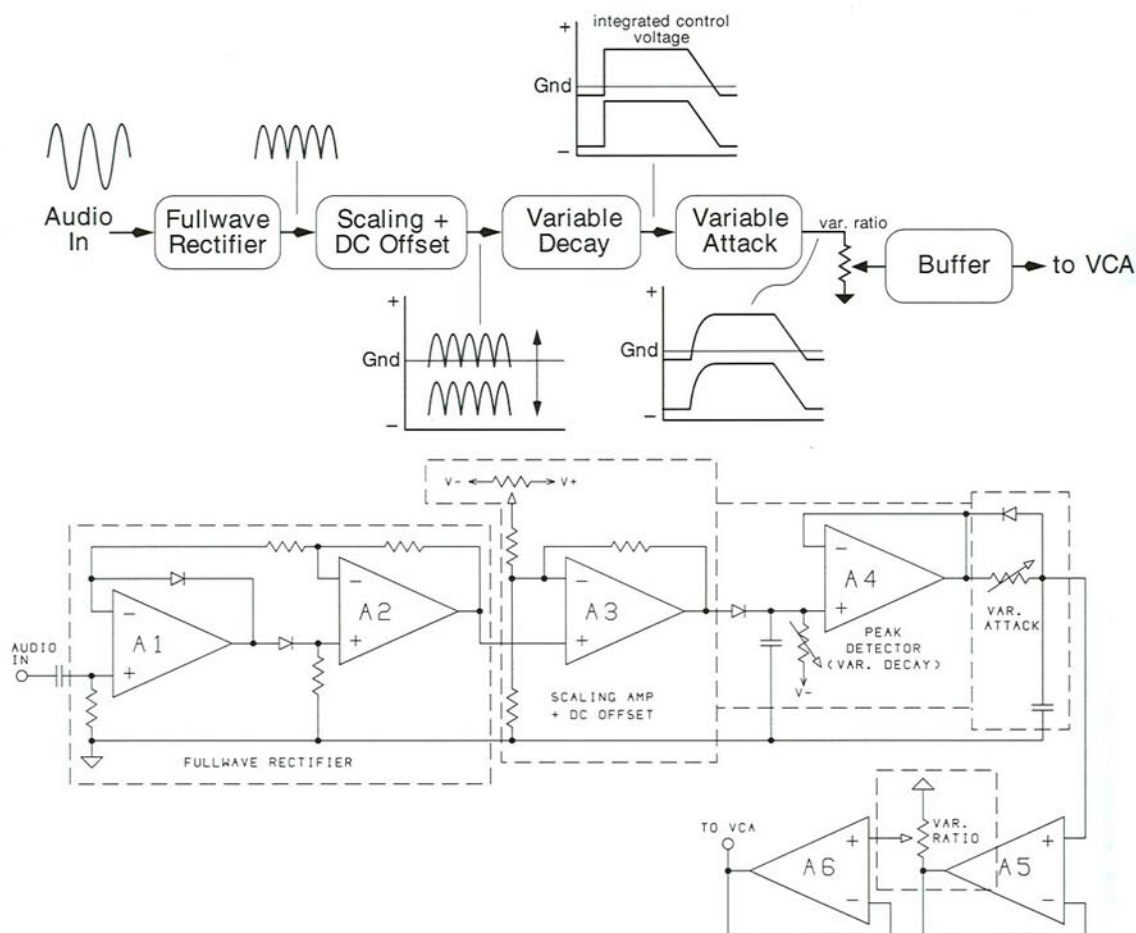
Fig. A79. Top—Block diagram of what happens in a typical level detector. Audio is converted to pulsing DC in fullwave rectifier. Rectified audio feeds a scaling amp that amplifies signal if needed, and applies a variable DC offset that determines the threshold. In this case, the diode coupling to the variable-decay network creates an internal threshold of ~0.6V. Integration occurs as an implicit function of variable decay; the control voltage emerging from this stage has been smoothed to DC. The decay network has no significant effect on attack, so the control voltage feeds a variable-attack network, then couples to a pot which controls ratio. Bottom—Specific circuits used to realize the level detector functions. A1 & A2 form a precision positive fullwave rectifier whose pulsing DC output feeds scaling amp A3, which boosts signal if needed, and applies a DC offset. Because the next stage couples through a diode, it is apparent that the voltage must exceed ground plus one diode drop to get through. This defines the internal threshold. Effective threshold at the device input is determined by gain of A3 and the DC offset at A3's output. Scaling amp feeds a variable decay network buffered by A4. The now-smooth control voltage feeds a variable-attack network buffered by A5, whose output feeds a pot that varies the percentage of control voltage allowed to the VCA.

and tie the ungrounded end of the control pot to V+.

Variable Threshold & Ratio

The ideal compressor acts as a unity-gain buffer below threshold. Achieving this requires a means to prevent the control voltage from reaching the VCA until that voltage passes some set level, the threshold. One method imparts a variable DC offset to the path traversed by the control voltage. A diode placed between the threshold circuit and the VCA control port will not conduct until the voltage differential exceeds the diode's forward drop. Controlling the conduction point means controlling the voltage on both sides of the diode. Because the potential at the VCA control

port is usually fixed, setting a threshold involves controlling the DC offset on the level-detector side of the diode (Figs. A77, A78).

In the case of a VCA that uses a positive control voltage, the diode is oriented to pass positive voltage. Therefore, the DC offset applied to raise threshold must be negative, to force the input signal to generate a greater positive voltage to reach the diode's conduction point. Making the DC offset less negative lowers the threshold by moving the voltage closer to the diode's conduction point. The threshold circuit must not apply a voltage that causes the diode to conduct in the absence of an input signal, or the VCA's static gain will shift.
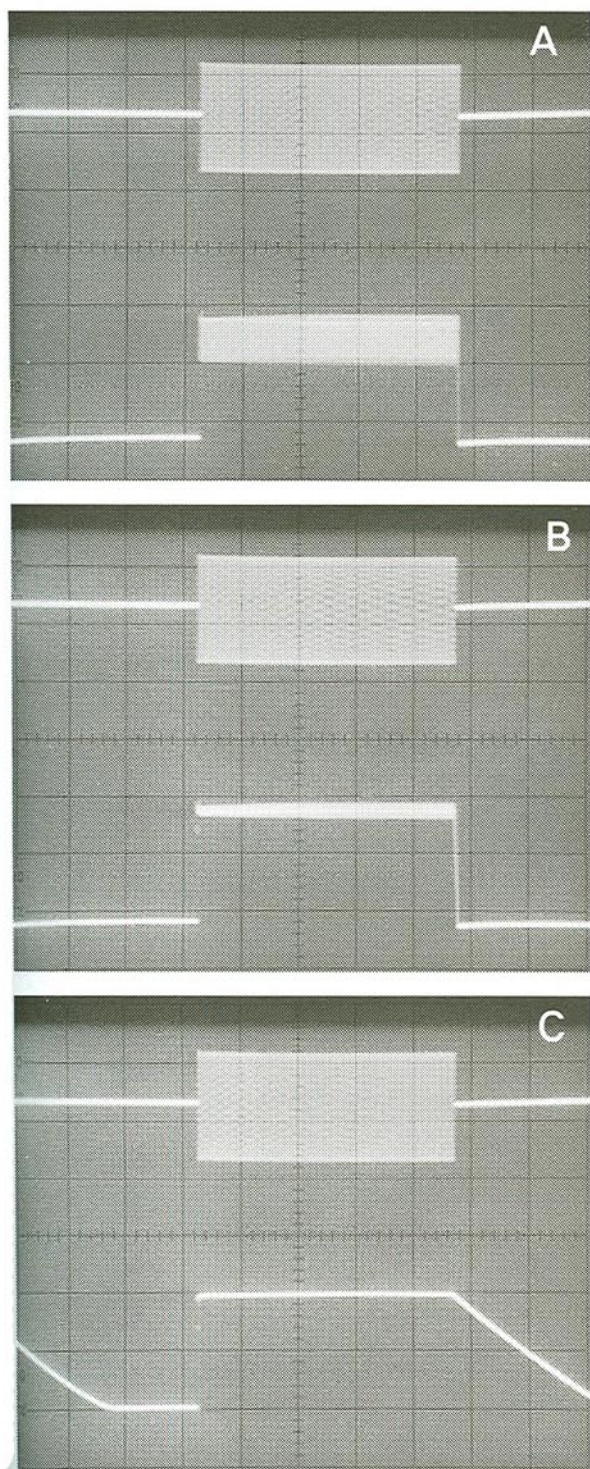
Fig. A80. I/O photos from Fig. A77, schematic 2. Tone burst input (top trace), output of A2 (bottom trace). A—Decay is at minimum. Remnants of the 5-KHz sinewave are still prominent in the level detector output pulse, and would distort the signal in a compressor. B—Decay has been increased; 5-KHz remnants start to smooth out by the process of integration. C—Decay at ~50 ms; AC in the input has been completely converted to DC in the output. This control voltage will not significantly distort the signal.

This scheme can enforce any threshold, from zero to infinite (effect off). Ten millivolts (−40 dBV) makes a useful lower limit for many effects.

Ratio is varied by altering the fraction of the control voltage allowed into the VCA control port. This circuit can be as simple as a resistive divider tied directly to the port, but use of a buffer reduces loading effects and gives smoother control over a wider range of ratios (Fig. A77).

Fig. A79 illustrates how these elements combine to form a basic level detector. Figs. A82 & A83 provide more detailed insights into level detectors for specific effects.

Characterizing Dynamic Controllers

While an experienced designer can predict much about compressor/expander behavior from the level detector's pulse response, assessment of musical behavior demands a tone burst fed through the full device. This reveals the effect on the music envelope. A tone burst generator makes an invaluable aid to designing and debugging dynamic controllers (see Project No. 33).

Advanced Techniques

Soft-Knee Compression

The manner in which a compressor shifts from linear transfer to compression-law transfer is known as its knee. Hard-knee compressors apply linear transfer below threshold, compression law above threshold. The higher the ratio, the more audibly sound hits the wall. Soft-knee compressors use a stepwise or sliding variable ratio (SVR). Here the linear ratio might measure 1:1 with a signal level of $0.5V_{p-p}$; 2:1 at 2V; 3:1 at 4V; and so on. Gradual transition masks the process, especially when the maximum ratio is high. What's more, SVR can be configured to raise the ratio to infinity should signals exceed the expected range, automatically limiting the level.

The builder looking to add SVR finds a glut of options. The simplest approach subjects the output of a conventional level detector to a voltage-dependent transfer function. As level detector output rises, a greater fraction of voltage arrives at the VCA control port. A smoothly sliding ratio can be realized by placing a VCA, multiplier, or voltage-controlled resistive divider between the level detector output and the VCA (Fig. A90). These networks feed off separate level detectors, or off the primary level detector, though the latter approach may demand steps to prevent ripple in the control voltage from compounding itself in the transfer block.

A diode string automatically alters transfer in discrete steps (Fig. A86). As each series diode conducts, the voltage passed takes an alternative path to a sum-
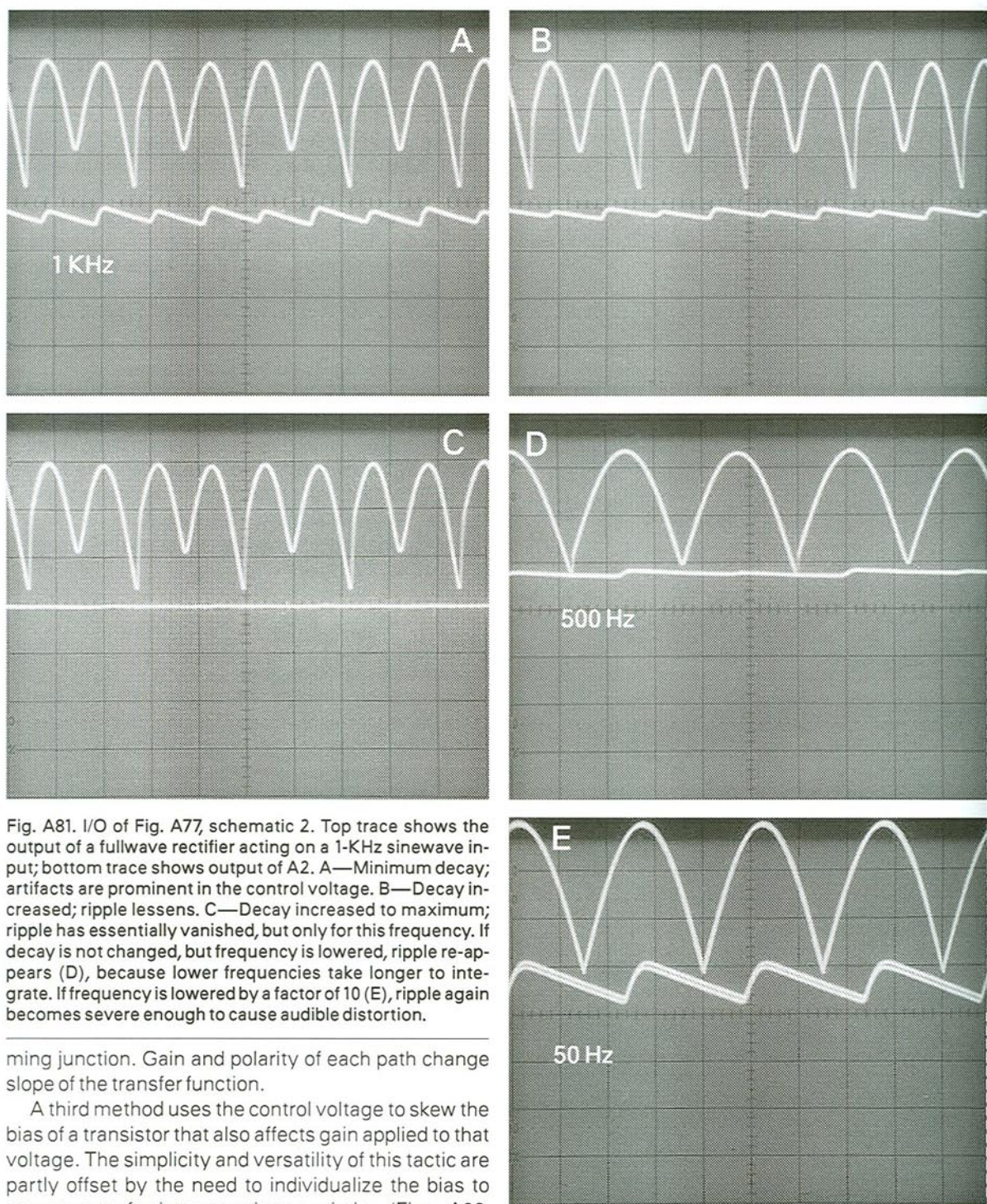
Fig. A81. I/O of Fig. A77, schematic 2. Top trace shows the output of a fullwave rectifier acting on a 1-KHz sinewave input; bottom trace shows output of A2. A—Minimum decay; artifacts are prominent in the control voltage. B—Decay increased; ripple lessens. C—Decay increased to maximum; ripple has essentially vanished, but only for this frequency. If decay is not changed, but frequency is lowered, ripple re-appears (D), because lower frequencies take longer to integrate. If frequency is lowered by a factor of 10 (E), ripple again becomes severe enough to cause audible distortion.

ming junction. Gain and polarity of each path change slope of the transfer function.

A third method uses the control voltage to skew the bias of a transistor that also affects gain applied to that voltage. The simplicity and versatility of this tactic are partly offset by the need to individualize the bias to compensate for inter-transistor variation (Figs. A88, A89).

Singly or combined, these methods achieve just about any transfer function desired, including curves that reverse polarity of slope.

Nonlinear transfer blocks find use in many guitar effects. They deform tremolo control feeds, generate distortion, and soften the knee in dynamic effects other than compression. Working up a transfer block by measuring and plotting voltages is tedious and time consuming. A basic ramp generator speeds the exercise (see Project No. 34).

The nature of soft knee prompts the player to rethink compression control settings. Hard-knee threshold defines the level at which the control voltage first alters gain of the VCA. In a soft-knee device this point is impossible to hear. Ratio varies with signal level. Soft-
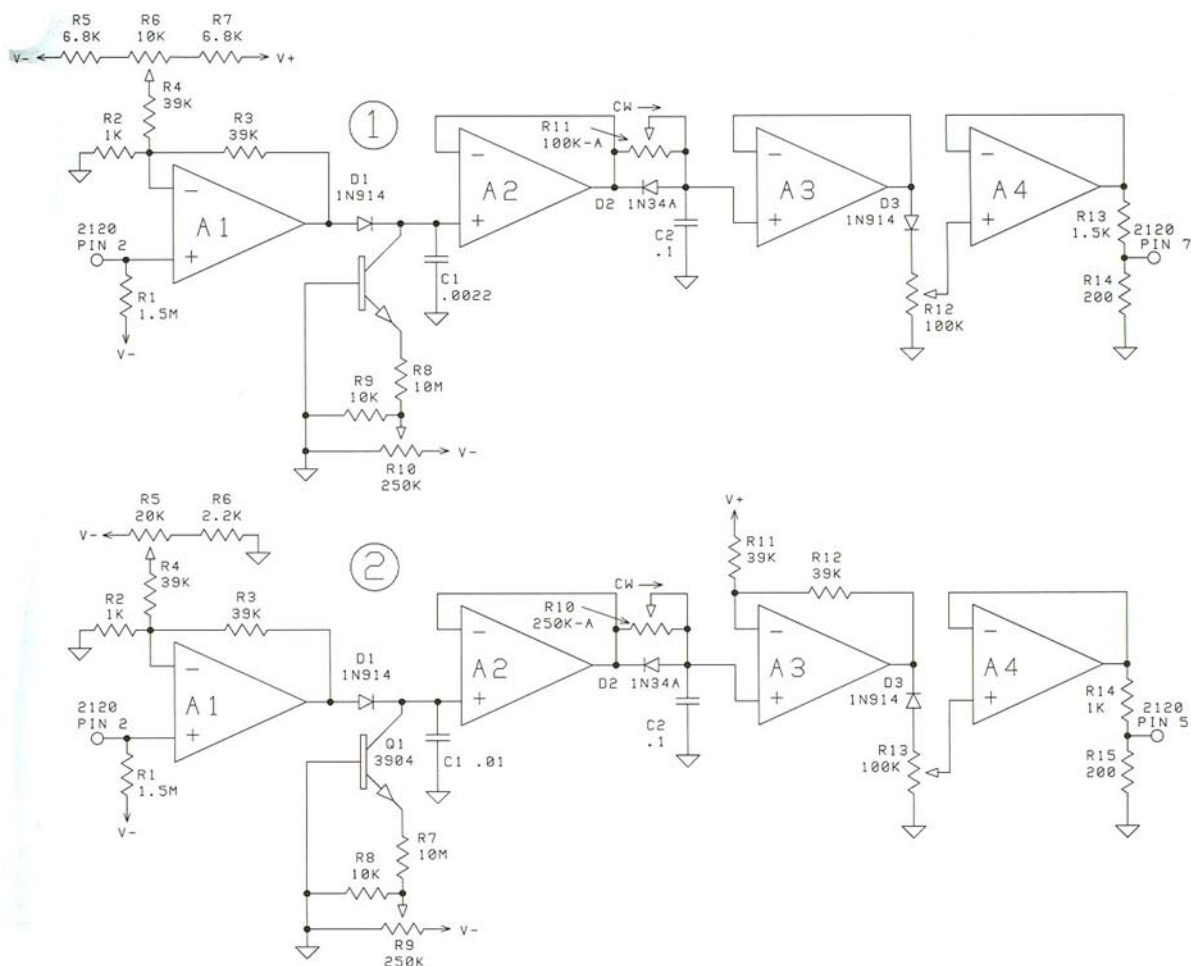
Fig. A82. 1—The level detector of Squeeze-O-Matic (Project No. 7). Compression demands a voltage whose magnitude rises with the input level. The level detector built into the 2120 provides a rectified log output at pin 2. A1 amplifies this voltage by a factor of 40; R6 applies a variable DC offset that determines threshold. Because the voltage generated is positive, compression requires either inverting the voltage, or using the 2120's '−' control port, pin 7, which reduces gain in response to a positive voltage. The latter method was chosen to save an inversion stage. Standard attack & decay networks follow A1. The output of A3 is a control voltage that swings positive in the presence of a signal, negative in the absence of signal. A downward compressor cannot allow negative voltage to reach pin 7, since this would boost gain, giving dual-mode compression. A diode at the output of A3 blocks conduction of negative voltage. Pot R12 controls the percentage of control voltage allowed through, buffered by A4. This scheme varies the linear ratio from 1:1 to about 25:1 (feedback control path); maximum ratio is determined by divider action of R13 & R14.

2—The level detector of Gate-O-Matic (Project No. 6). Downward expanders need a built-in tendency for the control voltage to swing to one extreme unless kept from doing so by a voltage generated by the input signal; the device activates in the absence of an input signal. The level detector's job is to generate a control voltage that offsets the natural tendency of the device to mute the signal. Since the 2120's built-in log amp normally generates a positive voltage, we choose negative voltage as the natural tendency. These polarity choices dictate use of pin 5, the VCA's '+' port, in which positive voltage boosts gain, negative voltage reduces gain. We create the negative tendency by tying R11 to V+, which causes the output of A3 to swing to its negative limit unless opposed by a positive voltage feeding its noninverting input. Ratio control is the same as that used for SOM. Minor changes have been made in attack and decay to suit the needs of gating, and to the DC offset network feeding A1, to alter the threshold range.

knee compression sounds subtle enough that the player cranks the ratio so high as to defeat the point. Players seeking audible compression should stick to hard-knee transfer. Soft-knee settings are best fixed by meter, rather than boosting the ratio to the point of blatant compression.

Limiting

As a constant process, compression modifies the overall sound of a track. Compression may fail to tame a peak that happens once in a session, making a need for a back-up mechanism known as a limiter. Compression supplemented by limiting allows lighter com-
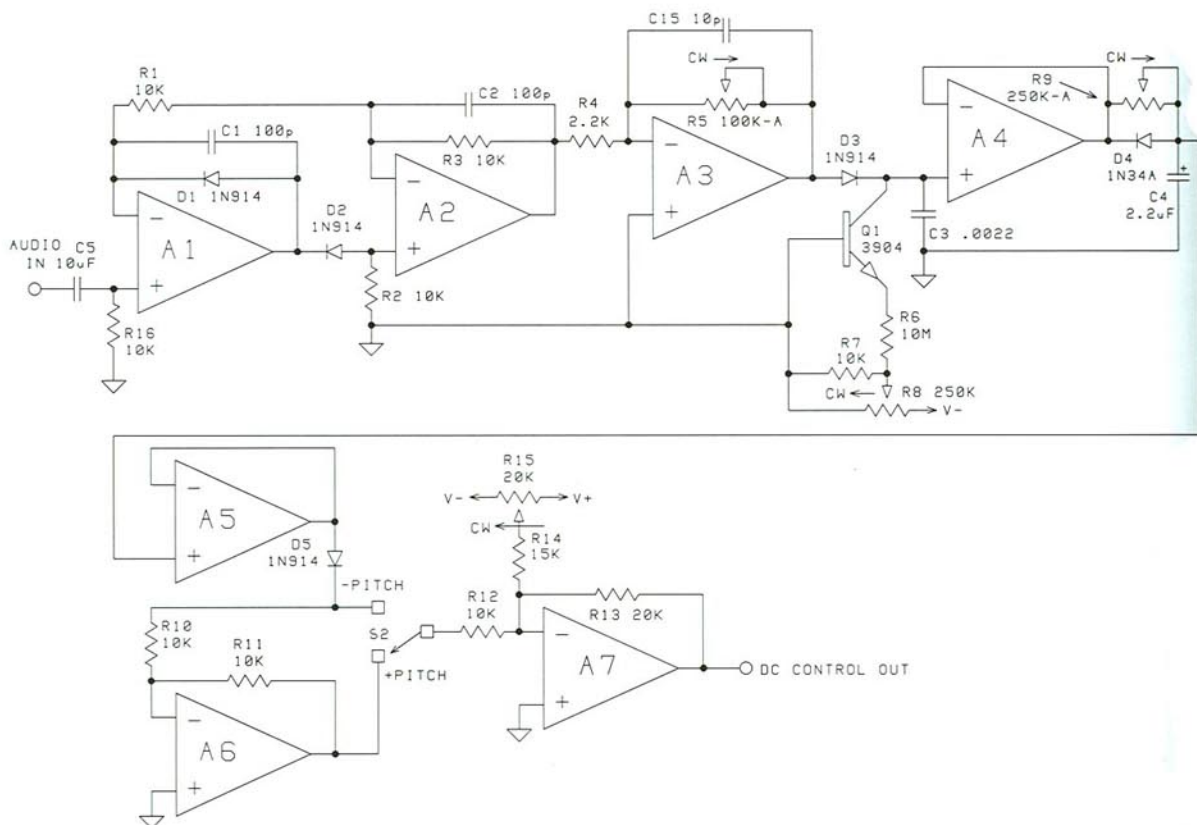
Fig. A83. The level detector of Envelo-Matic (Project No. 18). Because the VCF driven by this level detector responds over the approximate range −13.5V to +13.5V, the level detector's output should embrace this range as well. A1, A2, & associated components form a unity-gain precision fullwave rectifier whose negative output couples through R4 to inverting variable-gain stage A3. R5 varies gain 0–45. The now-positive output of A3 feeds a variable-decay network made up of Q1, C3, D3, & R6-8. R7 makes R8 a super-antilog pot that spreads the wide decay time evenly over the rotation of the pot. Decay contour is linear, to minimize control voltage ripple for a given total decay time. A4 buffers this stage, which feeds a variable-attack network made up of D4, C4, & R9. At this point we have a positive control voltage whose amplitude varies linearly with the input amplitude, and to which variable attack and decay have been attached.

A5 buffers the attack network output. D5 blocks conveyance of a small DC offset that crept in as part of the decay network. D5 also feeds unity-gain inverting buffer A6. One of S2's throws ties to the end of D5, the other to the output of A6, allowing choice of control voltages having identical magnitude but opposite polarity. These feed A7 through R12. The maximum voltage swing up to that point covers GND to op amp's positive limit, or GND to its negative limit. A7 has gain of 2, to enable this input voltage to cause A7's output to swing from its positive to negative limit.

R15 supplies a variable DC offset through R14 to set the resting DC at A7's output. Once set, that point shifts only in response to voltage generated by the audio input. S2 determines whether the shift moves the voltage higher or lower. R5 controls how far it moves by varying gain applied to the rectifier output.

pression without the risk of overload.

Limiting has vague and strict definitions. The vague sense implies compression at a linear ratio greater than 10:1, achieved by changing the compressor settings. Strict usage connotes a hard limit that the signal cannot pierce, coupled with very fast attack and decay.

Limiting sets new priorities for the level detector. Speed is king, the speed to catch the session's craziest rimshot; the speed to get out of the way in time not to bump the backslap echo. Compression set for 5 ms attack and 500 ms decay might as well be asleep at the wheel.

Proper usage demands signal management prior to the limiter. Only the rare, brief transient should trigger limiting. If a trial run shows this not to be the case, the compression settings probably need revision, or some other measure needs to be employed, such as auto-variable attack.

Limiters come in several varieties. Reactive limiters respond just as compressors do, but attack in microseconds, release in a few milliseconds. This approach can use a feedback or parallel control path. Cycling is inherent in the feedback control path, but inaudible on short transients (Fig. A91).

Staircase limiters use a multi-stage comparator as the level detector. Each comparator electronically switches gain in discrete steps. Switching takes a few microseconds. Distortion due to the discontinuous transfer function is ultrasonic and thus inaudible, but

may demand removal to avoid aliasing an analog-to-digital converter (Fig. A95).

Clipping limiters deprive the signal of headroom at a convenient point in the signal path. Op-amp diode clippers are preferred for the soft edges of the clipped waveform, resulting from AC logarithmic amplification (typical waveforms appear in the Appendix on Distortion). Another method limits the supply voltage available to a unity-gain buffer. Despite their crass approach, clippers function as transparently as reactive limiters if they act only on transients that pass too fast for the ear to register distortion.

Power-amp limiters limit in name only. They're simple, downward compressors defined by their control path. The level detector samples the power-amp output, cuts the signal to the power-stage input. The mechanism prevents power-stage clipping but not preamp overdrive, which many players use to achieve their sound. Power-amp limiters favor bass amps because bass players rarely want the power stage to clip. Photoresistors' native dynamic profile suits the job, making a power-amp limiter easy to build and to incorporate as a reversible mod.

Sustain

As a compression derivative, sustain comes in downward, upward, and dual modes. Downward sustain compresses the signal at high ratio and relatively low threshold. The amount of gain reduction falls as the note decays, sustaining the volume. Boosting the signal as much as possible before it reaches the compressor maximizes duration. Downward sustain uses

Fig. A84. Compression dB I/O curves. Hard knee is completely 1:1 below threshold, 4:1 above threshold. Soft knee shows indistinct threshold as a natural trait; compression begins well below nominative 0 dB threshold, but acts so subtly as to be inaudible. Soft knee curve #2 illustrates that function need not be smooth, and that ratio can continue to change with signal level, in this case reaching infinity.
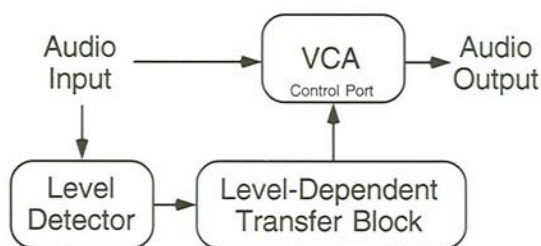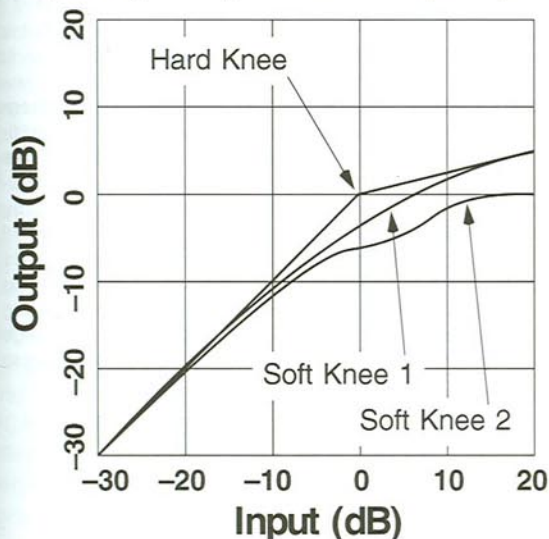
Fig. A85. Block diagram of a common means to implement soft-knee compression. The only difference between this and standard parallel-path compressor is the interposition of a level-dependent transfer network between level detector and VCA control port.

feedback or parallel control paths. Feedback control leans to subtlety, parallel control to a more forward sound, even to the point of sag.

Upward sustain boosts gain once the note decays below threshold. The result sounds more assertive than downward sustain but poses greater obstacles for the builder. Where downward sustain is naturally quiet, upward sustain magnifies system noise. Upward sustainers supply any gain desired, but lose their natural sense past ~20 dB. Higher gains are useful for 'rad' sustain, or to establish feedback, at the cost of conspicuous operation and overt noise.

In an upward sustainer, attack determines how fast the VCA reduces gain on receipt of a signal above threshold; decay determines how fast gain rises as the signal falls below threshold. Upward sustain needs fast attack to prevent clipping that otherwise happens when a large signal meets a VCA applying high gain.

A feedback control path yields upward sustain that sounds so natural that only A/B comparison betrays it, but this approach is prone to cycling. Gain rises to the point that system noise or some minor transient flips the VCA into unity-gain mode, often accompanied by an audible pop. An upward sustainer minimizes cycling by:

▶ isolating upward sustain from downward sustain in a dual-mode device
▶ limiting boost to 20 dB
▶ inserting a slight attack delay
▶ nulling the DC offset that some VCAs develop when applying boost
▶ using a low-gain level detector
▶ using a variable-ratio network similar to that which enables soft-knee compression; in this case, SVR reduces the boost voltage as the signal decays past a certain point, avoiding maximum gain at minimum input levels
▶ having the VCA default to unity gain when the input falls below a set level; this mimics the prior approach through a separate control path
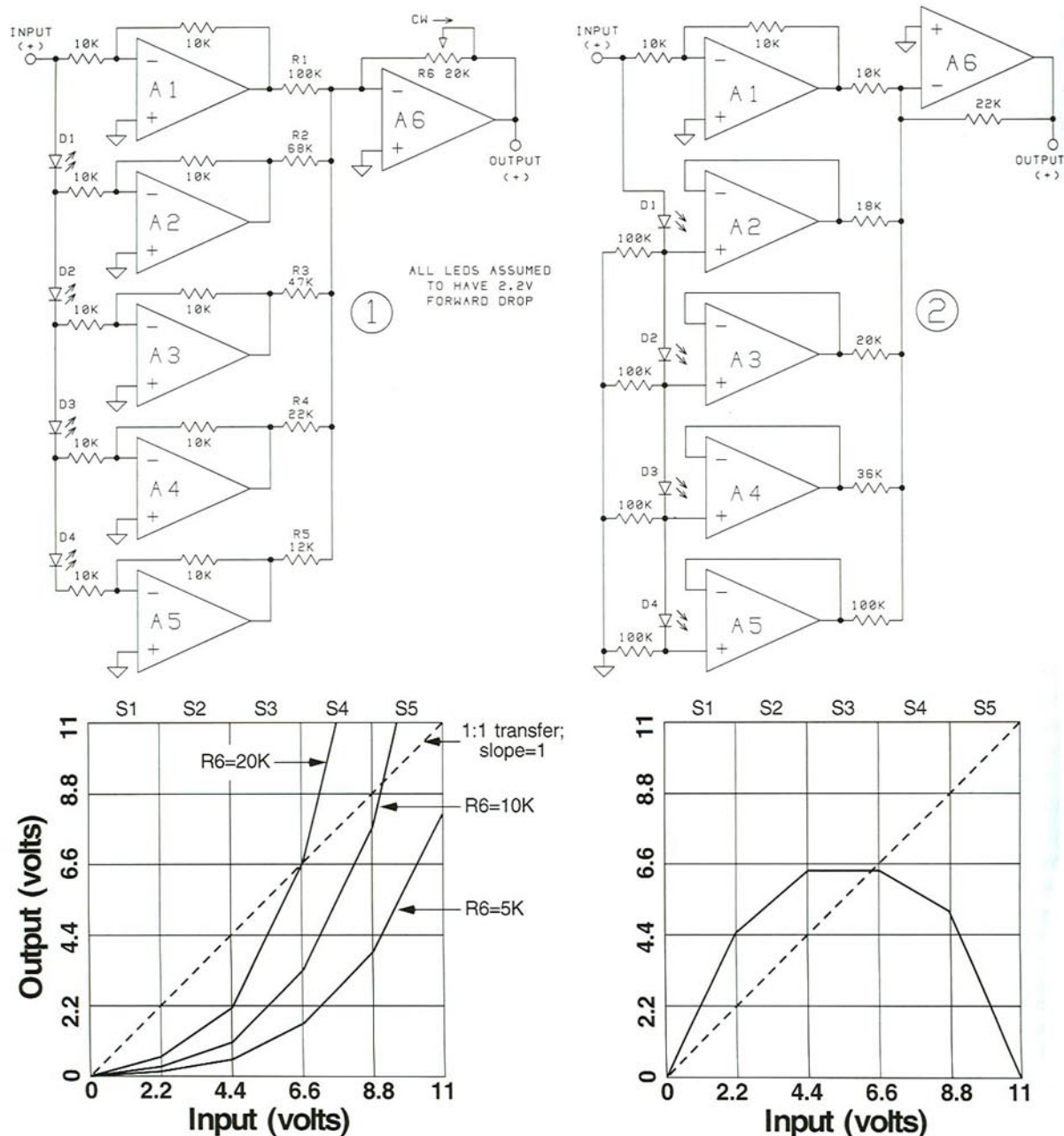
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ALL LEDS ASSUMED
TO HAVE 2.2V
FORWARD DROP

Fig. A86. 1—Diode-ladder transfer block; placed between raw level-detector output and VCA control port. Slope of first segments S1 is determined by ratio of R6 to R1. When control voltage input reaches D1's forward conduction point (~2.2V), additional voltage through that path sharpens the slope of segment S2. Succeeding diodes in the ladder conduct as input voltage continues to rise, conveying progressively greater voltage until, at about 9V, input = output. Above that point, output exceeds input. Transition point of each segment is determined by diode's forward drop; slope of each segment is determined by the ratio of coupling resistor to R6, and the running sum up to that point. Breakpoints of individual segments can be changed by using diodes having different forward drops; slope of segments can be changed by changing values of coupling resistors R1-R5. Overall amplitude of transfer function changes with setting of R6. 2—A variation on the theme, using noninverting segments to alter slope in the negative direction. This sort of curve can soften the action of an upward expander by lowering the expansion ratio as signal level rises. It is apparent that mixed inverting/noninverting blocks can cause slope to change direction, several times if needed. Circuits requiring more than four diodes can use LED bargraph modules or diode arrays to save board space.
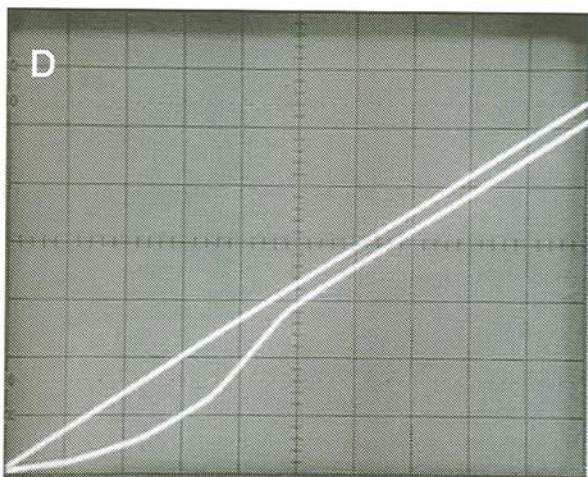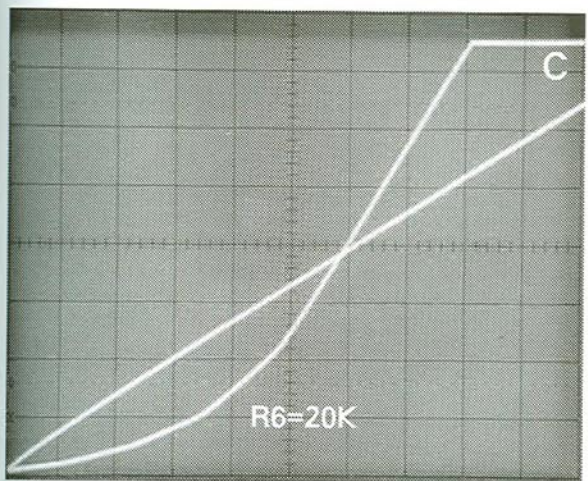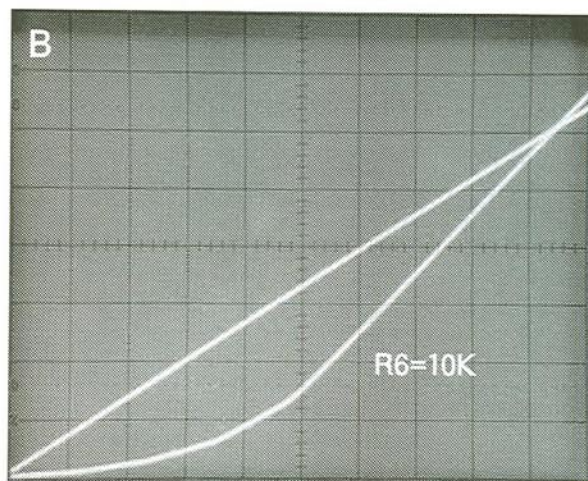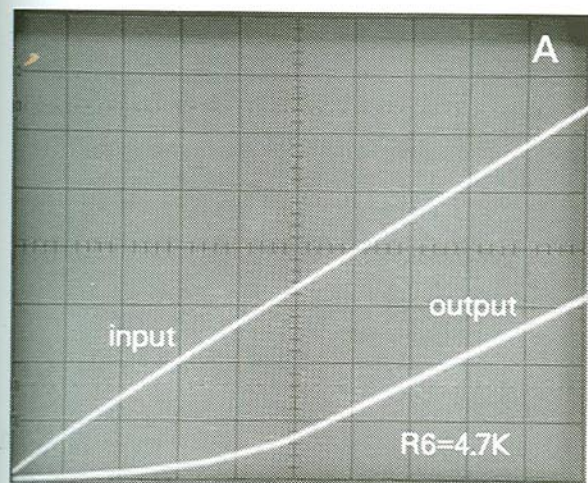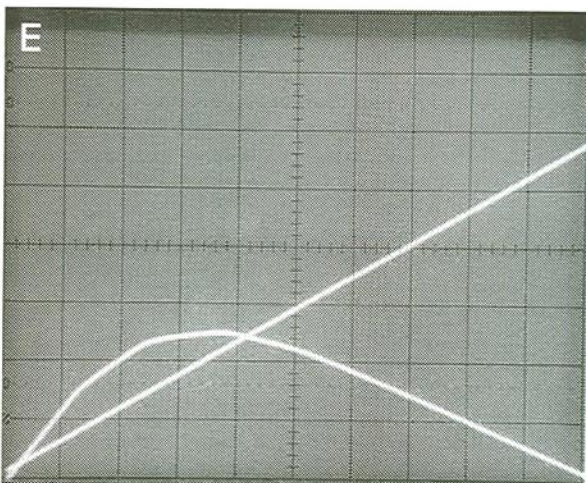
Fig. A87. A–C depict I/O of Fig. A86, schematic 1, using red LEDs having an average forward drop of 1.82V. A—R6=4.7K. B—R6=10K; this is a useful soft-knee compression curve. C—R6=20K; also useful, if effect benefits from having control voltage output exceed input at high levels. D—Shows the effect of changing final op amp A5 from an inverter to a follower; instead of continuing to increase in a positive direction, slope decreases. E—Depicts I/O of Fig. A86, schematic 2. In this case, initial positive slope grows less positive as each diode conducts, finally becoming negative. This type of curve is useful in, say, an upward sustainer, to prevent runaway gain at minimum input level; or an upward expander, to decrease the expansion ratio as input level rises. All photos scale 2V, sweep 500 μs.

A parallel sustain control path resists cycling but sounds less natural than feedback. Volume may actually rise as the note decays. Parallel control works well to initiate feedback or to achieve a flagrant effect.

Dual-mode sustain is practical and effective, but putting modes at odds magnifies the quirks of both. Cycling may intensify because gain flips between boost and reduction. Dual-mode sustain can use a single VCA receiving two control feeds, or separate VCAs; in effect two independent sustainers. In the latter case, overlap of upward and downward thresholds disposes to oscillation as the stages push against each other.

The builder working up a sustainer can use just about any VCA, but will find that certain circuits suit specific goals. Upward sustain exposes what may be the chip-VCA's worst flaw. Chips act quietly as gain reducers, but prove noticeably noisier than op amps when supplying boost, particularly if gain exceeds 20 dB. LDR-based optocouplers make VCAs out of low-
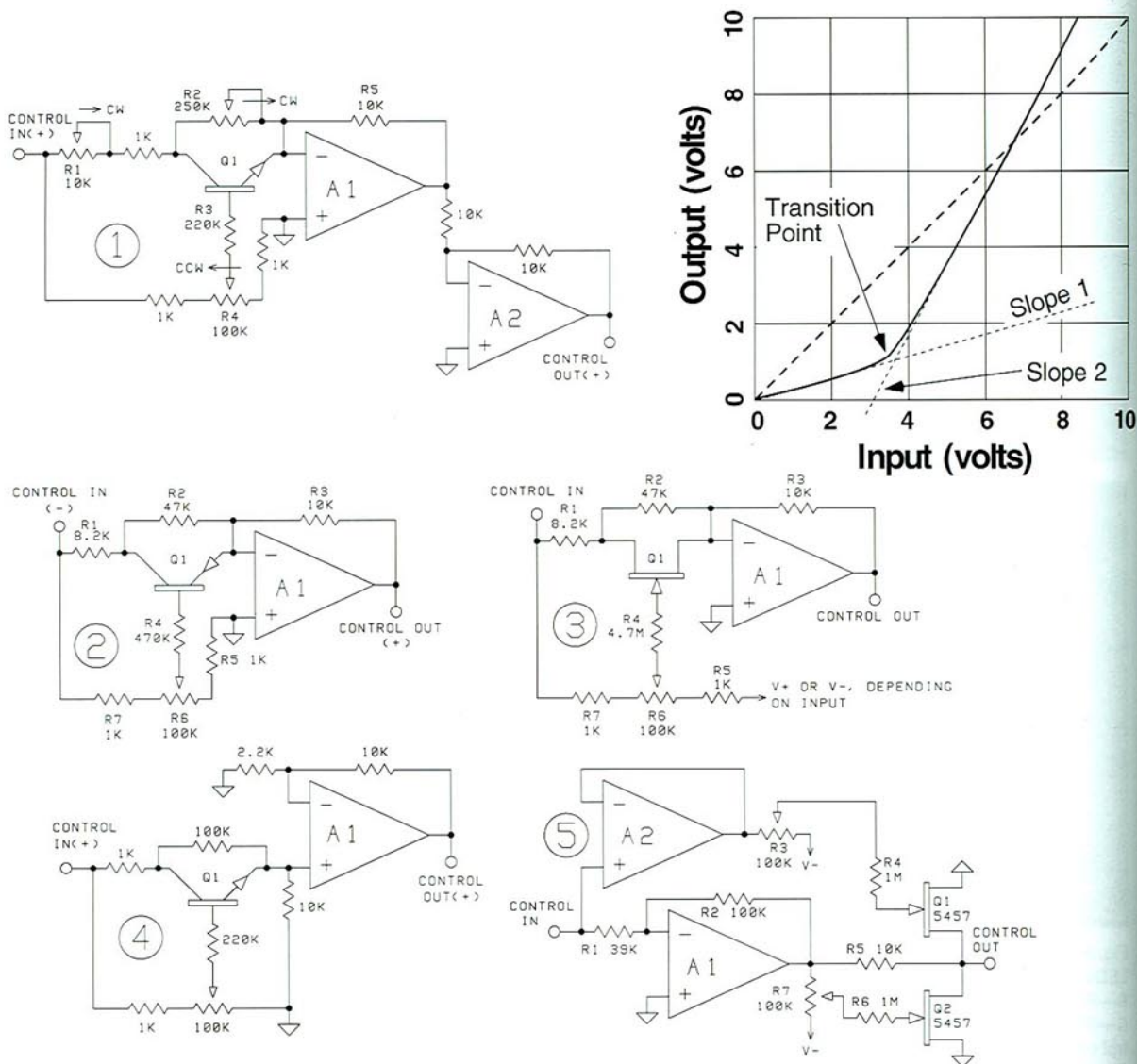
Fig. A88. Transistor-based nonlinear transfer blocks. 1—As positive input voltage rises, Q1 turns on at a point determined by setting of R4; R3 affects sharpness of knee. R2 controls Slope 1 (Q1 OFF); R1 controls Slope 2 (Q1 ON). Prior to transistor turn-on, transfer magnitude = R5÷(R2+R1+1K); once Q1 turns ON, transfer magnitude = R5÷(1K+R1). Q1's ON-resistance is negligible relative to fixed resistances. Output of A1 is inverted in A2 to clarify the process. 2—A similar setup using PNP transistor to react to negative input voltage. 3—An FET can transfer positive or negative input, but curves differ. 4—Use of noninverting amp driving output of simple external divider. 5—A more complicated setup in which FETs shunt the output of an op amp by divider action. Q1 is controlled by input, Q2 by output. This is useful, for instance, in a distortion generator. Countless variations and combos are possible, such as placing transistor in op-amp feedback loop. The simplicity, versatility, and economy of this approach makes it a prime contender in soft-knee devices, and many other stomp-box functions, such as distortion generators. Chief drawback is the need to bias each transistor, due to intersample variation.

noise op amps or discrete-transistor amps, and are preferred for sustain where the need for quiet boost predominates. The LDR's slow natural decay makes chip VCAs better candidates when dynamic response takes precedence.

Finer Points of Level Detectors

Working up a refined level detector requires the builder to define several variables beyond attack, decay, ratio, and threshold. They include:

► audio-to-DC conversion mode
► gain
► bandwidth
► attack & decay contour
► program-dependent functions

Audio-to-DC Conversion
While every level detector converts audio to DC, the change may take several forms. A *simple average* rectifies and integrates the signal, integration usually happening in the variable-decay network. Proportion-
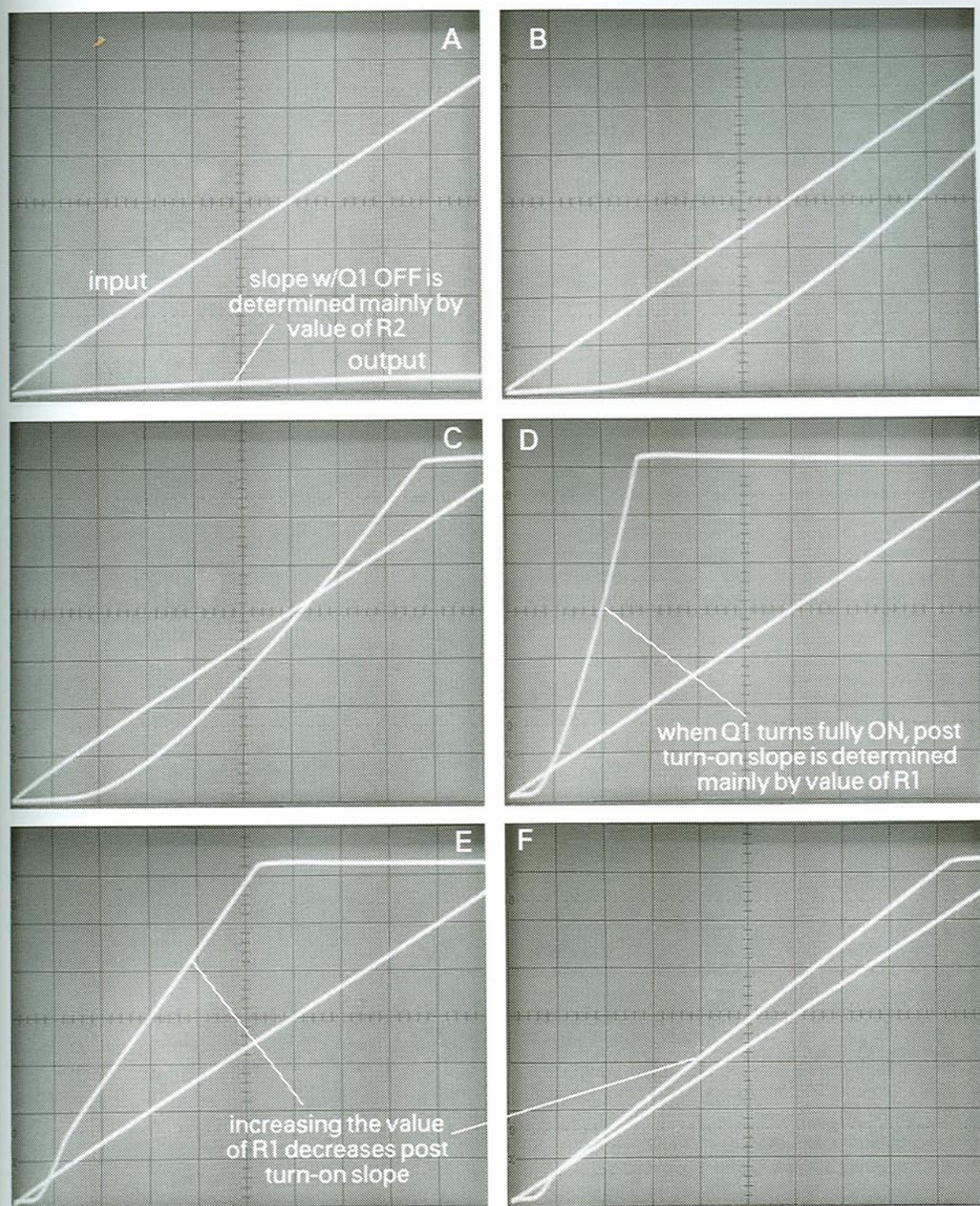
Fig. A89. Photos depict I/O of Fig. A88, schematic #1. A—R1 fully CCW, R2 fully CW, R4 fully CW; in this state Q1 is OFF; output slope = 10K÷251K. B—R4 is turned ~30% CCW; positive input voltage reaches Q1's base, turning it partly ON; output slope now equals 10K÷[({Q1's C-E resistance} in parallel with R2) + 1K]. C—R4 turned to ~40%; Q1 turns ON sooner in ramp cycle. D—R4 fully CCW; Q1 turns ON at a very low input voltage. E & F—Pot settings same as for 'D,' but value of R1 is increased, decreasing post-turn–on slope. This circuit—all circuits of Fig. A88, in fact—can produce a much greater variety of I/O curves than these few photos suggest, easily ascertained with a ramp generator and a scope. Scale all photos 2V, sweep 500 µs.
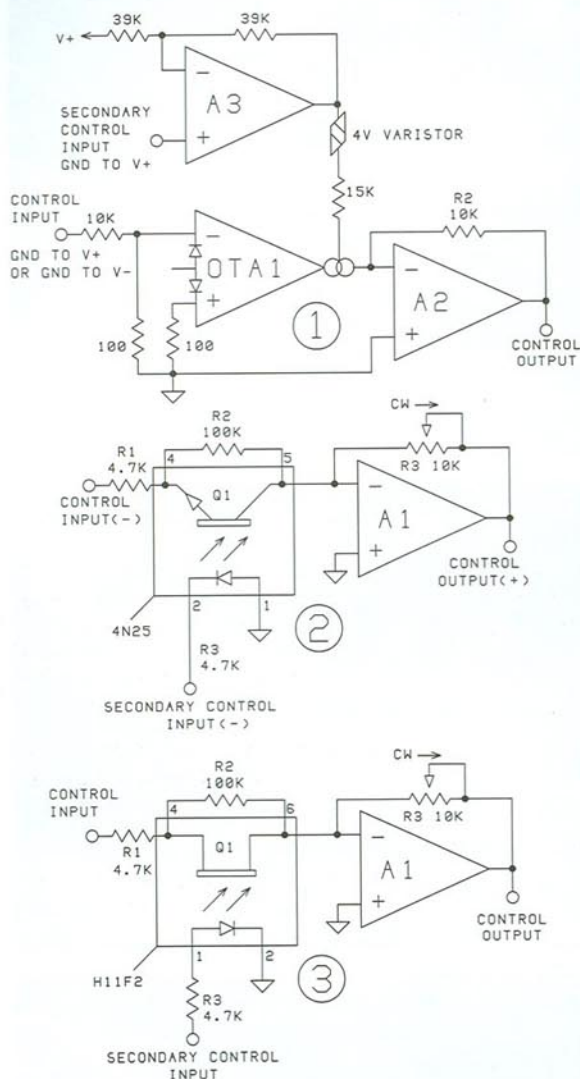
Fig. A90. 1—Level-dependent transfer block (LDTB) using low-voltage varistor. 2—LDTB based on optocoupler. When Q1 is OFF, transfer magnitude = R3÷(R1+R2). With Q1 fully ON, magnitude = R3÷R1. Reverse Q1's collector/emitter for positive input voltage. 3—LDTB using FET-based optocoupler works the same way, but can handle positive or negative voltages; secondary control must be positive.

al transfer means that a peak registers as a peak, affecting sound accordingly.

A *log average* precedes the rectifier with an AC log amp. Log conversion smooths compressor action by raising low-level sensitivity and blunting the impact of peaks. The rectified log amp's output is nominally limited to ~0.6V_{p-p}, requiring boost to yield a useful control voltage. The 'squashed' output waveform allows faster decay for the same amount of ripple compared to a simple average.

The *modified log average* gives the level detector a split personality. A resistor in series with the logging diodes allows response to follow the log curve at low levels, but to swing linear at high levels (Fig. A92–4).
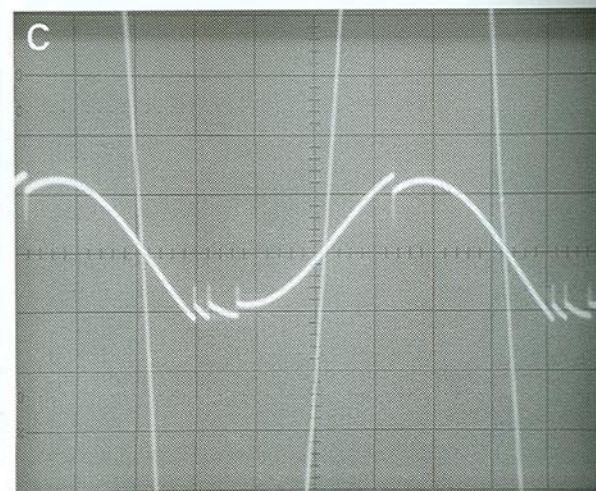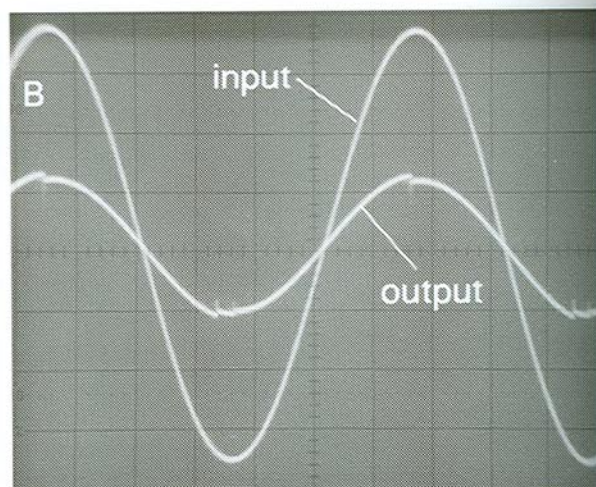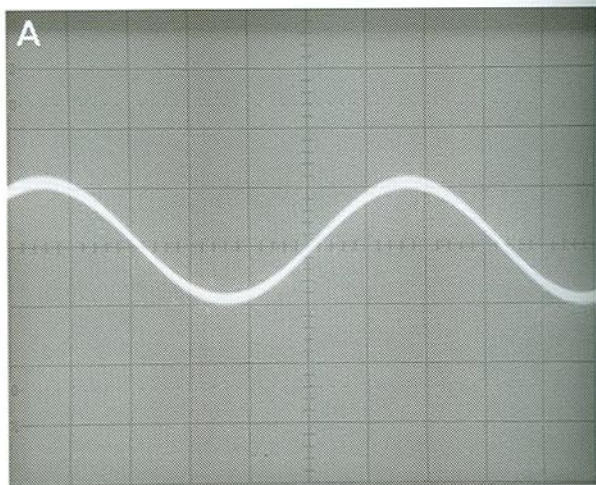
Fig. A91. I/O of reactive limiter (Fig. A94, schematic #2). A—Input superimposed on output (whose scope trace is inverted). In this photo, input level is below threshold. B—Input exceeds threshold, comparator trips, switching VCA to minimum gain; feedback path results in cycling, whose artifacts are visible at peak and trough of sinewave. C—Cycling intensifies as signal level rises. Raising the value of C1 reduces cycling at the cost of longer decay and increasingly audible limiter action. Switching artifacts are not audible on brief transients. All photos 1 KHz sinewave, scale 500 mv.
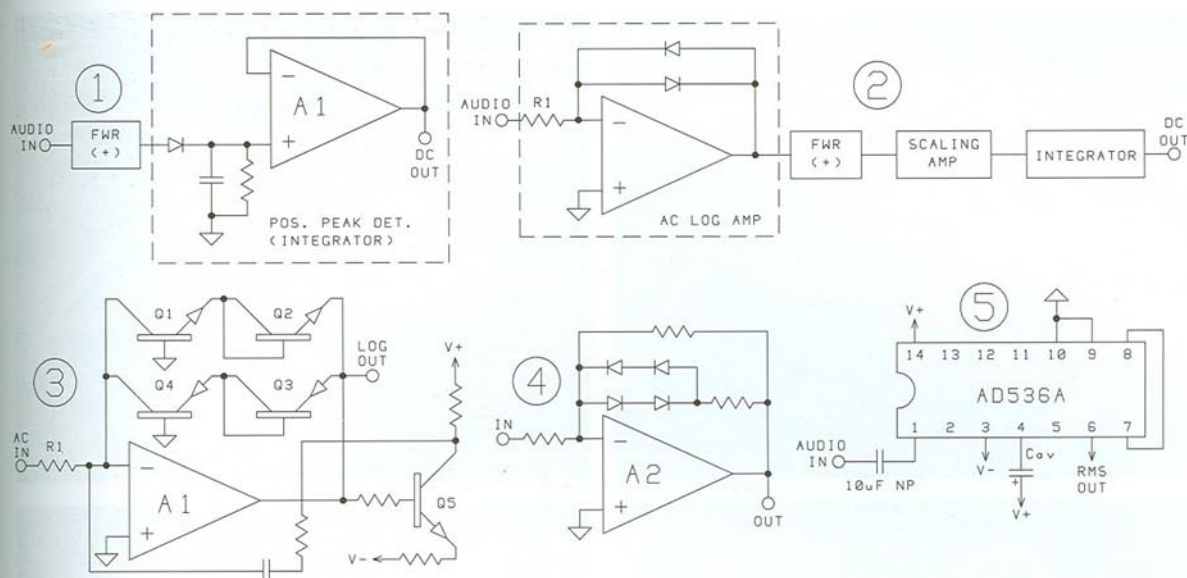
Fig. A92. Different types of AC–DC conversion. 1—Simple average is achieved by integrating the output of a fullwave rectifier. 2—Log average precedes rectifier with AC log amp, in this case a diode clipper. Raw output of log amp is nominally limited to twice the diodes' forward drop, ~1.2V_{p-p} for 1N914s, so rectifier's output will be limited to ~0.6V_{p-p}. Most circuits demand a higher control voltage, so a scaling amp often follows the rectifier. 3—While the diode clipper is adequate for most level detectors, it is neither accurate nor temperature-stable. This transistor-based circuit is more accurate, and includes positive high-frequency feedback to improve the converter's high frequency response (Ref. 43). Temperature stability requires additional circuitry. 4—A modified log amp can take many forms, one shown here. Use of multiple diodes in series, or diodes having forward drops different from those of 1N914, alters the maximum output amplitude; resistance in series with diodes tends to linearize the output once diodes conduct; resistance in parallel with network reduces low-level gain. 5—Basic AD536A RMS–DC converter circuit.

The rectifiers built into the NE570 show this pattern, which aids low-level tracking while obviating a scaling amp to obtain the high-level control voltage. A resistor in parallel with the logging diodes retains the log response to peaks but lowers gain below the diodes' forward conduction threshold. This pattern helps prevent cycling in feedback-controlled upward sustainers. Intermediate functions flow from shifting values of the input, parallel, and series resistances.

The *root mean square* (RMS) transform is given by:

$$RMS = \sqrt{\frac{1}{T}\int_0^T [f(t)]^2 dt}$$

where $f(t)$ indicates voltage, integrated over the period 0–T. Stated more simply:

$$RMS = \sqrt{\overline{V^2}}$$

Root mean square provides a power assessment that is independent of waveform shape. RMS can be thought of as the DC voltage that generates the same amount of heat in a resistor as does the audio signal. The function is realized in RMS converter chips, such as AD536 and SSM2110.

Traits of multiple transforms combine by means of a linear OR block, configured to pass the higher or lower of two or more voltages.

Each of these conversions, and others too obscure to list, flavors a dynamic effect with a slightly different sound. Whil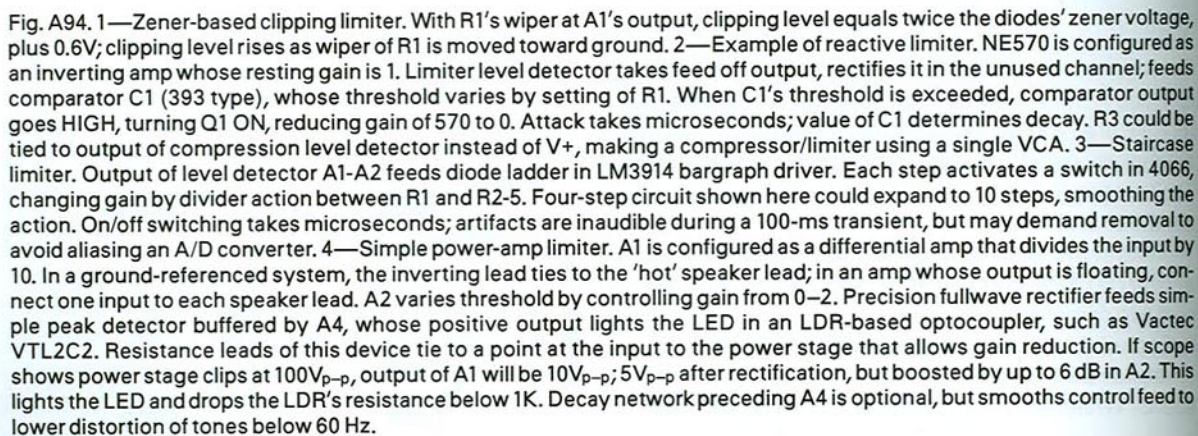e RMS compression has a reputation for smoothness, overall sound depends at least as much on control path, threshold, and ratio as on AC–DC conversion mode. In general, the log conversion cloaks compression by reducing the response to wide level swings. Exaggerated compression comes from a simple average acting in a parallel control path.

Level Detector Gain

Level detector gain is to some extent implicit in AC–DC conversion mode. Choice of gain per se depends on what the level detector has to do. A wide-band companding system needs very high gain to get a control voltage from less than 10 millivolts, to enable it to track low-level signals. An upward sustainer needs a low-gain level detector to better sense decay

Fig. A93. Left graph shows linear vs. exponential attack. Apparent that, for similar performance, linear attack needs to happen much faster. Right graph compares linear and exponential decay. For same time to total decay, linear much slower initially, yielding a smoother control voltage than does exponential decay.
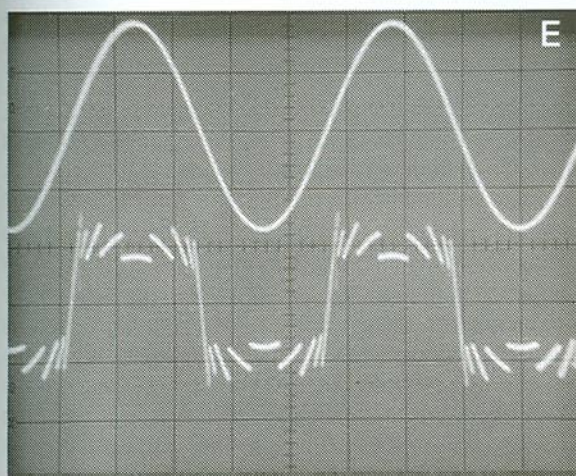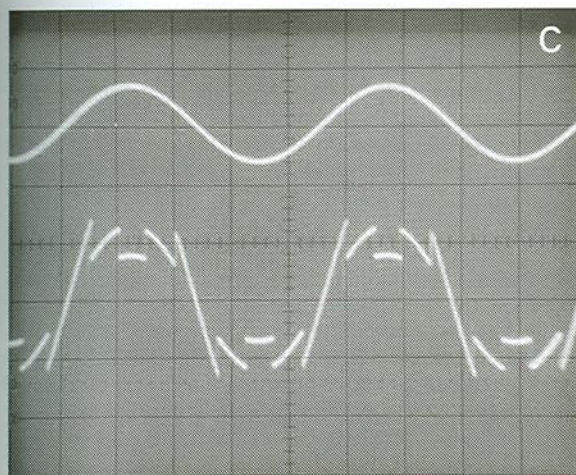

173
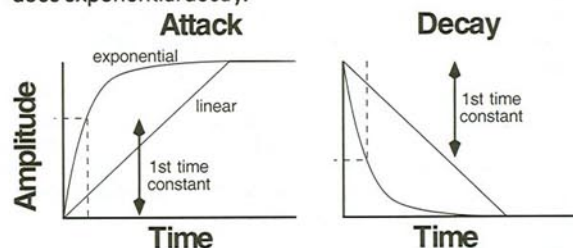
Fig. A94. 1—Zener-based clipping limiter. With R1's wiper at A1's output, clipping level equals twice the diodes' zener voltage, plus 0.6V; clipping level rises as wiper of R1 is moved toward ground. 2—Example of reactive limiter. NE570 is configured as an inverting amp whose resting gain is 1. Limiter level detector takes feed off output, rectifies it in the unused channel; feeds comparator C1 (393 type), whose threshold varies by setting of R1. When C1's threshold is exceeded, comparator output goes HIGH, turning Q1 ON, reducing gain of 570 to 0. Attack takes microseconds; value of C1 determines decay. R3 could be tied to output of compression level detector instead of V+, making a compressor/limiter using a single VCA. 3—Staircase limiter. Output of level detector A1-A2 feeds diode ladder in LM3914 bargraph driver. Each step activates a switch in 4066, changing gain by divider action between R1 and R2-5. Four-step circuit shown here could expand to 10 steps, smoothing the action. On/off switching takes microseconds; artifacts are inaudible during a 100-ms transient, but may demand removal to avoid aliasing an A/D converter. 4—Simple power-amp limiter. A1 is configured as a differential amp that divides the input by 10. In a ground-referenced system, the inverting lead ties to the 'hot' speaker lead; in an amp whose output is floating, connect one input to each speaker lead. A2 varies threshold by controlling gain from 0–2. Precision fullwave rectifier feeds simple peak detector buffered by A4, whose positive output lights the LED in an LDR-based optocoupler, such as Vactec VTL2C2. Resistance leads of this device tie to a point at the input to the power stage that allows gain reduction. If scope shows power stage clips at $100V_{p-p}$, output of A1 will be $10V_{p-p}$; $5V_{p-p}$ after rectification, but boosted by up to 6 dB in A2. This lights the LED and drops the LDR's resistance below 1K. Decay network preceding A4 is optional, but smooths control feed to lower distortion of tones below 60 Hz.
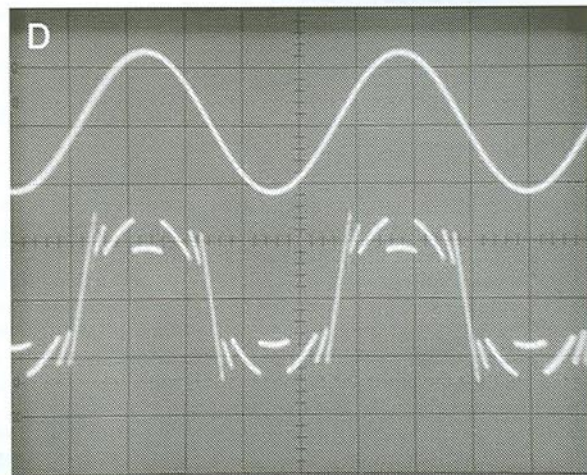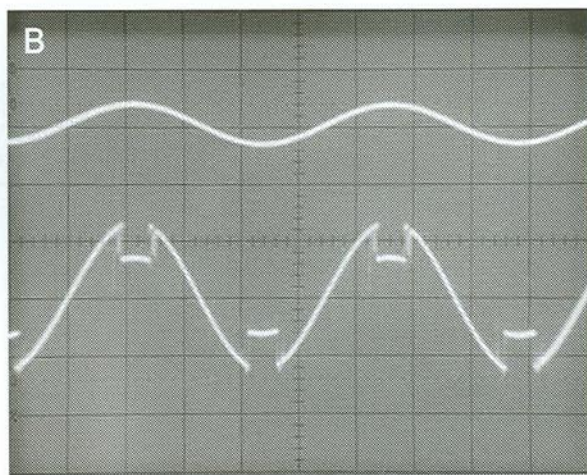
Fig. A95. I/O of staircase limiter, Fig. A94, schematic #3. Threshold is ~1.3V_{p-p}. In photos B–E, a new notch appears as each comparator's threshold is passed. Rough appearance is due to relatively few stages, and is easily smoothed by adding more stages. Switching times measure in microseconds; the ultrasonic artifacts could alias an A-D converter. All photos this page scale top trace 2V, bottom trace 500 mv; sweep 1 ms.

of the program, and to resist cycling.

Level Detector Bandwidth

Just about any signal that enters the level detector generates a control voltage. This is sometimes desirable. A compressor built for spot use may be called upon to compress narrowband audio anywhere between 20 Hz and 20 KHz, and to do so according to the control settings. This demands that the level detector's bandwidth equal that of the signal path. Unrestricted response proves undesirable when it allows one audio band to generate the bulk of the control voltage. For example, in a wideband feed, strong bass might audibly modulate treble. Wideband dynamic controllers, such as companding systems, sound more natural with a bandpass filter placed ahead of the level detector to emphasize frequencies to which the ear is most sensitive. Stomp-box compressors tuned for specific instruments attain part of their singularity from such filters.

Some signals always need exclusion, such as the 19-KHz FM radio pilot tone, scan frequencies of TV and computer monitors, AM radio signals, etc. Filters to exclude these bands are found in some rackmount pieces, but are rarely required in pedals.
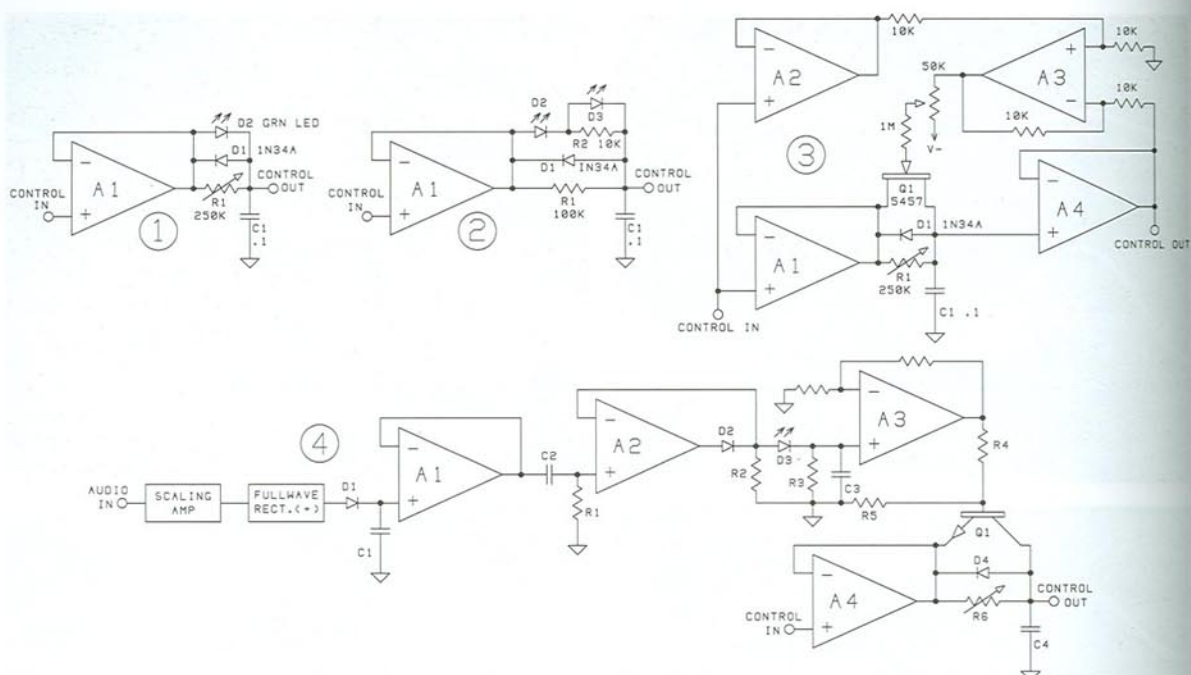
Fig. A96. 1—Simplest program-dependent attack circuit adds diode D2 in parallel with standard attack network D1/R1/C1; assume R1 at 250K. If program audio attacks slowly, C1 has time to charge, such that no significant differential voltage develops between D2's anode and cathode. If program attacks quickly, and has sufficient amplitude, C1 won't have time to charge, and the difference voltage reaches D2's forward conduction threshold, leading to instantaneous attack that lasts until the charge in C1 builds up. Design is sensitive to both amplitude and rate of change. Different thresholds can be chosen by selection of diodes having different forward drops. 2—A variant on this theme uses series resistance to create a three-step attack. When differential voltage exceeds D2's drop, attack accelerates by a factor of 10, due to the time constant of R2 with C1. If transient is big and fast enough, the differential across R2 surpasses D3's forward drop, giving instantaneous attack. 3—FET-based circuit is more complex. Subtraction of input from output by A3 yields an error voltage whose amplitude varies with amplitude and rise time of input pulse. The positive error voltage turns Q1 on, accelerating attack until error voltage falls below turn-on point. This circuit allows variable threshold by setting of R1. 4—Running a leading-edge detector in parallel with the main control path. Audio is conditioned in a scaling amp; converted to DC by fullwave rectifier, then peak-detected by D1-C1, which smooths out most of the ripple while not materially slowing the leading edge of the program. A1 output feeds differentiator C2-R1; augmented diode D2-A2 allows only the positive voltage to pass. Because very low frequencies leave ripple in the control feed at that point, the signal is further detected by an LED, whose high forward drop essentially eliminates the ripple. The resultant positive control voltage feeds scaling amp A3; thence to R4-R5, whose values determine Q1's sensitivity. Fast attack in the program generates a positive pulse that turns Q1 on, speeding attack over the time constant of R6-C4. The control voltage could also drive a phototransistor in an optocoupler. (In circuits 1–3, it is important that the signal feeding the auto-attack network have been integrated, as by passage through a variable-decay network.)

Attack & Decay Contour

Attack and decay networks govern their respective functions by varying the time taken to charge or discharge a capacitor. Charging through a resistor leads to an exponential curve, characterized by fast rise slowing to a plateau. The curve results because the cap's voltage changes as it charges. The time constant of the RC pair tells the time needed to charge or discharge 63.2% of the total voltage.

A linear curve attends charging or discharging through a constant-current path, which frees current flow from dependence on voltage. Thus, attack and decay are easily configured for linear or exponential modes (Fig. A93). Modes differing from these are realized by feeding the output of a standard network through a nonlinear transfer block.

Control contour matters most in decay, because decay chiefly determines control-voltage ripple. For the same time from a fully charged to a fully discharged state, linear decay changes less and thus integrates more than exponential decay. Linear decay also happens to allow long decay using small capacitors, and facilitates voltage control of the function. Exponential decay is preferred for a voltage meant to mimic decay of a note, because picked or struck strings normally decay exponentially.

Program-Dependent Functions

Ideal compression happens discreetly, yet confines the signal to the bounds of the medium. Most studio vets have learned that some types of program unmask the process or defeat the control settings. In the case of decay, a loud peak followed by a soft passage exposes compressor action if decay has been set
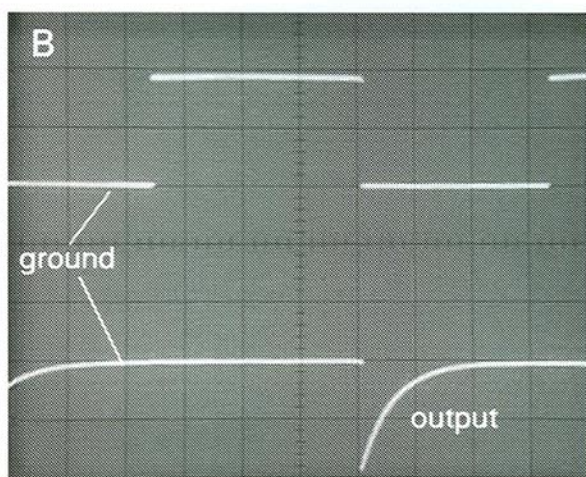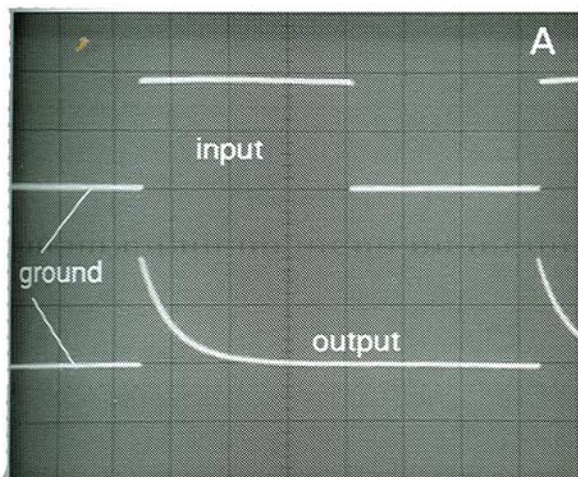
Fig. A97. Schematic illustrates one way to distinguish the leading from the trailing edge of a signal. A—Squarewave simulates program with infinitely fast attack and decay. A1 buffers the pulse, which feeds simple differentiator C1-R1. A2 is configured as an augmented diode that passes only positive voltage; output is a purely positive pulse coincident with attack; does not react to falling edge of program. B—A3 is configured as an augmented diode that passes only negative voltage; reacts only to falling edge of pulse; output is a negative voltage, even though input pulse is purely positive. Musical signals exhibit slower attack and much slower decay, but detector generates useful control voltage in both cases.
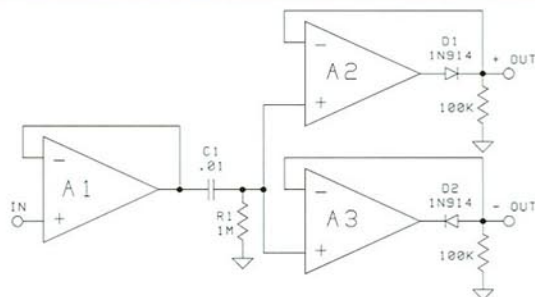
longer than ~500 ms; a *pianissimo* rebounds audibly during that time. In the case of attack, the slight delay often used to retain a sense of dynamics leaves the track vulnerable to overload by the loud, fast transient, such as a rimshot. These instances crave a level detector that senses the event and departs from fixed attack and decay.

The simplest program-responsive attack circuit keys on the voltage difference that develops between the input and output of an attack-delay network (Fig. A96–1). If the program attacks slowly, the attack cap has time to charge, and only a small difference voltage develops. If the program attacks fast, the cap does not have time to charge, and a large voltage develops. A diode in parallel with the attack resistor conducts when this difference voltage exceeds its forward drop. At that point attack becomes instantaneous if the raw diode is used, or accelerates by some fraction if the diode sees a series resistance smaller than the main attack resistance. The voltage required to trigger instant attack is varied by using diodes having different forward drops. 'Stacking' diodes and resistors in parallel with the attack resistor gives a network whose attack varies in three or more steps (Fig. A96–2). A related approach uses the difference voltage to activate a transistor wired in parallel with the attack resistor (Fig. A96–3). This method constitutes a feedback control loop that continuously tunes attack to the pro-

gram. A third option uses a leading-edge detector running in parallel with the level detector to generate a control voltage that, through a control block, varies attack according to the content of the program (Fig. A96–4).

Program-dependent decay requires sensing a rapid fall in level. One method runs two decay networks in parallel, one of which decays several times faster than the other (Fig. A100–1). Both networks feed a subtractor. If the program decays slowly, little voltage difference develops. But when a soft passage immediately follows a peak, the disparity in decay creates a large voltage difference that hastens decay of the main network. Another approach uses a falling-edge detector. Differentiation yields a positive voltage from the leading edge of a transient, but a negative voltage from the falling edge, even when the transient swings purely positive. A diode discriminator selects the negative voltage, which then speeds decay (Fig. A100–2). The falling-edge detector demands moderate pre-integration of the signal. Both approaches are practical, easily adjusted, and mate smoothly to the stock linear-decay network.

A 'decay' version of the diode method of auto-variable attack proves practical. The approach combines both auto-variable functions in a block of remarkable simplicity (Fig. A102). The output of this network tracks the music envelope closely, but without the severe ripple that accompanies fast attack and decay.
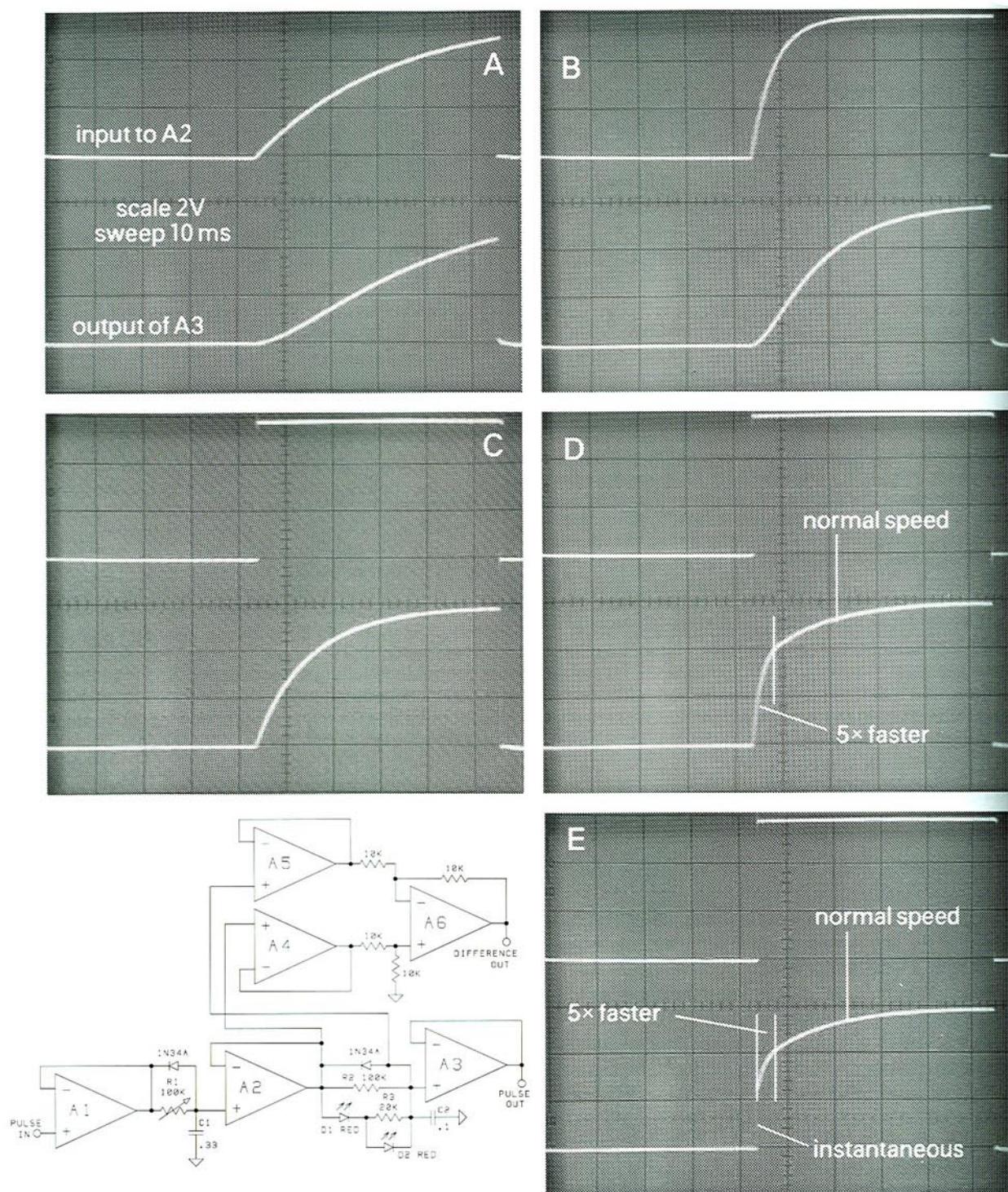
Fig. A98. Schematic shows auto-variable attack network D1-D2-R2-R3-C2, whose output is buffered by A3. Feeding this network is an artificial attack generator, R1-C1, which variably slows the leading edge of a positive pulse feeding A1. Because C1 is three times the value of C2, the attack of the input can be slowed below the time constant of R2-C2. In photos A–C, D1 & D2 have been removed from circuit. Photo A shows slow leading edge is further slowed by R2-C2; attack speeds up in photos B & C, but output of A3 is slowed by time constant of R2-C2. In photo D, D1 has been plugged into circuit; now output of A3 exhibits two phases; initial portion rises five times faster than second portion due to the smaller time constant of R3 with C2. In photo E, D2 has been added. Now output shows three distinct phases; when rate of attack and pulse amplitude are sufficient to overcome D2's forward drop, attack becomes instantaneous, then slows to time constant of R3-C2, finally slowing to time constant of R2-C2. Verifying action of circuit is easy. Dim the lights in the lab and watch D1 & D2 as attack is accelerated by lowering the value of R1; first D1, then D2 flashes with each pulse.

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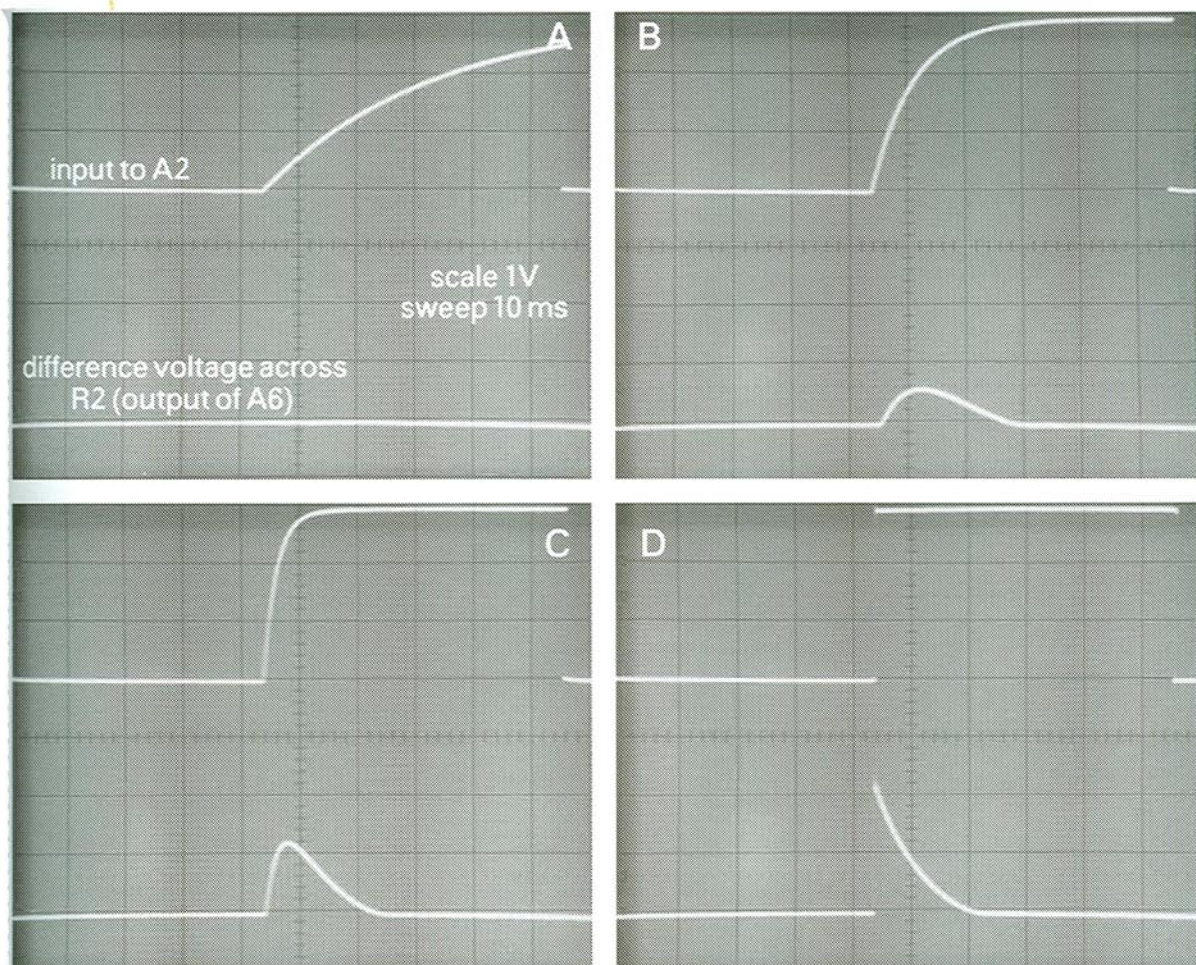
Fig. A99. Another way of looking at auto-variable attack. This series of photos illustrates the difference voltage that develops between the ends of the attack resistor, R2 in Fig. A98. Top trace input, bottom trace difference voltage. A—Input pulse rises more slowly than the time constant of R2-C2; C2 cap has time to charge, no difference voltage develops. B & C—As attack accelerates, a difference voltage appears and grows. D—Instantaneous attack generates a comparably fast difference voltage. This voltage can be used to trigger a switch across the attack resistor, speeding attack when the network senses fast attack in the program. Concept adapts to parallel or feedback control.
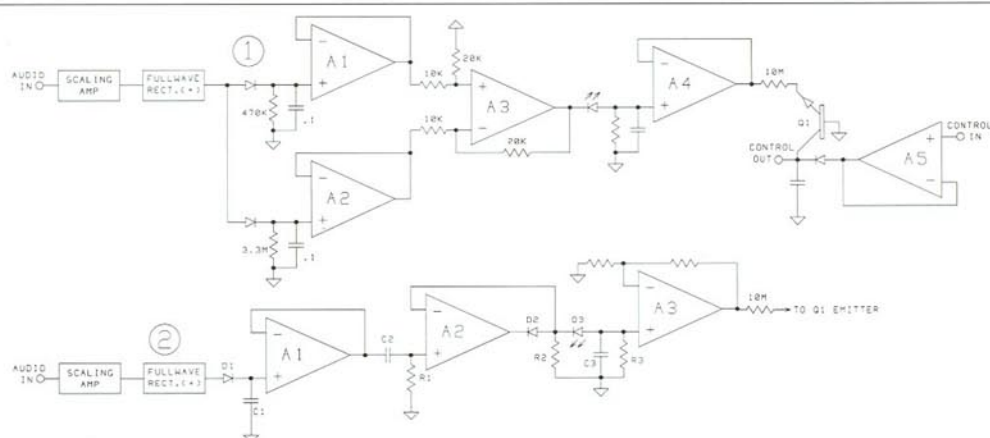
Fig. A100. 1—One means to achieve program-variable decay. Output of scaling amp/positive rectifier feeds two peak detectors, identical but for their decay rates. Because both detectors see identical amplitudes, subtractor A3's output is proportional mainly to the rate of change between the two decay networks. When program decays slowly, little differential voltage exists. The quicker the program decays, the greater the negative voltage generated, and the faster the decay. This voltage applies directly to 10M resistor tied to Q1's emitter, or through a summing amp that allows manual decay control which will shift when program decays quickly. Also, a pot tied between A4's output and ground would let the user control the magnitude of the auto-variable effect. 2—Similar results realized by a falling-edge detector. But for polarity of the control voltage output, circuit functions identically to the leading-edge auto-variable–attack circuit previously described.

A
input 1V/div.
output 200 mv/div.
sweep 20 ms

B

C
1V/div.

D
1V/div.
200 mv/div.
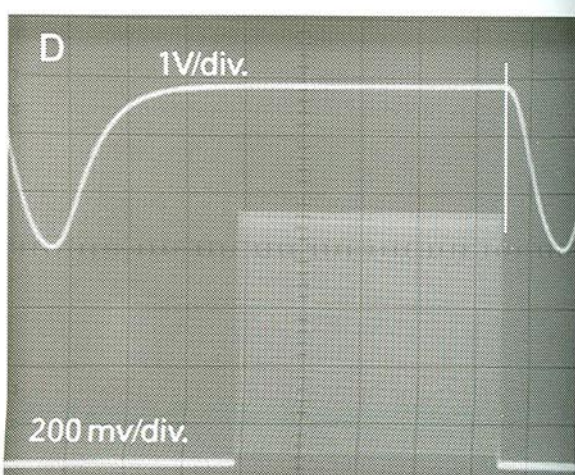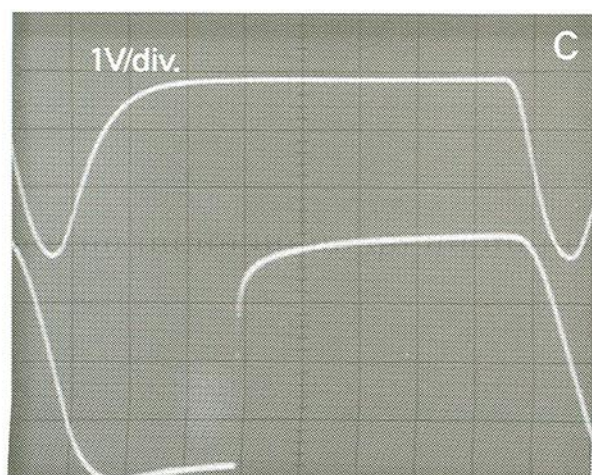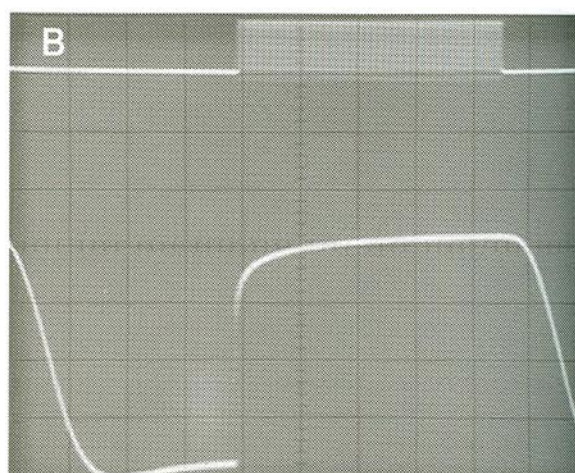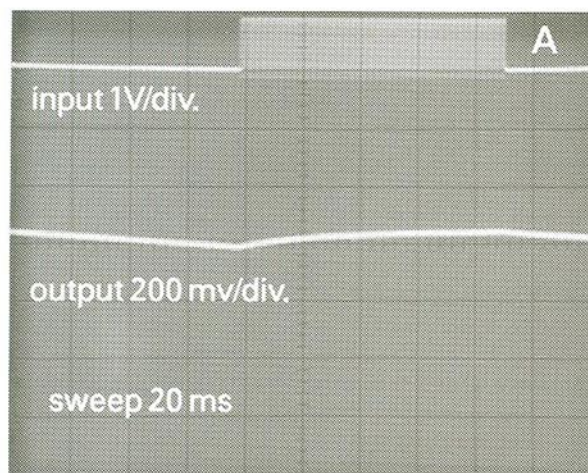
Fig. A101 (right). Schematic illustrates a linear decay circuit (integrator) with switchable auto-variable decay. When S1 ties R1 to ground, positive charge stored in C1 decays linearly, over ~4 seconds. In photo A, top trace shows tone burst output of fullwave rectifier; bottom trace shows integrator output; long decay results in little change between bursts. In photo B, S1 selects output of A3; now decay takes less than 40 ms. In photo C, top trace shows output of A3, a negative voltage produced only when the program decays quickly. This voltage causes a negative current to flow through Q1, neutralizing the positive charge in C1. The faster the program decays, the larger the output of A3, and the faster the decay. D—Fullwave rectified input burst now appears on bottom trace; illustrates that the decay-sensing network reacts only to falling edge of program.
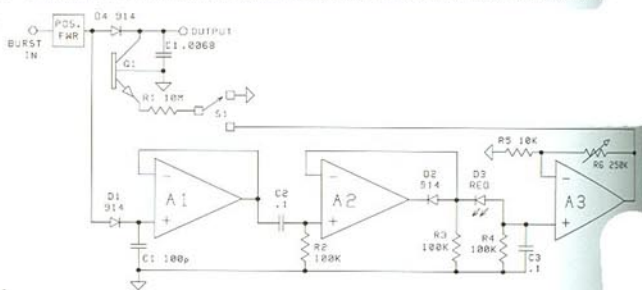
Fig. A102 (left). Auto-variable attack & decay in one simple stage. Key step involves pre-integration of rectified audio by D1-C1, buffered by A1 (064 or similar FET-input type). Charge in C1 leaks off by various paths, giving predominantly linear decay. This provides substantial integration, despite decay on the order of 80 ms. At 20 Hz, ripple at output of A1 is $600mv_{p-p}$, against rectified input amplitude to D1 of $5.5V_{p-p}$; ripple in the control voltage output is only $\sim30mv_{p-p}$, becoming negligible above 40 Hz. Auto-variable attack involves R1-R2-D2/3/4 and C1, by mechanisms previously described. The only decay path for the charge in C1 is through D5 and R3, which has a long time constant with C1. However, if the program decays quickly, the voltage at A1's output falls, creating a voltage differential across R3; if that difference exceeds the forward drop of D6, decay tracks that of the input signal, down to D6's forward drop, at which point the time constant of R3-C2 determines remaining decay. Circuit is easily tuned to specialized applications by changing diode types, or the values of resistors and caps.
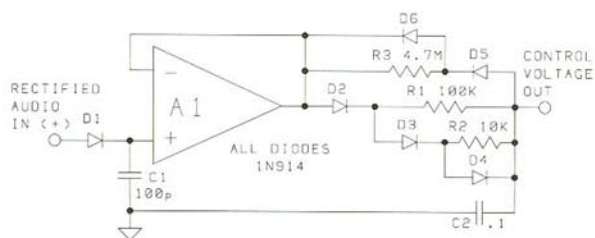
Project No. 33
Burst-O-Matic

Designing and testing compressors and expanders is greatly aided by an audio-band tone burst generator, especially one that allows control of burst depth, to reveal the effect of attack and decay on musical transients. As a bonus, Burst-O-Matic throws in a pulse output to aid testing pure attack and decay networks.

Circuit Description

IC1-a & -b and associated components form a sine-wave generator whose frequency measures ~5 KHz. IC1-b output couples through R8 to one terminal of a CMOS switch inside IC3. The other terminal of this switch ties to an artificial ground produced by IC1-c and associated components. Signal couples through pot R9 to input of voltage follower IC2-a and to the output path through R10-C4.

IC1-d supplies the bias reference used by IC1-a/-b and IC2-a.

IC4 is configured as a variable-rate squarewave generator whose output controls the switching rate of IC3. When IC4 pin 5 goes HIGH, switch IC3 turns on, shunting most of the signal present at IC3 pin 1. When IC4 pin 5 goes LOW, switch opens and allows most of the sinewave burst to pass to R9 and the output path. The setting of R11 determines how much signal gets shunted, and thus controls burst depth.

Raw IC4 output also couples to voltage follower IC2-b, through R12-R13 to the pulse output.

Fig. 33–1. Burst-O-Matic prototype board. C4 in this early version was an electrolytic cap; current version uses 0.01 μF monolithic.

C4

Use
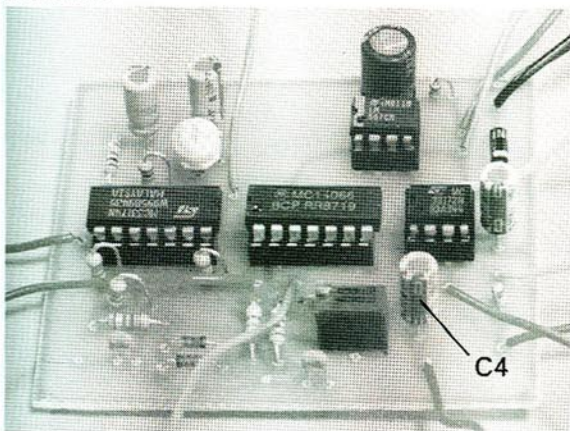
Pots have these functions:

R9	burst output level
R11	burst depth
R12	pulse output level
R17	pulse/burst rate

Initial settings: R9, 11, 12, 17 fully CW. Place trimpot R16 in center position. First, trim the sinewave generator. Power up, connect scope probe to pin 14 of IC1, trim R6 to give a sinewave of ~4V$_{p-p}$. Trim R16 to give symmetrical burst output (Fig. 33–4A, top trace; Figs. 33–4C & D show output w/pot improperly trimmed). Take all panel pots through full range and note effects on pulse & burst outputs.

BURST-O-MATIC PARTS LIST
Resistors
R1 4.7K
R2, 3, 7 2.2K
R4 47K
R5 1.8K
R6 1K multiturn trimpot
R8 100K
R9 1M audio-taper pot
R10 100
R11 100K pot
R12 5K audio-taper pot
R13 470
R14, 15 10K
R16 10K trimpot
R17 10K pot
R18 1K
Capacitors
C1, 2 0.01 μF, 5% or better
C3, 5, 7 10 μF aluminum electrolytic
C4 0.01 μF monolithic, 20%
C6 47 μF aluminum electrolytic
Semiconductors
D1, 2 1N914
D3 1N4001
IC1 MC33174 quad low-voltage op amp
IC2 MC33172 dual low-voltage op amp
IC3 4066 quad bilateral switch
IC4 NE567 tone decoder
Miscellaneous
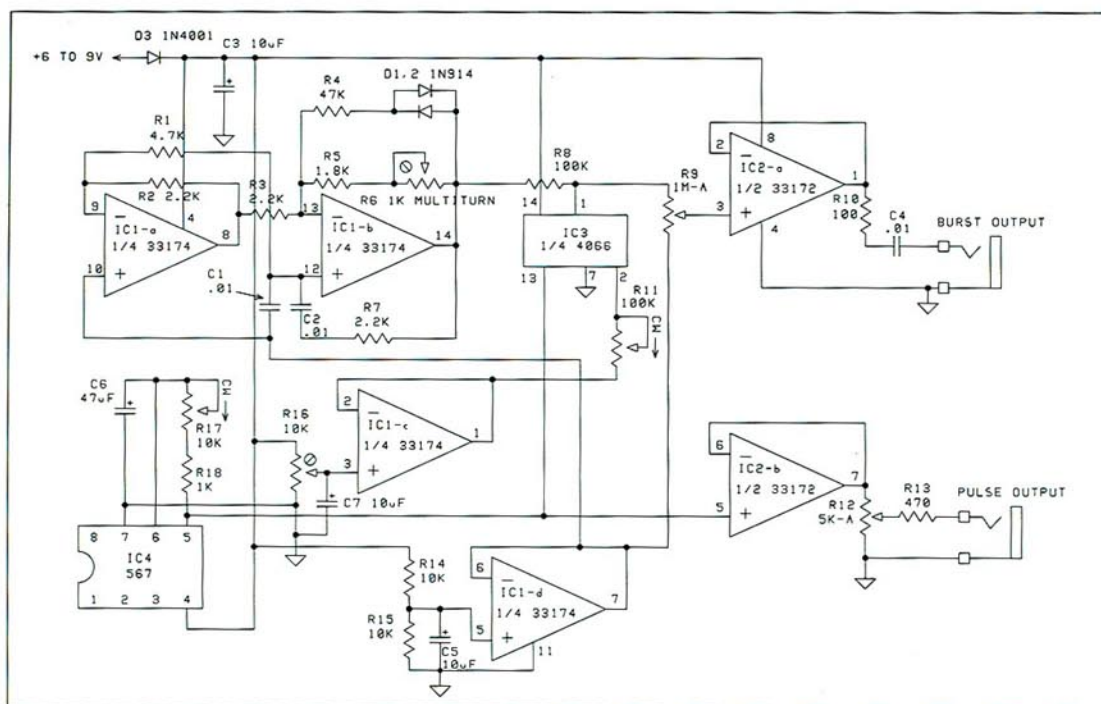circuit board, solder, wire, jacks, etc.

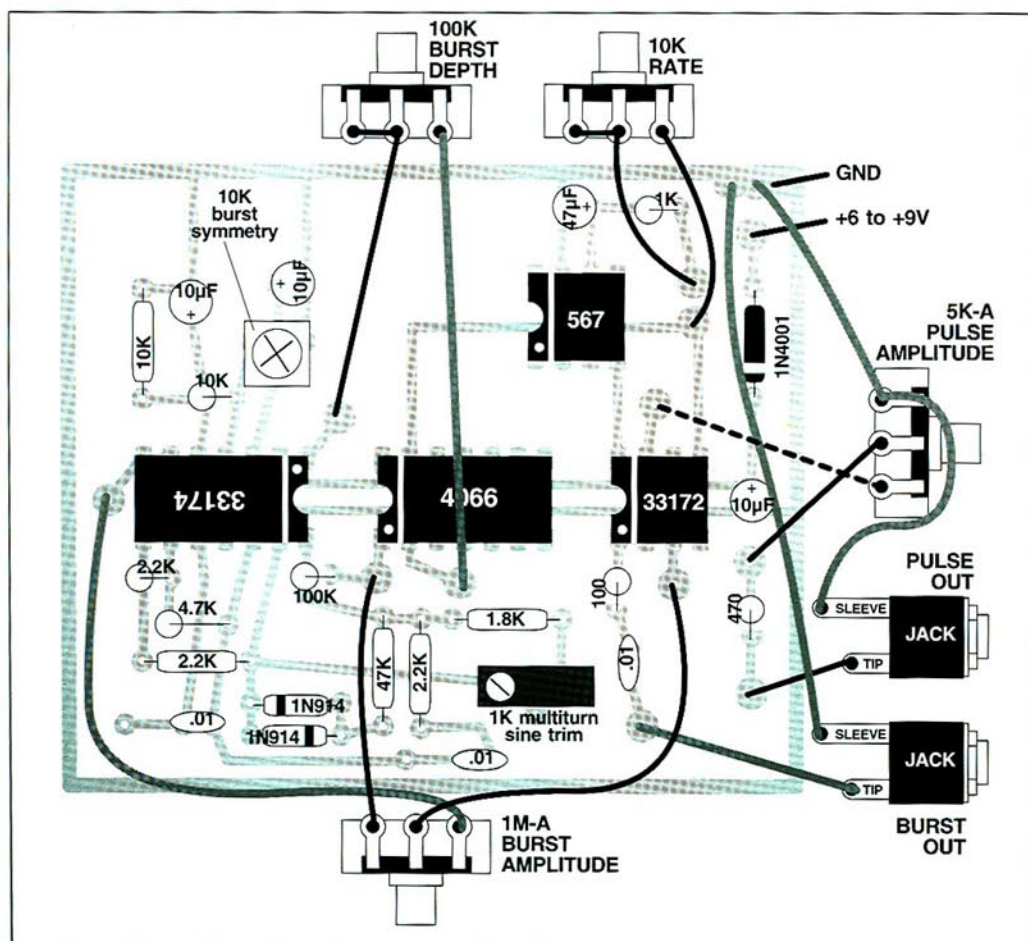Fig. 33–2. Burst-O-Matic schematic.
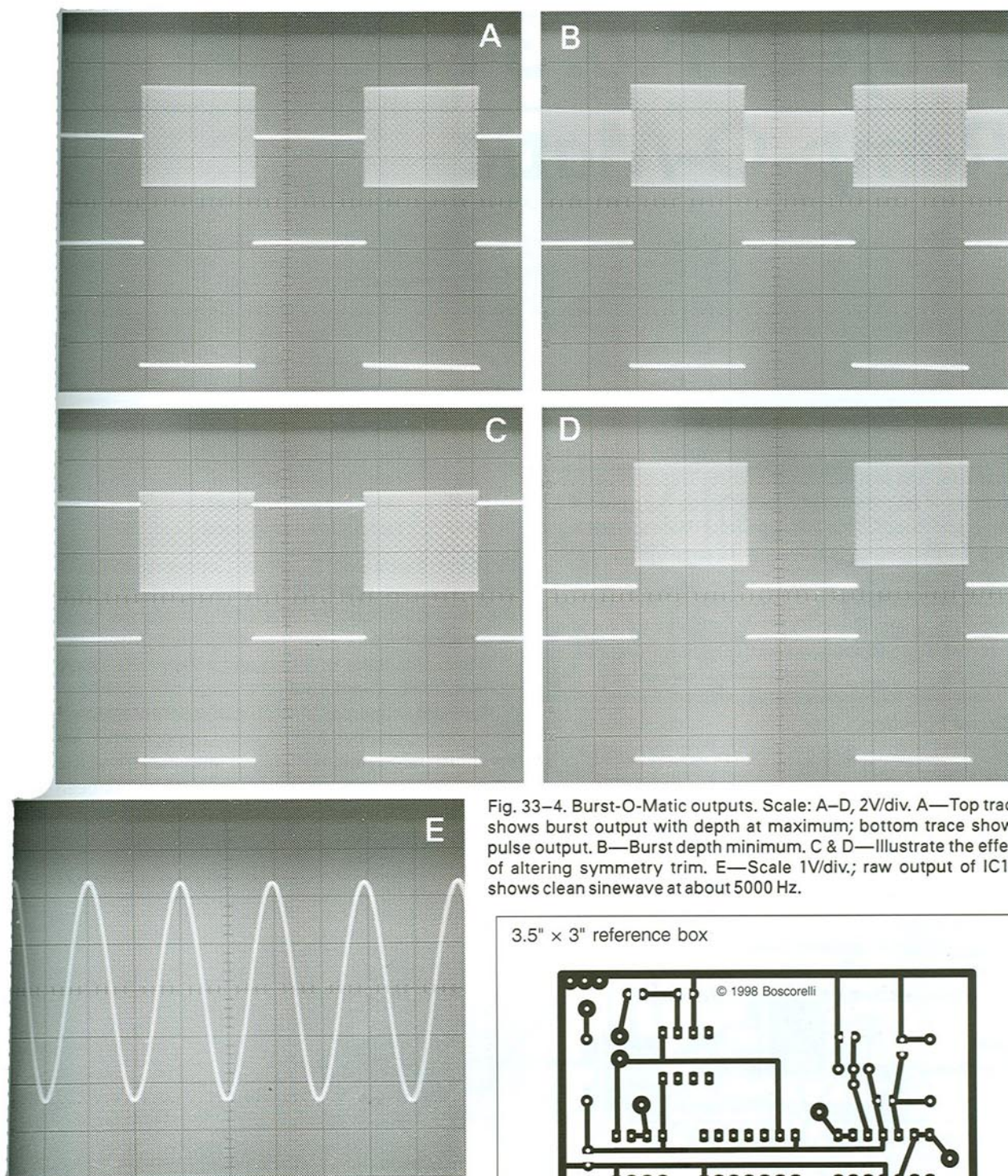
Fig. 33–3. Burst-O-Matic layout & wiring diagram.

Fig. 33–4. Burst-O-Matic outputs. Scale: A–D, 2V/div. A—Top trace shows burst output with depth at maximum; bottom trace shows pulse output. B—Burst depth minimum. C & D—Illustrate the effect of altering symmetry trim. E—Scale 1V/div.; raw output of IC1-b shows clean sinewave at about 5000 Hz.
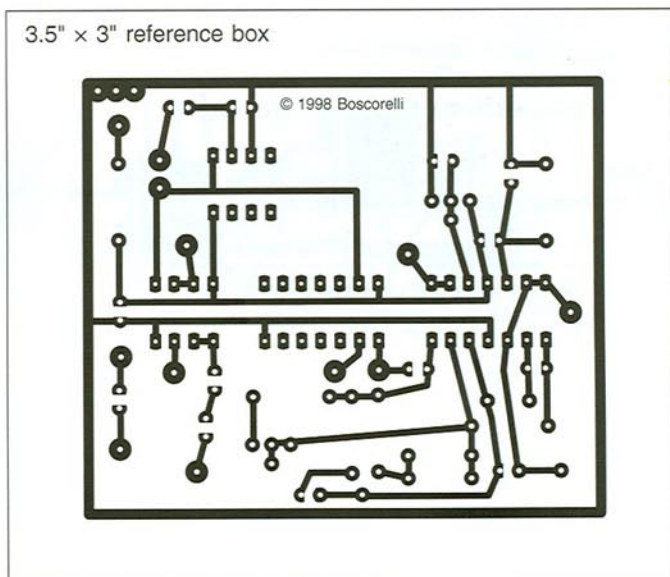
3.5" × 3" reference box

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Fig. 33–5. Burst-O-Matic circuit board.

Project No. 34
Ramp-O-Matic

One way to work up a nonlinear transfer block is to measure and plot individual points. This approach is tedious. A quick, intuitive option feeds the network a ramp and displays the I/O on a scope. What could take hours of plotting can be done in minutes. All you need is a ramp generator. Tuning nonlinear transfer blocks (especially complex, direction-changing designs) becomes a snap. Ramp-O-Matic is great for testing those what-if networks, and reveals discontinuities at a glance.

Circuit Function

Q1–3 and associated components form a variant on a well-known linear positive ramp generator. R2 varies frequency. Ramp output is taken off Q3's emitter, coupling directly to noninverting input of IC1-a, which applies gain of ~4. R8 nulls the offset at IC1-d's output, which is due to the offset that exists at Q3's emitter.
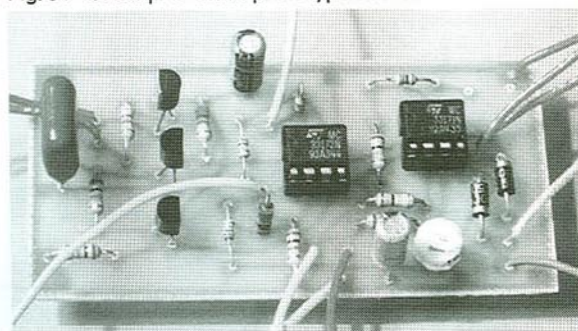
IC1-a output couples through R10 to unity-gain inverting buffer IC1-b. SPDT switch S1 selects the positive or the negative ramp, feeding it to pot R17, which varies ramp amplitude. R12 controls output DC offset. Ramp couples through R15 to the output path.

Use

Switches & pots have these functions:

S1	ramp select positive/negative
R2	ramp frequency
R8	ramp internal DC trim
R12	ramp output DC offset
R17	ramp output amplitude

First, trim the internal DC offset. Connect output pin 1 of IC1 to scope, power up the device, trim R8 so that the ramp is equidistant from IC1's positive and negative output limits.

Ramp amplitude and frequency vary with supply voltage; the box works over the range ±7.5 to ±15V.

Connect scope probe to output, confirm action of S1 by noting positive and negative ramp; confirm action of R2 ramp rate, R17 output amplitude, and R12 output DC offset. Ramp-O-Matic's utility has been demonstrated earlier in this Appendix.

Fig. 34–1. Ramp-O-Matic prototype board.

RAMP-O-MATIC PARTS LIST
Resistors
R1, 3 4.7K
R2 10K pot
R4, 5, 6 10K
R7, 10, 11 22K
R8 100K trimpot
R9 30K
R12, 17 100K pot
R13, 14, 16 100K
R15 100
Capacitors
C1 0.33µF
C2, 3 10µF aluminum electrolytic
Semiconductors
D1, 2 1N4001
IC1 MC33172 dual op amp
IC2 MC33171 op amp
Q1 2N5457 field-effect transistor
Q2 2N2646 unijunction transistor
Q3 2N3904 NPN transistor
Miscellaneous
S1 SPDT switch
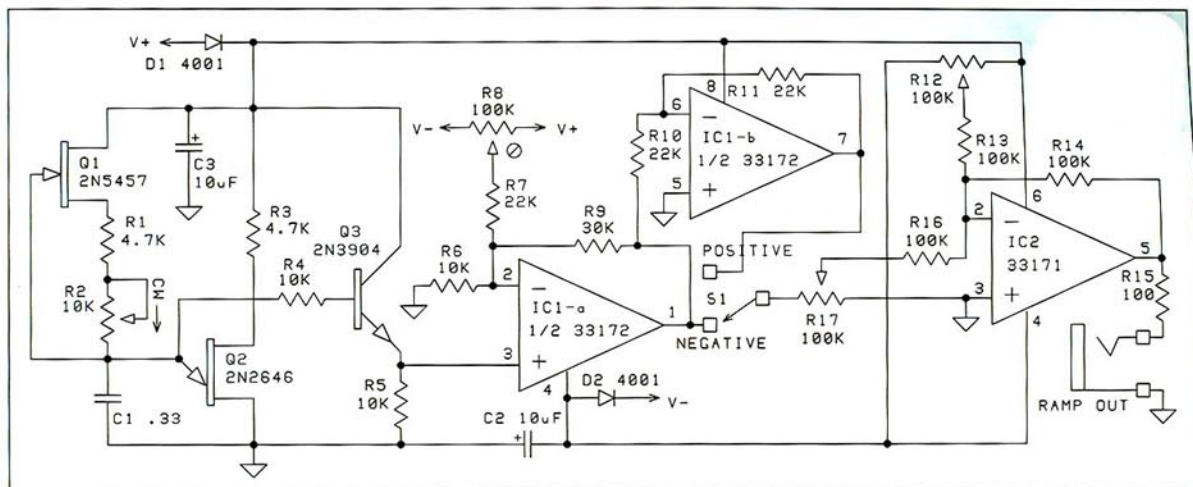wire, solder, circuit board, output jack, etc.

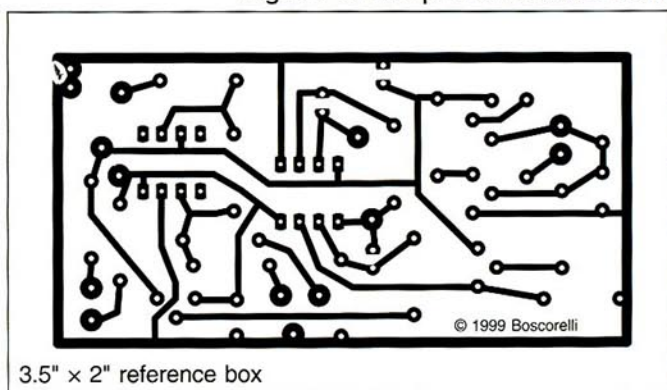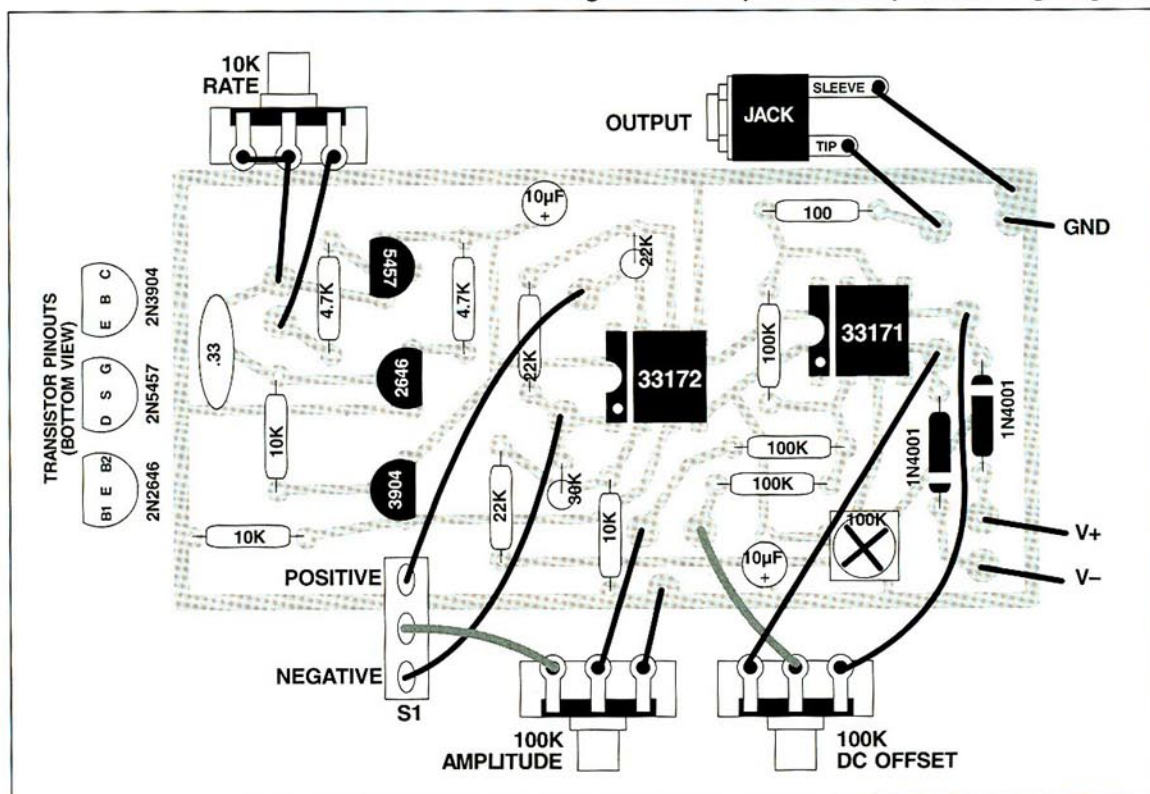Fig. 34–2. Ramp-O-Matic schematic.

Fig. 34–3. Ramp-O-Matic circuit board.

© 1999 Boscorelli

3.5" × 2" reference box

Fig. 34–4. Ramp-O-Matic layout & wiring diagram.

Distortion

Distortion spawned the industry of which this book is an outgrowth. In fact, fuzz paved the way for a galaxy of pedals now gracing the market. The effect spans the gamut from sweet—Nokie Edwards' lead on "Only The Young"—to the fresh-paper-cut edge of "Who Are The Brain Police?" Like 'verb, a touch of distortion rounds out the sound, imbuing tone with hints of strange instruments. Stomp boxes dazzle by how tastily they distort.

Distortion qualifies as desirable or not, depending on context. It cleaves naturally into that occurring in the amp, during normal operation or abuse, and that produced by outboard engines.

Distortion In The Amp

Amps hold the seeds of so much distortion that some players add no outboard fuzz. Every amp distorts, but not all the time and not always audibly. Distortion due to *headroom exhaustion* flows from the fact that no stage can deliver more than the available voltage. Solid state tends to go straight from linear transfer to clipping. Decapitation of waveforms at the tips gives an edge some players find desirable. Pushed further, the process replicates sine-to-square conversion (Fig. A110). By contrast, tube stages squash the signal long before clipping commences. Headroom exhaustion forms an essential part of tube sound, discussed in its own Appendix.

Crossover distortion (Fig. A112) occurs in push-pull amplifier stages. By definition, half the stage handles the positive voltage swing, half handles the negative swing. No distortion results if the stages hand off perfectly. If the hand-off runs awry, a failure of linearity called a deadband is created; the resultant crossover distortion has as diagnostic a look in solid state as in tubes, and is easily generated at levels from subliminal to overpowering. Crossover distortion has been cited as one facet of "Marshall sound."

Transformer-related distortion is confined to the tube scene. The transformer's primary winding ties to the output tube(s); its secondary winding to the speaker(s). The resultant system contains heavily reactive elements, often complicated by a negative feedback path, introducing distortion terms even below the level of transformer saturation. Saturation injects gross harmonic distortion. The saturation point hinges on power and frequency; lows usually saturate before highs.

Some gain stages act as mixers, in the Ham-radio sense of the word. Studio parlance uses mixing to

Basic Stomp-Box Distortion Modes

Sine-to-Square Conversion
    methods: comparator, Schmitt trigger
    sound: archetypal harsh
Headroom Exhaustion (clipping)
    methods: drive a stage to its voltage limit
    sound: depends on degree of overdrive
    and available headroom
AC Logarithmic Amplification (a/k/a
    diode clipping)
    methods: the name is the method
    sound: not as harsh as true SSC; changes
    with mods to log amp
Fullwave & Serial Fullwave Rectification
    method: precision fullwave rectifier
    sound: mixture of smooth and harsh;
    better dynamic tracking than
    squarewave and diode-clipped feeds
Squashing
    methods: linearly biased CMOS inverter,
    overdriven OTA; triode stage
    sound: ranges from 'warmth' to
    diode-like clipping
Crossover Distortion
    methods: feed signal through improperly
    biased push-pull stage; many diode
    circuits
    sound: scant effect on individual notes
    unless severe; chords acquire a chunky
    quality
Intermodulation Distortion
    method: multiply axe feed against itself in
    a 4QM
    sound: variable; usually accompanied by
    harmonic distortion
Harmonic Distortion
    methods: most distortion methods
    generate frequencies that are even or
    odd multiples of the primary tone; in that
    sense, they all make harmonic
    distortion
    sound: depends on severity, and whether
    even or odd, and high- or low-order
    harmonics predominate

designate an algebraic sum. Mix-born distortion comes from four-quadrant multiplication. In this sense, mixing two different frequencies yields an output from which the input frequencies have been nulled, plus two new terms, the sum of the input frequencies, and their difference. Amp stages don't act as pure mixers, but can mix a percentage of the signal. Two types of distortion flow from this process. *Intermodulation distortion* describes the result of mixing two different frequencies. The output of an amp fed a 1-KHz tone and a 620-Hz tone contains those two tones, but also portions of 1620 Hz and 380 Hz (Fig.
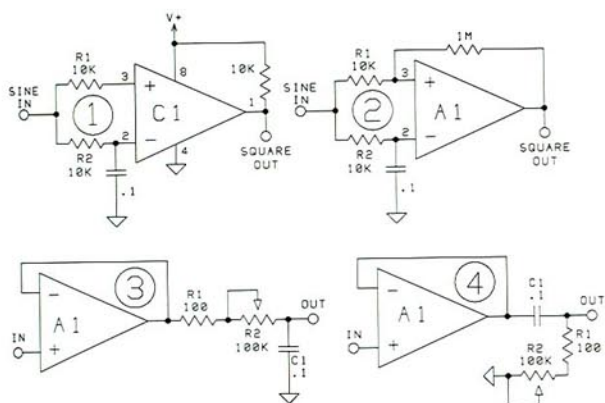
Fig. A103. 1—Comparator, such as LM393, configured as sine-to-square converter. 2—Op amp configured as comparator; positive feedback avoids linear response. 3—Basic integrator; effect on squarewave illustrated in Fig A107–C, D. 4—Basic differentiator; effect on squarewave illustrated in Fig. A107–E, F.
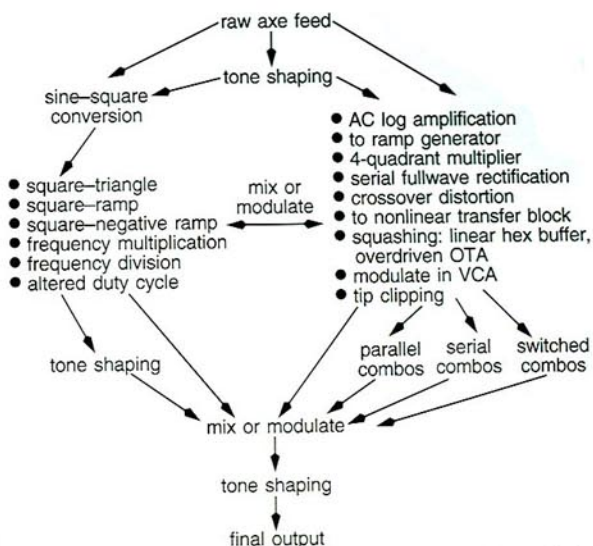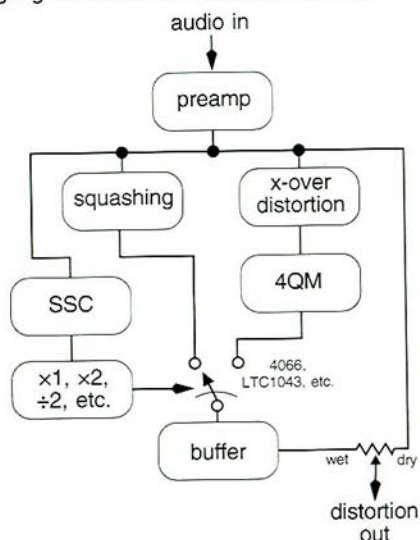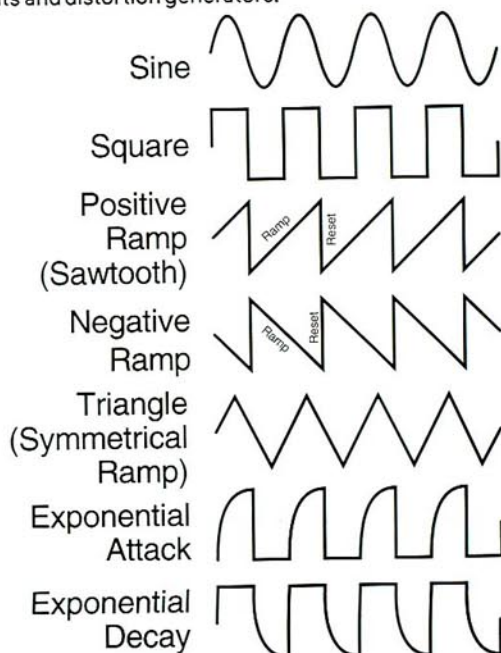

Fig. A105. Basic distortion tree. A more detailed rendering becomes unmanageable due to interconnects and overlaps. Anything is possible; no rules apply.

A113). The percentage of new tones tells the intermodulation distortion. IM's key aspect is that the new tones don't represent integer multiples of the original frequencies. Mixing a single frequency with itself gives a difference term of zero. The sum term equals twice the input, making the mixer a doubler of isolated sinewaves. The creation of integer multiples of the fundamental tone is known as *harmonic distortion*, and more typically results from clipping and squashing.

The amp output is not the end of the line. One final transform happens in the speaker, a complex mingling of drive level, damping, impedance, and cone materi-

al. Speaker nonlinearity accounts for distortion whose role amp makers now appreciate. The fact that the speaker cannot accurately reproduce what the amp feeds it probably holds part of the magic.

Outboard Distortion: Fuzz Boxes

Compared to distortion in the amp, fuzz is deliberate, deviant, blatant; typically dominant. Sounds in nature trace to vibration, a sinusoidal phenomenon. Human hearing discerns a sinewave as right and pure; altered sinewaves as distorted. So, distortion generators corrupt sinewaves. Though the number of possible per-

Fig. A104. An example of multiplexed distortion. Preamp feeds parallel distortion paths, each producing a different sound; also feeds sine–square converter optionally subjected to frequency division or multiplication, which controls toggling rate between the distortion paths.

Fig. A106. Typical waveforms encountered in electronic circuits and distortion generators.
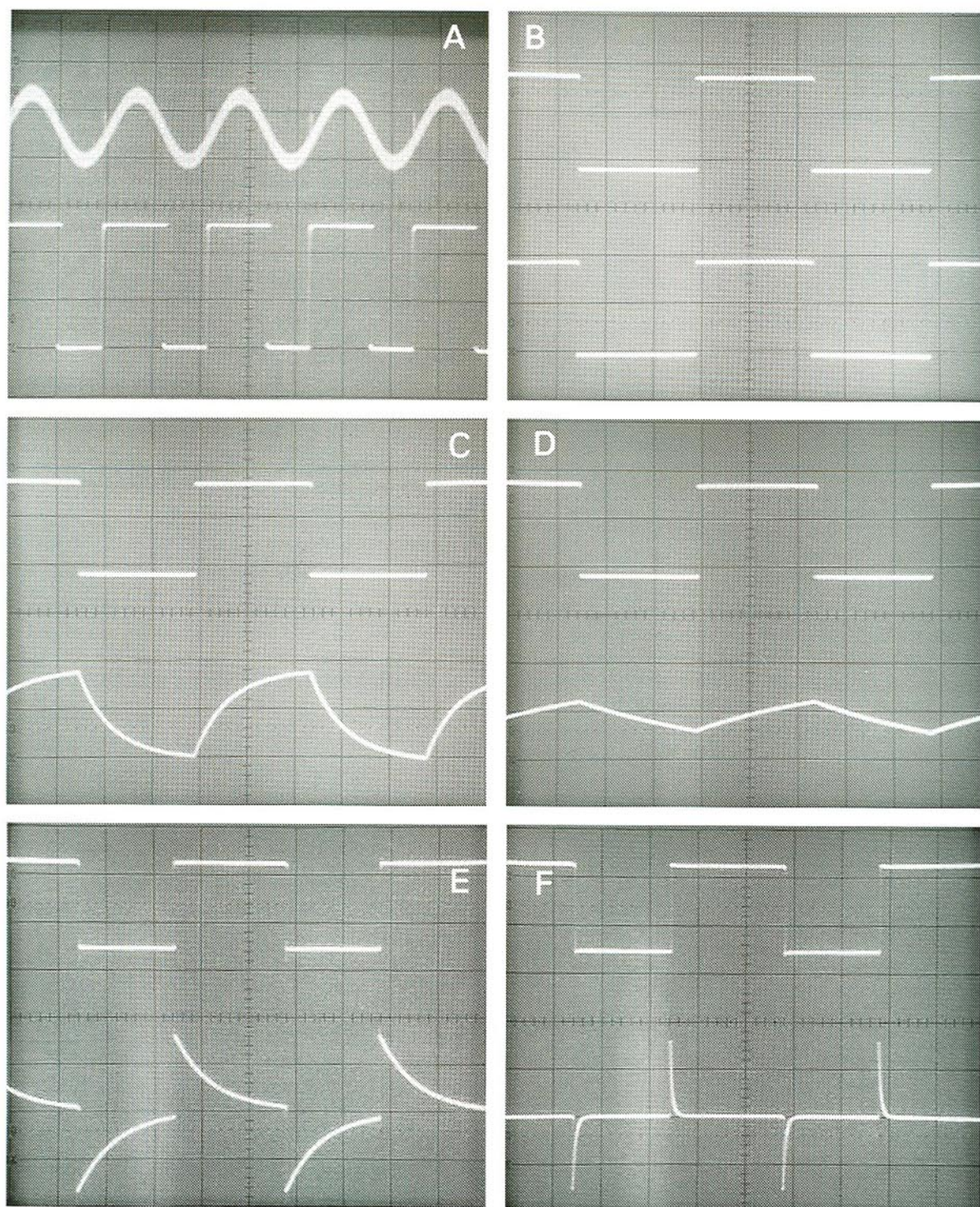

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Fig. A107. A—I/O of Fig. A103–1. B—I/O of Fig. A103–3 & 4, when integration and differentiation are canceled by setting of R2; input essentially equals output. C—I/O of Fig. A103–3 (integrator); increasing value of R2 reveals exponential curves typical of a capacitor charging and discharging through a resistor. D—R2 at maximum; output approximates a triangle wave; also note significant drop in amplitude. A given amount of integration affects different frequencies differently, something not apparent from these photos. E & F—I/O of Fig. A103–4, a basic differentiator. As value of R2 is progressively lowered; network selects progressively higher frequencies. Integration and differentiation affect the amplitude of sinewaves, but alter the shape of non-sinewaves.

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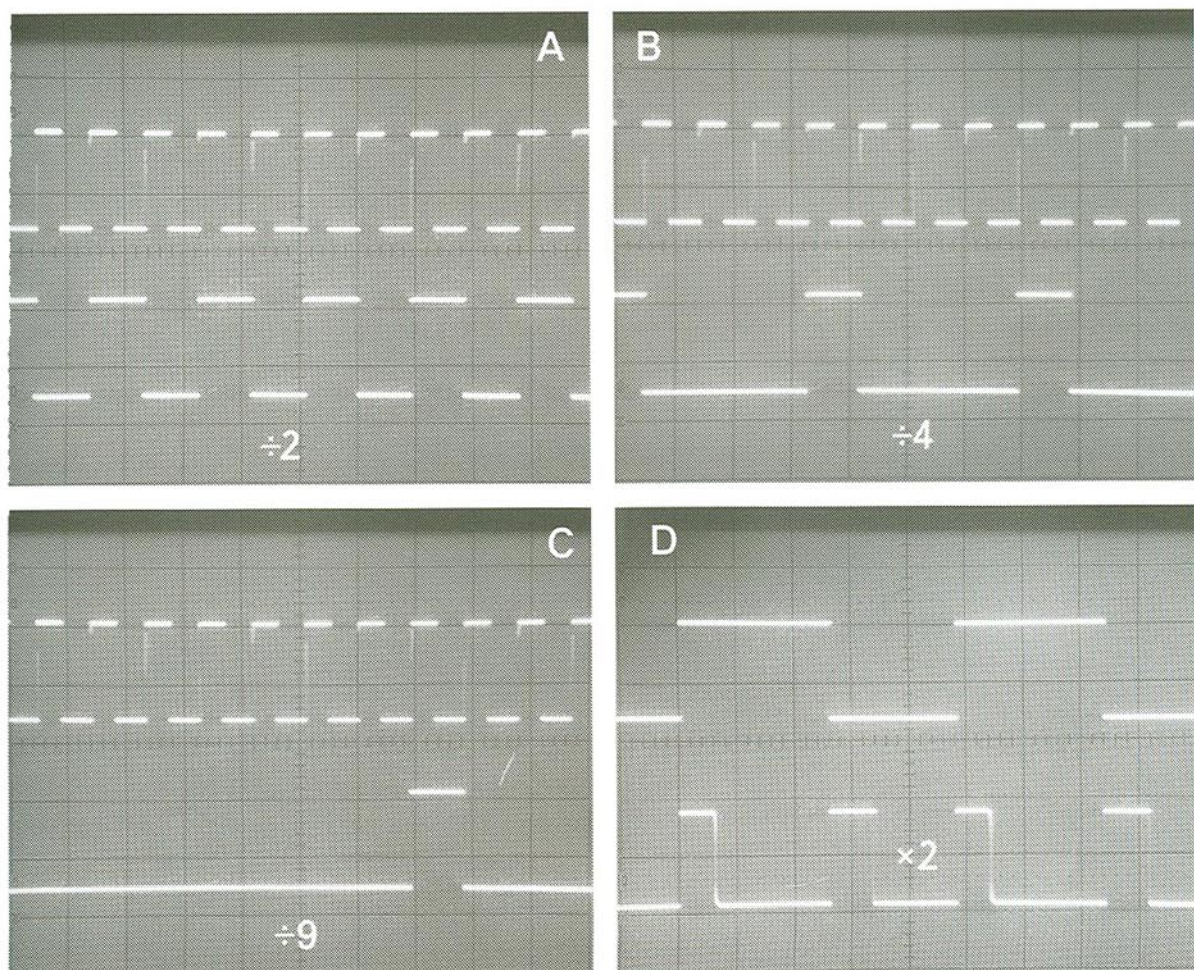
Fig. A108. 1—Dual flip-flop wired as two independent divide-by-2 blocks; if wired in series, results in divide-by-4. The 4013 can be particular about inputs and/or supply voltage; most consistent results result from passage of signal through at least one CMOS buffer before it reaches the 4013. 2—Squarewave frequency doubler; exclusive OR gate changes state on leading and falling edges of input; duration of resultant pulses depends on time constant of RC network; trim to suit input frequency range. Output may need squaring up in one or more buffers. 3—4017 configured to divide-by-n, in this case, n is any factor between 2 and 9. Photos above illustrate selected I/O of these circuits.
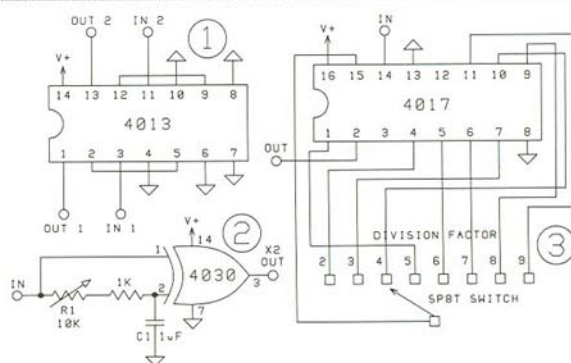
versions is large, a few predominate out of ease of creation and control.

Sine-to-Square Conversion (SSC)

SSC makes a natural first step in a buzz box because squarewaves sound harsh, and lend themselves to further, readily realized changes. True SSC uses circuits having only two output states, LOW and HIGH. Examples include zero-crossing detectors and Schmitt triggers. The output has perfectly sharp shoulders, at least in a musical time frame (Fig. A107–A).

Once a squarewave is obtained, the leading and/or trailing edges are easily altered by the same attack and decay networks used in level detectors. The main difference is that the RC time constants measure microseconds instead of milliseconds.

Phase-shift prior to SSC changes the point at which zero crossing occurs, shifting phase between wet and dry, resulting in comb-filtering if wet is then mixed with dry.

Squarewaves lend themselves to altered rate. Musically apt changes double or quadruple, halve or quarter the frequency. Division by two is achieved by a
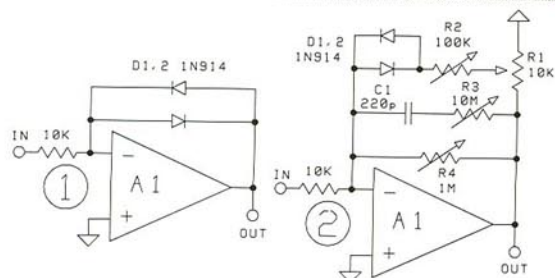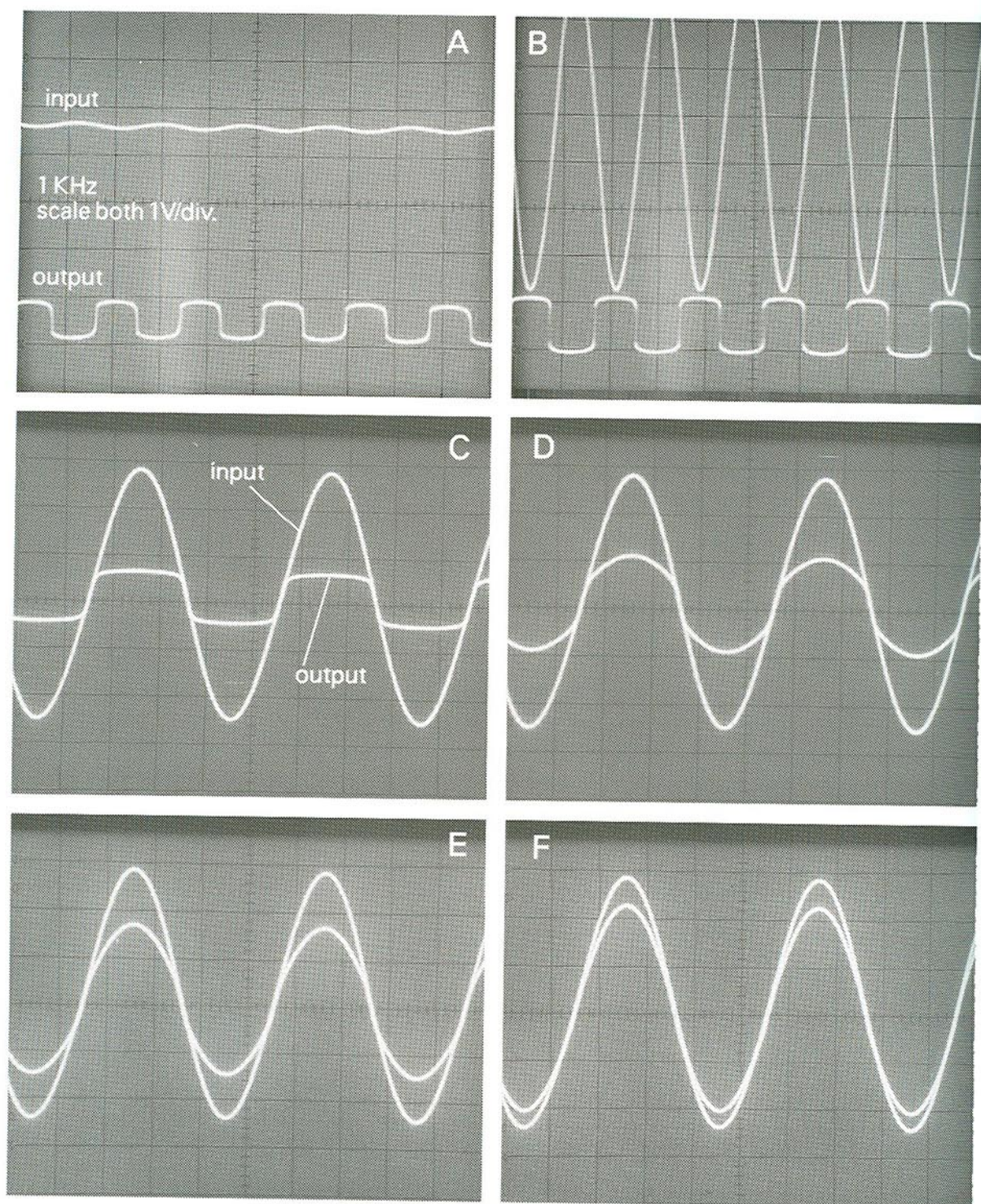
Fig. A109. Photos A & B show I/O of schematic 1. A—Low-level input (top trace), diode clipper output (bottom trace). B—Input increases to several volts$_{p-p}$, but output has increased minimally, due to log amplification. Rounded shoulders of diode-clipped output give a softer sound than sharp shoulders of a true squarewave. Photos C through F depict I/O of schematic 2, in which setting of R1 is varied, affecting contour of waveform after diode turns on. Circuit allows manipulation of other variables affecting sound, e.g., high frequencies roll off when R3 is decreased from its 10M value; R4 varies sensitivity of circuit by limiting maximum gain available in the feedback loop.
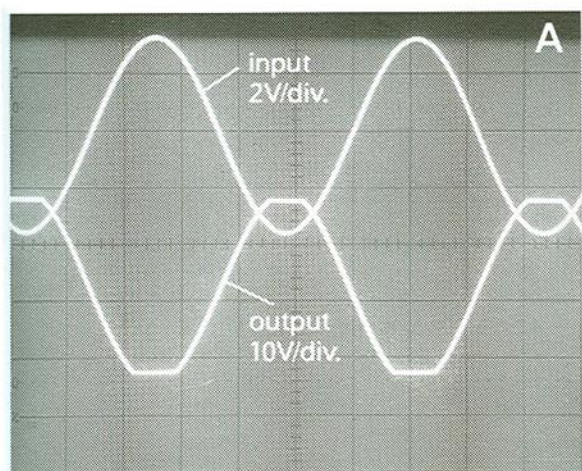
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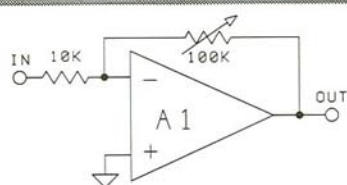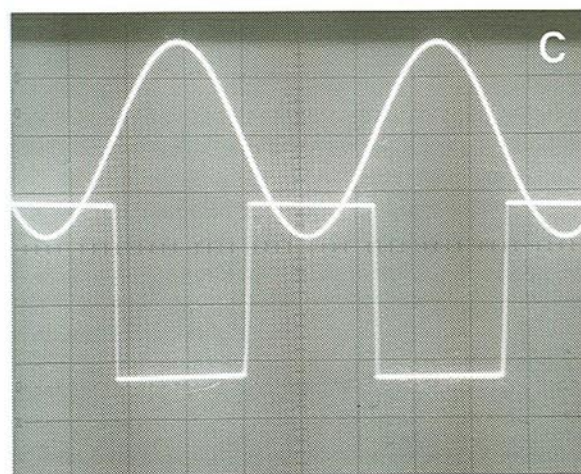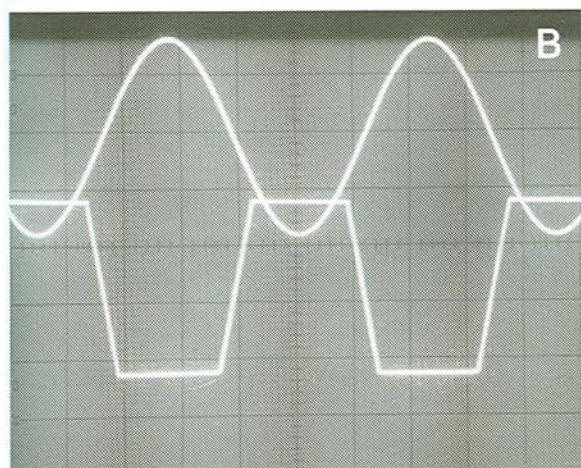
Fig. A110. I/O of 1 KHz sinewave feeding TL071 configured as inverting amp whose gain varies 0–10 depending on setting of 100K pot. A—Barely clipping. B—Gain increases; clipping becomes more pronounced. C—In effect, sine-to-square conversion.

flip-flop; division by four, by two flip-flops in series. A 4017 counter divides by a selectable factor from two to nine (Fig. A108). Frequency division finds use outside distortion in a class of boxes known as dividers. Frequency doubling is achieved many ways, one of the simplest being an exclusive-OR gate (Fig. A108–D).

Squarewave sound changes with *duty cycle*, the percentage of time spent high relative to the time spent low. Altering the duty cycle of an existing squarewave usually requires differentiation, to convert it to a pulse train; each pulse then triggers a one-shot with a variable duty cycle.

Integrators and differentiators—lowpass and high-pass filters, respectively—alter the amplitude of sinewaves, but change the shape of squarewaves. Integration removes high-frequency products; differentiation removes low-frequency components (Fig. A107–C to F)

Diode Clipping

Perhaps the most versatile single distortion step, diode clipping isn't clipping at all, but AC logarithmic amplification. Two diodes wired parallel/reversed in the feedback loop of an op amp give a particular gain curve. The amp attempts to apply its open-loop gain below the diodes' conduction threshold. This is one reason some fuzz effects are noisy. Once the signal overcomes the diodes' forward drop, output voltage becomes proportional to the log of the current through the diodes. The wave plateaus at about twice the diodes' forward drop; $1.2V_{p-p}$ for silicon diodes, $0.7V_{p-p}$ for germanium diodes. As expected from its appearance (Fig. A109), diode clipping sounds softer than SSC.

Diode clipping allows manipulation of several variables during creation. Reducing the op amp's low-level gain with parallel resistance keeps the fuzz but reduces sensitivity and noise. Raising post-turn-on gain with series resistance improves dynamic tracking without losing distortion altogether. The diode's forward conduction point can be altered with a pot tied to the output; a capacitor in the feedback loop skews frequency response. Sound also changes with diodes having higher or lower forward drops than 1N914s. Despite their sloping shoulders, diode-clipped waves qualify for many of the transforms applicable to squarewaves.

Diode clipping and SSC are self-limiting because their products have fixed amplitudes. Indeed, both processes are meant to ignore input amplitude. Restoration of dynamic tracking demands explicit steps.

Perverting Sinewaves

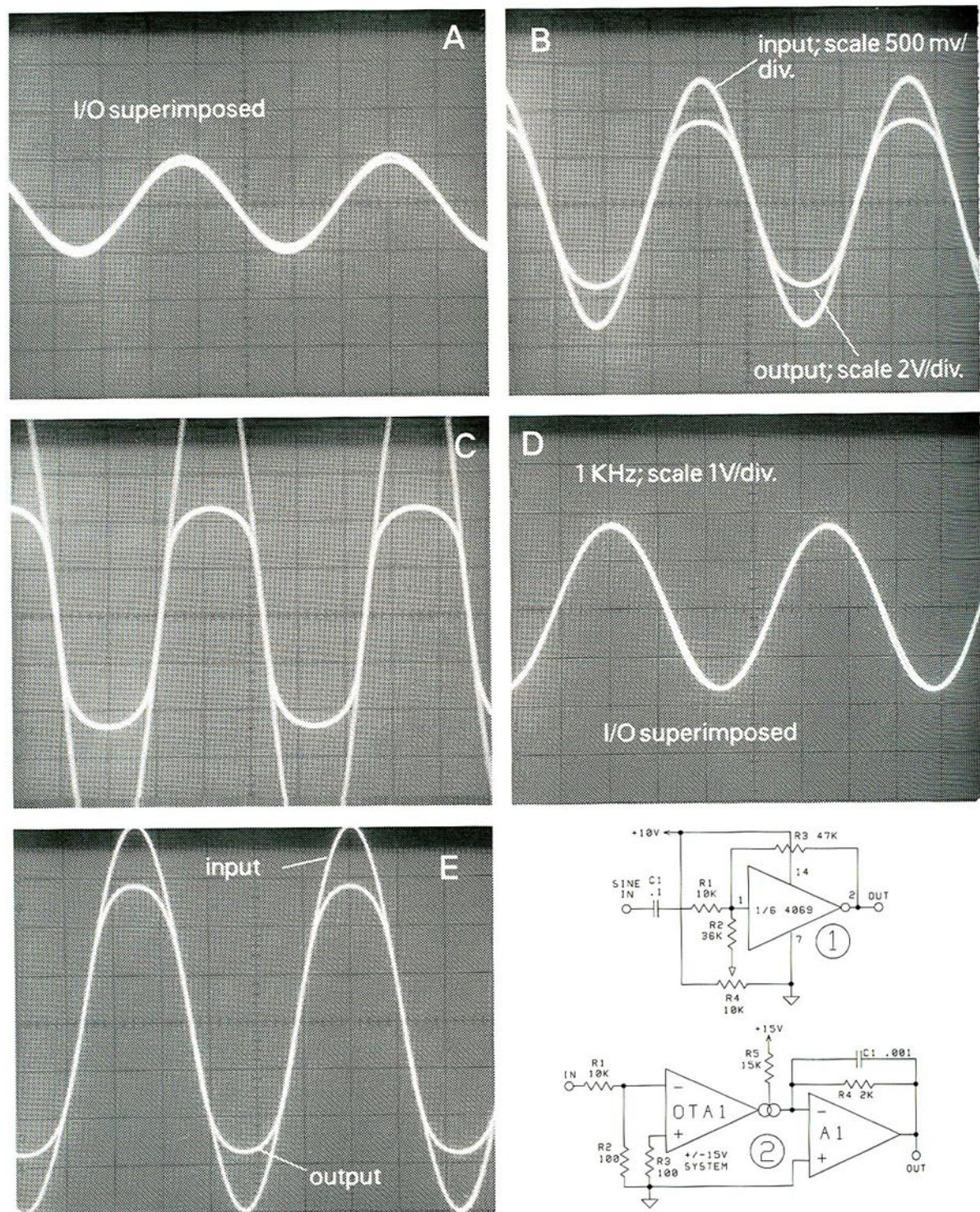Attacking a sinewave without prior squaring yields a varied class of sounds that tend to retain dynamic

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Fig. A111 (photos above, and on facing page). Photos A/B/C depict I/O of schematic #1. Waveforms are superimposed to highlight the amplitude-dependent differences; output has been inverted to facilitate comparison; bias has been trimmed for best symmetry. A—Transfer is essentially linear up to an input of ~1V$_{p-p}$. B & C—As input level increases further, output becomes progressively squashed. Photos D & E depict I/O of schematic #2; OTA exhibits linear transfer up to about 3V$_{p-p}$; squashing grows after that point, ending finally in a function that resembles diode clipping. Distortion threshold can be raised or lowered by changing value of 10K input resistor; this also changes gain, so I–V conversion resistance will have to be altered to keep unity gain. F—I/O of OTA driven by triangle wave; input & output superimposed; linear transfer at low levels. G & H—Triangle rounds off as level increases. I—Applying a DC offset to the OTA results in asymmetrical squashing.
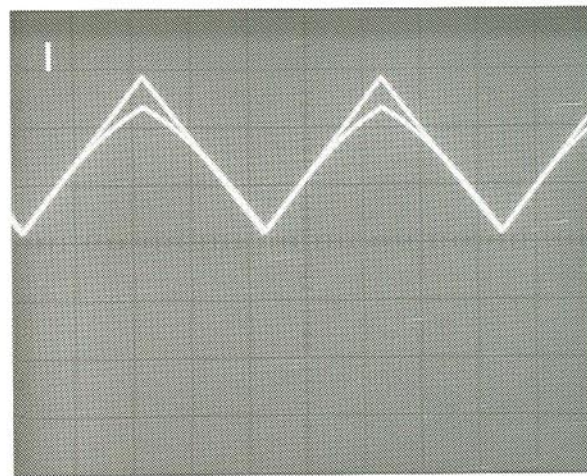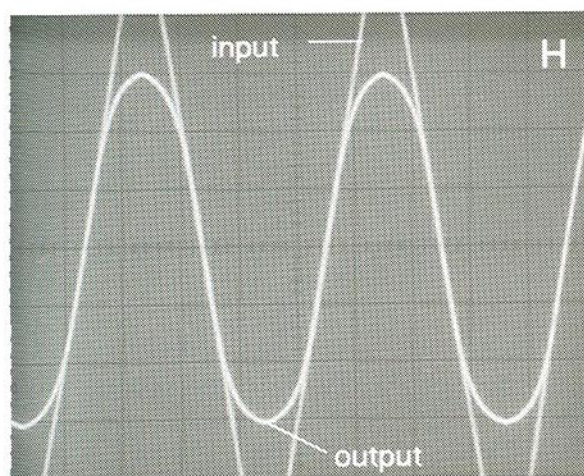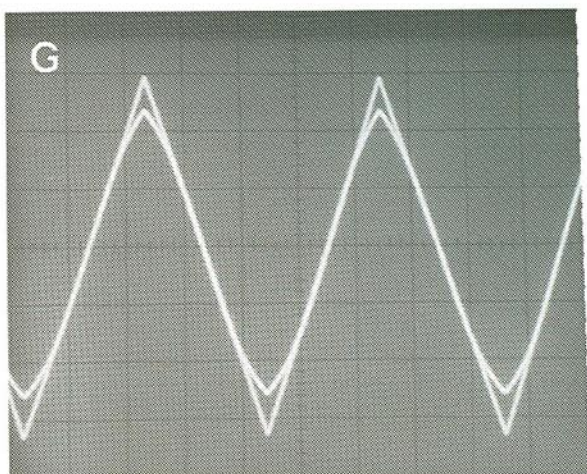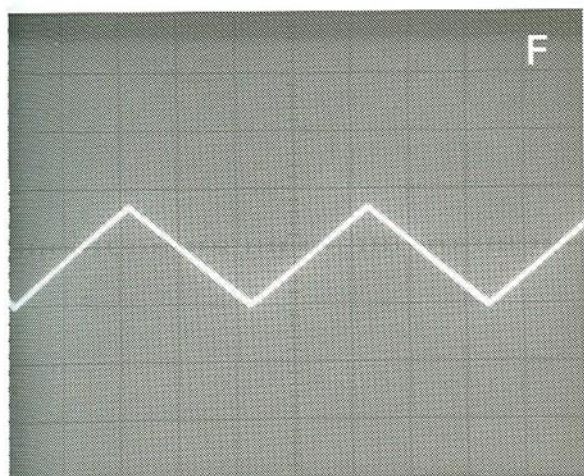
tracking. Fullwave rectification generates 2nd harmonics, along with high-order 'fuzz' products that sound smoother than squarewaves. One variant of rectification unbalances the rectifier, so that alternating pulses have different amplitudes. Serial fullwave rectification doubles frequency with each step.

Driving a ramp generator with a sinewave creates an output whose shape changes with input level; a very high level generates a ramp, a lower level creates a variable nonlinear output.

Sinewaves were meant for crossover distortion, which needs a sloping waveform to express itself.

Multiplying the signal against itself in a 4QM has varied effects. A pure sinewave emerges doubled in frequency; complex sinewaves typical of an axe feed acquire harmonic and intermodulation distortion, the latter giving rise to ring-modulator–like sounds.

Squashing describes one aspect of the tube-amp transfer function (Fig. A111). Sinewave tips take on a rounded contour that intensifies with level. Solid state replicates this aspect in linearly biased inverting CMOS buffers, such as a 4069. The output contour is rounder than that of a diode clipper, and usually asymmetrical unless trimmed. A similar effect results from overloading the input of a transconductance amp. The higher the level, the flatter the wave becomes until it resembles diode clipping.

All these transforms are fair game for nonlinear transfer functions.

Miscellaneous Distortions

A distortion pedal can use any circuit whose output represents a perversion of the input. Some of the tastiest modes flow from abusing systems meant to resist distortion. For instance, FM systems are designed to convey signals with immunity to interference; yet modulator and demodulator tolerate only so much deviation. Past a certain point, they lapse into distortion whose character shifts with drive level. A phase locked loop sounds particularly distressed when losing lock. Choice of a low-frequency carrier aliases the system, producing audio-band spuriae.

Smoothing ripple from the control feed forms an essential part of designing a compressor, because control-voltage ripple distorts audio. Perverted control feeds make distortion tools out of compressors. VCAs are two-quadrant, rather than four-quadrant multipliers. The control feed alters the musical envelope, but
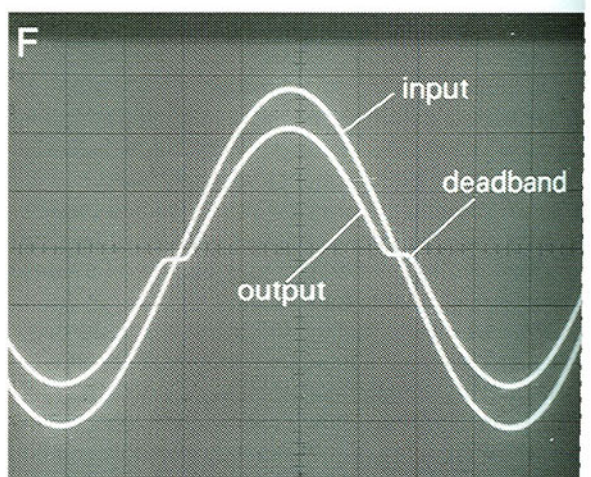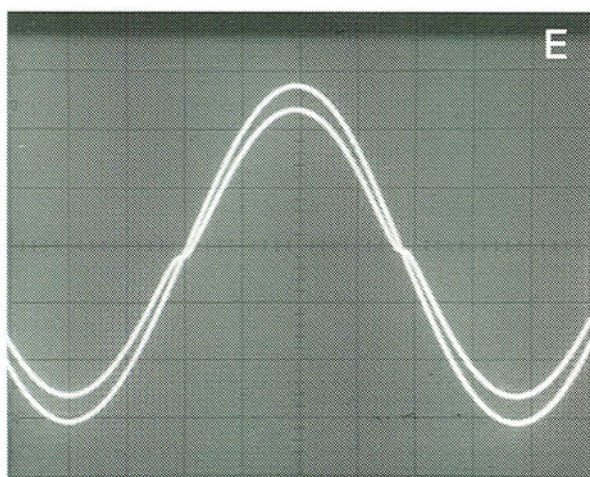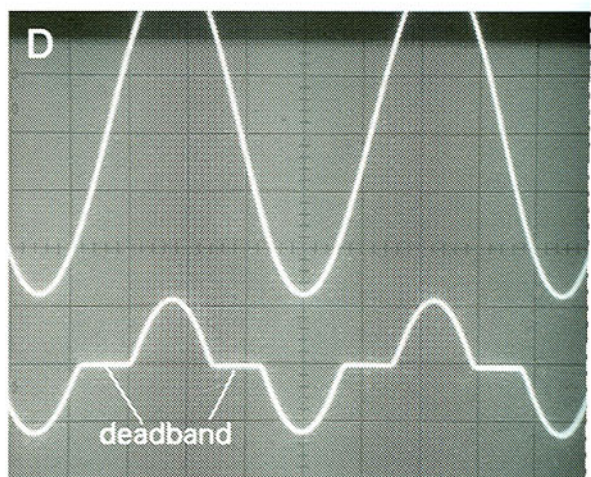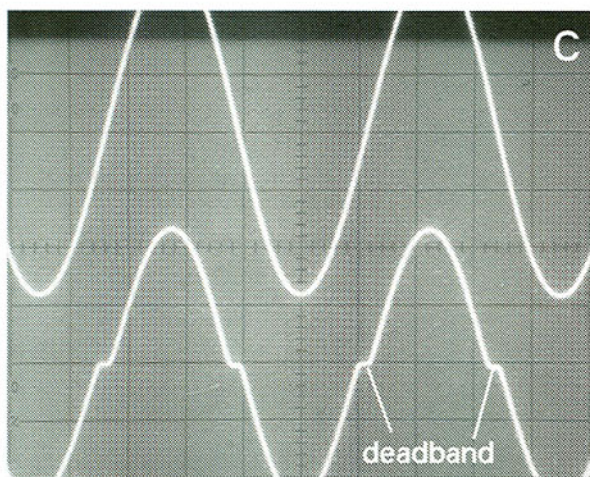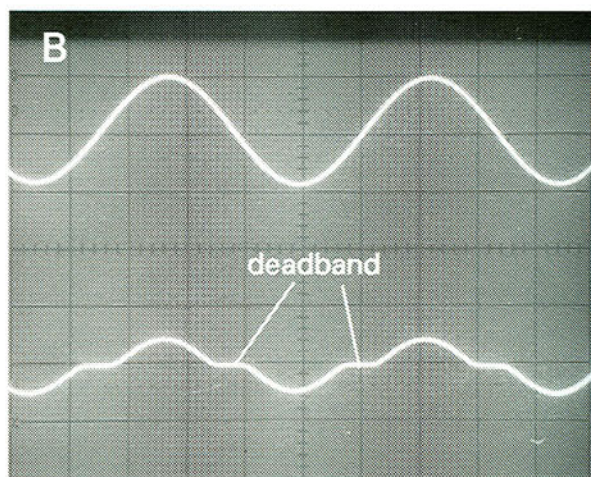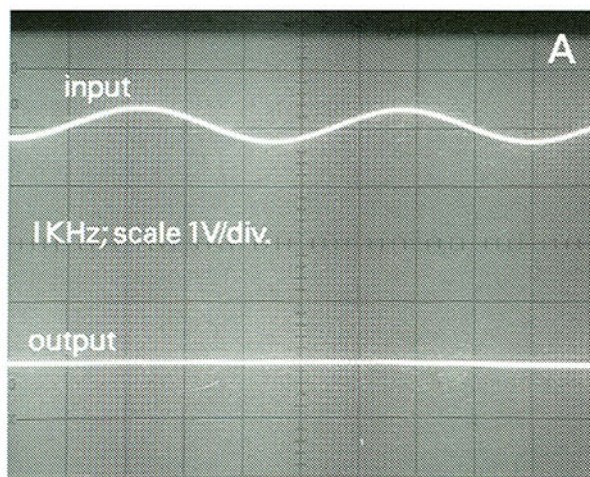
Fig. A112. Photos A to D depict I/O of schematic #1. A—Input fails to reach 1N914 diodes' conduction threshold, so no signal gets through. B—Signal has overcome diodes' conduction threshold; marked deadband is present. C—Signal increases, deadband becomes proportionately less. D—Identical to photo C, but green LEDs have been substituted for 1N914s; greater forward drop accentuates deadband. E & F— I/O of schematic #2. Variable bias applied through R1 alters relative bias of push-pull pair, variable deadband distortion. Many other ways exist to realize crossover distortion in transistor and tube circuits.
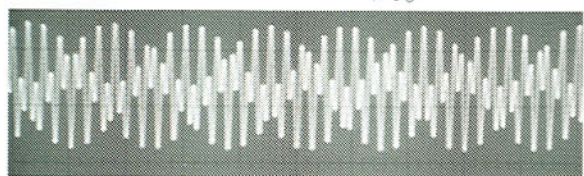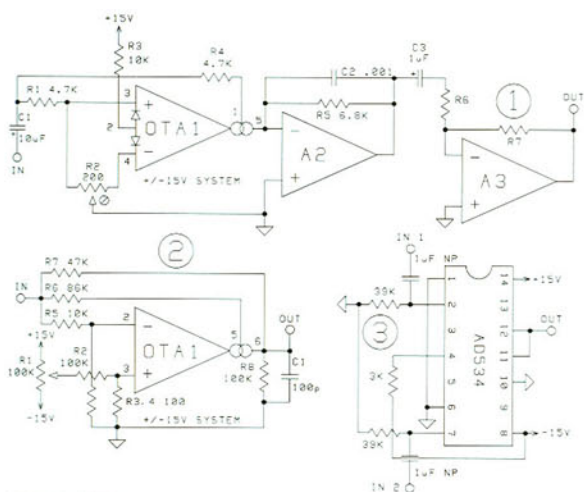
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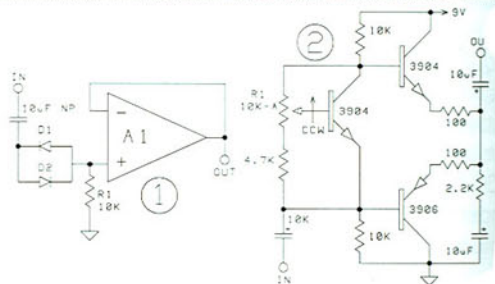
Fig. A113. Generating harmonic and intermodulation distortion using four-quadrant multipliers. 1—Half an LM13600 configured as 4QM. Builder adapting this circuit to other uses should note two strong DC offsets. Input exists below ground, I–V converter A2 output exists well above ground. Output amplitude is low, requiring scaling amp A3 to give desired amplitude. Trim 200-ohm pot to balance multiplier. 2—Functionally similar circuit based on LM3080; trim R1 to balance multiplier. Input exists below ground; output varies above or below ground, depending on trim setting. 3—Simpler, but much more expensive approach uses AD534 precision analog multiplier. 3K resistor sets internal scaling amp such that output is about 38% of input, fairly high for a 4QM. Circuit shown allows multiplication of two different inputs. Multiplying a signal against itself affects amplitude as well as frequency. This causes multiplier to act as an expander that quickly reaches clipping if overdriven. Photo shows output of the 534 circuit fed two different sinewaves.

does not produce the sum and difference terms of four-quadrant multiplication.

Multiplexed distortion involves electronic toggling between different distortion paths (Fig. A104). The switch can be driven rhythmically, randomly, or by a pulse train derived from the input. Alternating between distortions give a sound distinct from summing them. Discontinuities due to switching add a third component to the mix.

Nothing is too bizarre when it comes to fuzz, e.g., feeding audio through a silicon controlled rectifier, or an unbypassed 3V regulator.

Tone Shaping

Tone shaping changes the character of distortion in two ways. First, many distortion transforms exhibit thresholds. Altering the frequency profile ahead of the distortion block changes the population of frequencies that reach threshold and undergo transformation, and those that pass unmolested.

The distorted signal contains new frequencies above the fundamental, or below it in the case of frequency division. EQ accentuates or mutes these new tones.

Cooking Up Distortion Boxes

Distortion seeks less a practical end than the genesis of new sound, preferably christened in the font of cool. That attitude bred the adage that no rules rule distortion. Rules or not, some approaches enjoy greater success than others. The builder new to buzz boxes might find these guidelines helpful:

► Avoid redundancies. Diode clipping followed by SSC serves little point.
► Leave room for the effect to express itself. Some distortions blunt the impact of others. SSC ahead of crossover distortion defeats the crossover distortion because the deadband needs sloping sides to manifest.
► Keep two equalizers handy, one to apply before distortion, one after. EQ sometimes vivifies transforms that sound lackluster raw.
► Few great-sounding distortion boxes involve a single transform. The more fruitful approach combines multiple transforms (serial and/or parallel) with tone shaping.
► Consider the worth of a tape library correlating instruments and playing styles subjected to specific types of distortion. Each recording should follow a routine that covers the fretboard, and includes single notes and chords, plus a range of dynamics. A box that does nothing for individual notes might be found to imbue chords with body. Players who use several axes with significantly different output or tone should include a segment with each axe. A tape library can remind the builder of paths that lead to dead ends, and help correlate waveforms with sound.
► Give wacky approaches a try. DM6 arose that way. Distortion is one area where the builder can cook up sounds immune to ripoff.

Sonic Correlates of Distorted Waveforms

This Appendix treats distortion as another ingredient with which to cook up new sound. The builder need not understand distortion circuits to use them; but understanding why transforms give certain sounds helps guide the builder seeking specific sounds. A harmonic profile aids the analysis. All non-intermodulation distortion can be characterized as a fundamental tone plus a blend of harmonics. The fundamental is always a sinewave; each harmonic is an integer-multiple of the fundamental. Harmonics are further de-
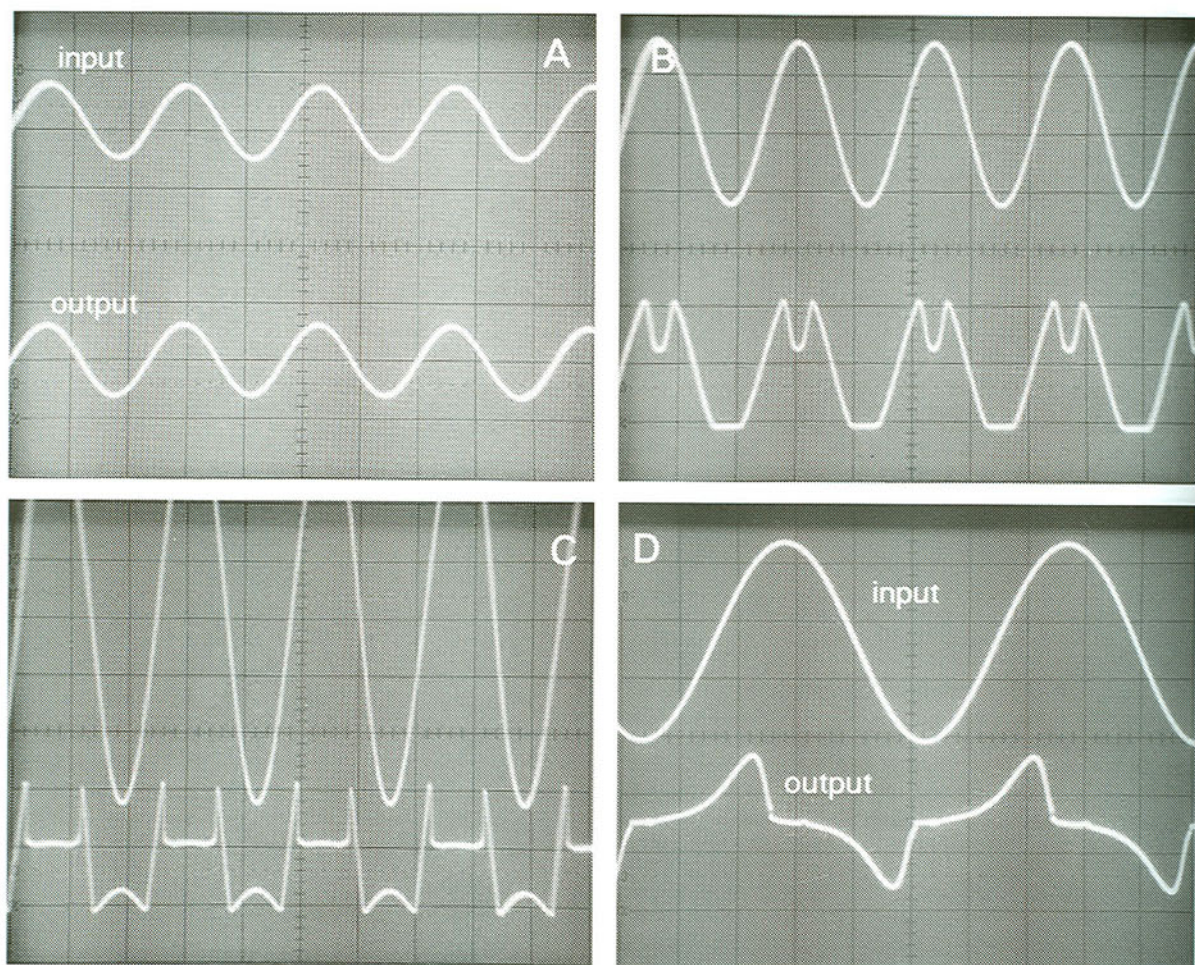
Fig. A114. A–C show I/O of FET-based nonlinear function block, shown in section on soft-knee compression (Fig. 88–5). Output has been inverted to facilitate comparison. A—Linear transfer below threshold. B—FETs turn ON at points on the positive and negative swing varied by potentiometer settings. C—Distorted waveform has nothing to do with a trademarked superhero. D—I/O of 42TU016 transformer; 20 Hz, scale 5V/div.; illustrates distortion due to transformer core saturation.
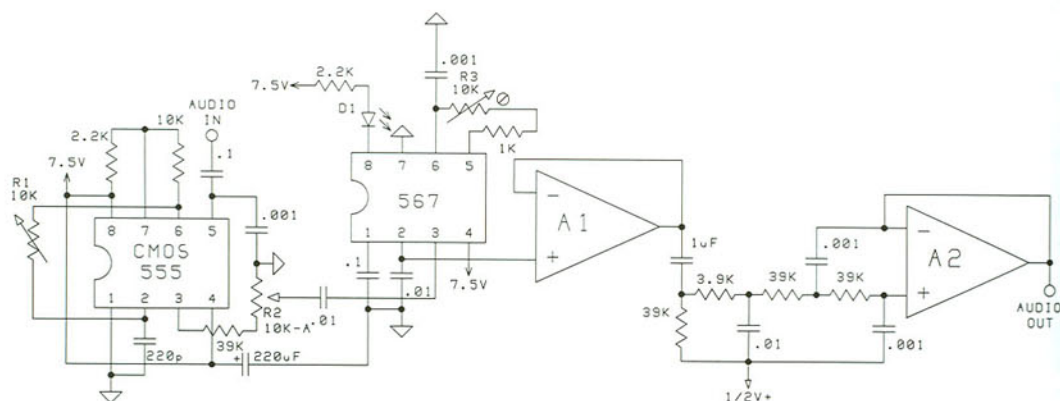

Fig. A115. Schematic shows a bare-bones FM system. LM555 timer chip acts a frequency modulator; carrier frequency ~200 KHz, trimmed by R1. Output feeds R2, which varies FM signal level feeding phase locked loop demodulator NE567; R3 trims PLL frequency. To test this setup, center R1 & R2, trim R3 so that LED lights, indicating that the PLL has locked. Connect circuit to axe and amp (start with amp volume LOW). Overdriving the 555 occurs at audio levels above ~100 mv$_{p-p}$ and yields raucous but rich distortion. The sound of the PLL losing lock can be heard by turning R2 CCW; be aware that sudden and large amplitude shifts that may attend this move. Sound also varies when carrier is taken off center by changing the setting of R1. DM6 operates on these principles, but uses a 40-KHz carrier, and the slope detector formed by co-resonant transducers in place of a PLL. Layout is critical to keep carrier from coupling to 567 input by parasitic paths. Some designs may require running 555 and 567 off independent 9V batteries.
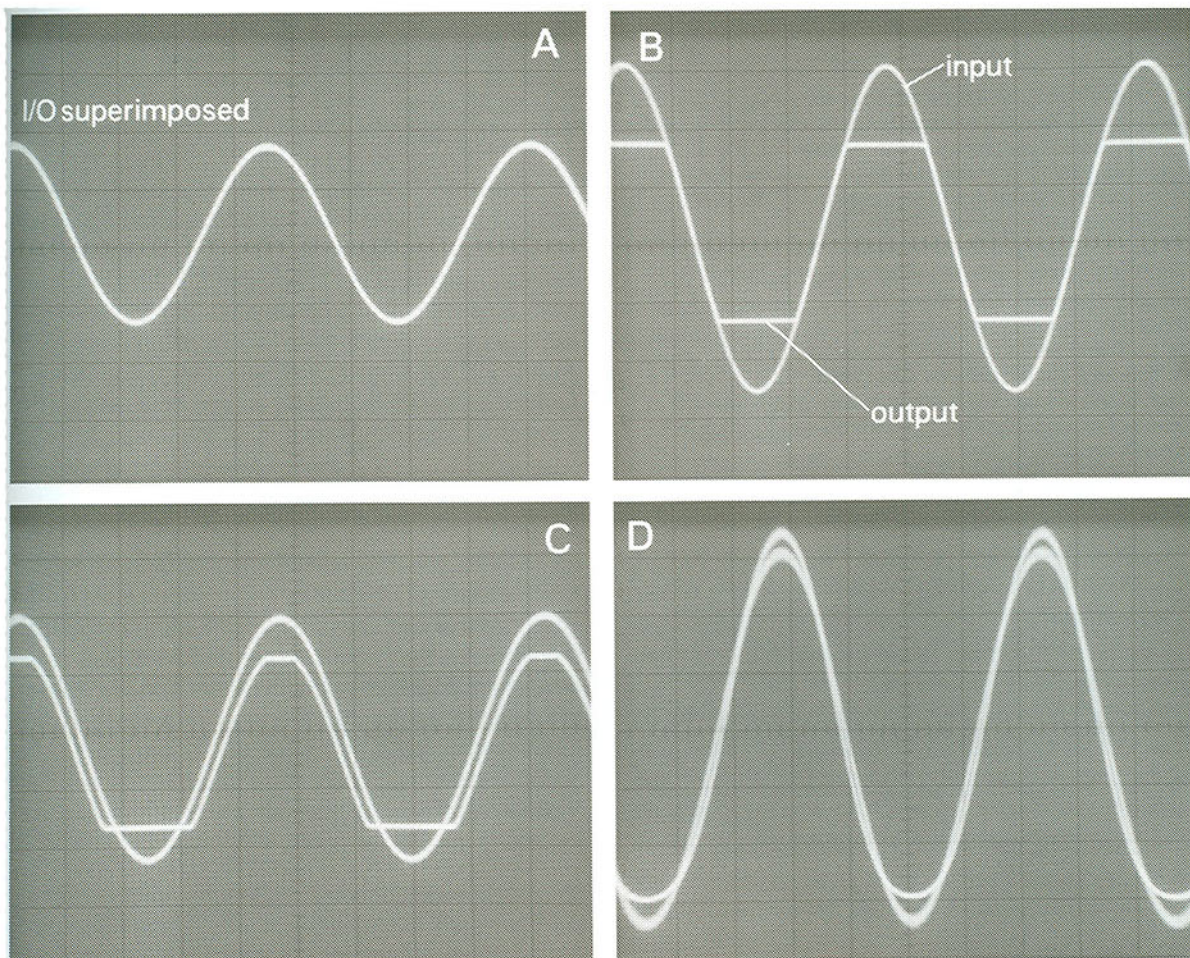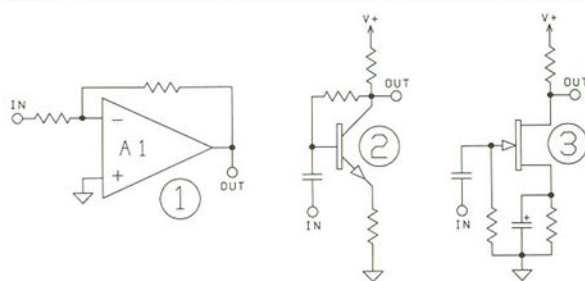
Fig. A116. Photos illustrate how different solid state stages behave at their headroom limit. A—Output inverted, superimposed on input; all three circuits are linear if kept within their headroom. B—Op amp goes straight from linear transfer to symmetrical clipping. C—Bipolar transistor also clips, but asymmetrically. D—FET squashes before clipping, but margin between squashing and clipping is slim compared to vacuum tube. All photos 1 KHz sinewave input; scope gain trimmed to match input and output amplitudes below clipping threshold.

scribed as even (second, fourth, etc.) or odd (third, fifth, etc.); and in terms of their proximity to the fundamental. Low-order harmonics are close to the fundamental in frequency; high-order harmonics are further away. For example, the second and fourth are low-order even harmonics; the eighth and the sixteenth are high-order even harmonics. The third and fifth are low-order odd harmonics; the ninth and the eleventh are high-order odd harmonics. Harmonic analysis of a squarewave confirmed Fourier's claim that it equaled the sum of an infinite series of sinewaves, the fundamental plus $1/3$ the third harmonic plus $1/5$ the fifth, and so on ad infinitum. Harmonic analyses are performed by computer, using Fourier's equations; or through spectrum analysis of the waveform under study. An audio spectrum analyzer is an extremely narrow bandpass filter swept from 20 Hz to 20 KHz. Each harmonic's amplitude is plotted relative to that of the fundamental (Fig. A116B).

Harmonic profiles have many uses. In a sense, they constitute sonic fingerprints. Naturally occurring harmonics account for timbre, which makes C from a trumpet unlike C from piano or guitar. Generally, even harmonics support the fundamental, lending the sound a musical or choral quality; odd harmonics oppose the fundamental, producing a woody, muted tone. A predominance of low-order harmonics tends to sound smooth or warm. Harmonics above the fifth sound harsh, especially if clustered between 1 KHz and 5 KHz, where hearing exhibits peak sensitivity.

Fig. A116B. Two illustrations of harmonic profile. A squarewave, as Fourier showed, represents the sum of an infinite series of odd harmonics related by amplitude: the fundamental plus a third the 3rd harmonic plus one-fifth the 5th, and so on. A sawtooth or ramp represents the sum of an infinite series of even and odd harmonics, also related by amplitude: the fundamental plus half the 2nd harmonic plus a third the 3rd, and so on. This type of harmonic vs. amplitude graph has predictive value in assessing distortion; musical instruments exhibit typical harmonic profiles.
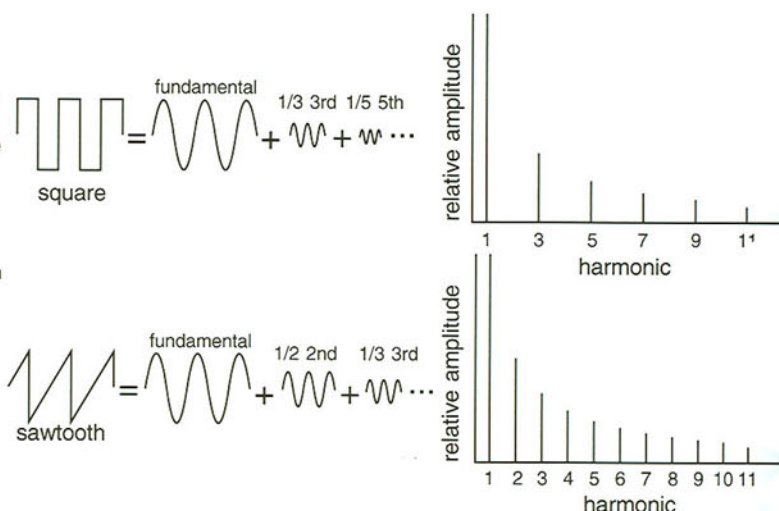
Harmonics affect perceived loudness. Orders above the 4th bear an impression of loudness known as the edge effect. For example, spectrum analysis of the sound of a trumpet reveals a rise in the 5th harmonic when the trumpet blares.

Several common transforms alter sinewaves by limiting their excursion. The resultant harmonic profile depends on symmetry and abruptness. *Symmetry* denotes the extent to which tip-altering modes affect the positive and negative swing equally. Symmetrical modes generate odd harmonics; unipolar modes make even harmonics (Fig. A124). *Abruptness*—the sharpness of the cramped edge—determines harmonic order. Soft edges generate low orders (2nd–4th), sharp edges create high orders (5th and above). In a squarewave, harmonic energy exceeds that of the fundamental, with much of the energy clustered above the 5th harmonic. In a squashed wave, harmonics roll off rapidly, less than 2% of total energy falling above the 5th harmonic.
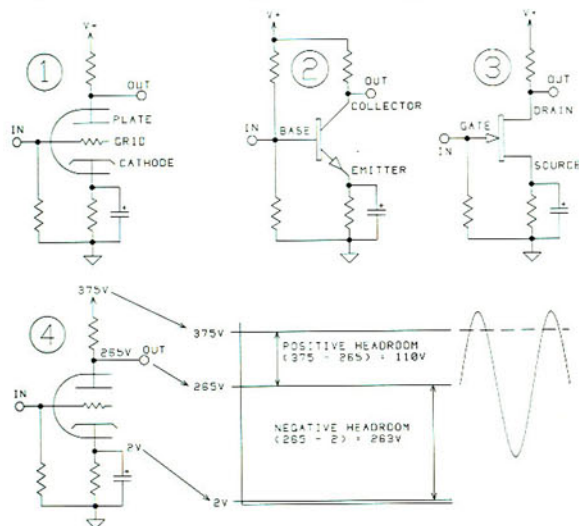
Tube Sound

To behold the guitar scene is to know an obsession; titled, to be sure, but harder to limn than the Louvre. *Tube sound* has a contextual definition. It means the warmth of a microphone with a triode preamp, but also the chime of an AC-30 and the wail of a '56 Pro. Even a CD player tempered by a triode claims the lineage of a Marshall stack. All these sounds emanate from tubes in the signal path; are, so far, impossible to get without tubes. A search for common factors fails to identify anything but thermionic valves.

Perhaps the most telling question asks how tube gain differs from solid state gain. Schematics yield few clues, for tubes and transistors share circuit topologies. The triode gain stage compares to the BPT's common-emitter amp, or the FET's common-source amp. All three stages invert, work at comparable relative operating points, and are biased class A (Fig. A117).

Circuits diverge in operating voltage. Transistors typically run at 12–50V, tubes at 250–450V. This gives tube amps a ratio of headroom to input level about 10 times transistor amps'.

In transistors, electrons and 'holes' flow through silicon doped with impurities. In tubes, electrons move in a partial vacuum. Sonic correlates of these facts re-

Fig. A117. 1—Triode schematic symbol; shows half a 12AX7 or similar tube, which contains two discrete triodes; heater filaments not shown. Tube's plate, grid, and cathode correspond to BPT's collector, base, and emitter; and to FET's drain, gate, and source. Like its transistor counterparts, this triode configuration inverts the signal. **4**—Illustrates headroom relative to operating point. In fact, OP shifts further upward when input reaches a level that drives grid positive relative to cathode, 2V in this case. The signal rounds off at the positive excursion long before hard clipping sets in.

main obscure.

One glaring departure is the output transformer, superfluous in solid state. Less obvious is a tube amp's low damping, the ratio of speaker impedance to output-stage impedance. Transistor amps possess very high damping.

Tubes are microphonic to a degree that might affect sound, even in amps culled of grossly microphonic samples. Transistors exhibit no significant microphonics.

Key differences emerge as input compares to output over a range of drive levels. The margin between linear transfer and clipping is slim in transistor stages. The typically biased triode resists clipping. Headroom exhaustion manifests as rounding off of peaks well before hard clipping sets in. Triodes also exhibit operating-point shift as input level rises, skewing the squashed waveform.

These differences lay an intuitive basis for solid-state emulations of tube sound.

Tube-Sound Emulation Patents
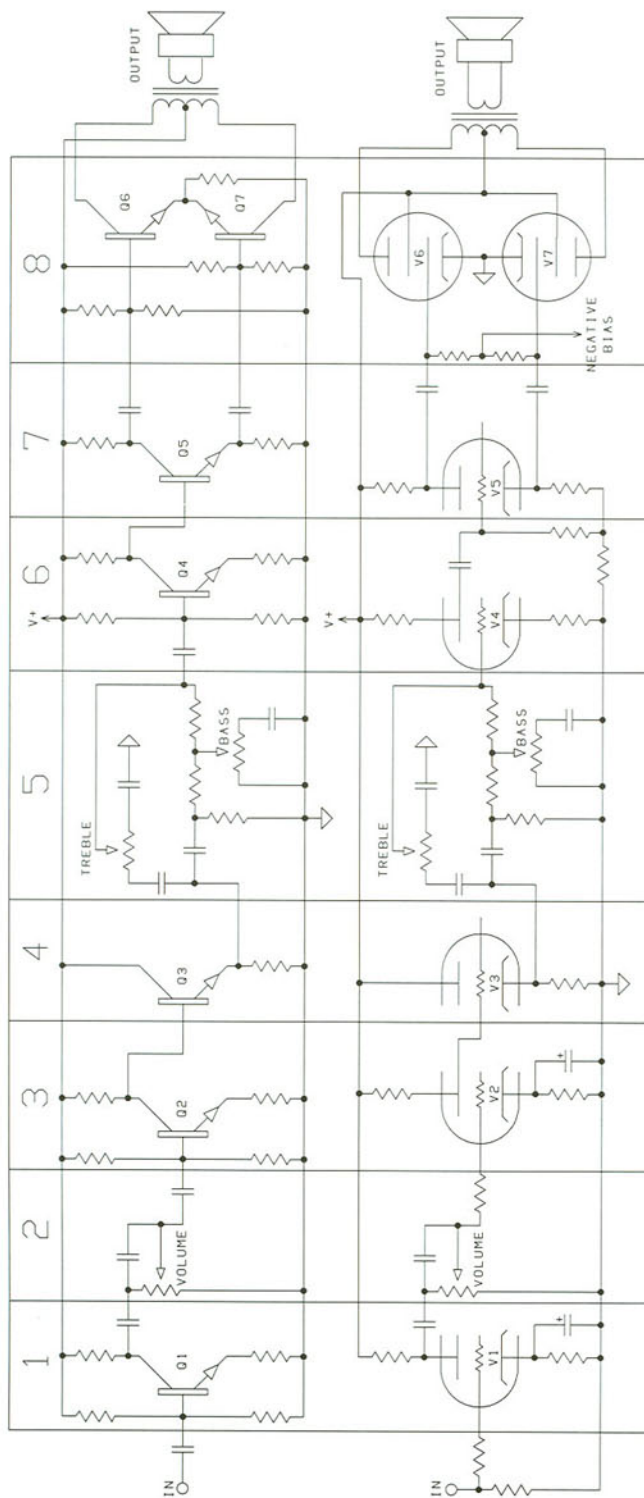Guitar amps evolved with a sense of dearth in straight gain. Clean amps, if not the font of bad, still finished behind amps that flunked the IHF distortion test. Senior pickers caught on quick. The rest of the world woke up when the industry fixed seeming flaws by replacing tubes with transistors. Though difficult to believe now, tube amps had a bad rap in the late fifties. Their nonlinearity was known from the lab; their bulk

8—Tube stage uses a pair of pentodes, biased class AB, configured as push-pull amp, feeding a transformer; can additionally squash the signal. Transistor stage uses two transistors biased class AB, also configured for push-pull operation, also feeding a transformer; linear unless clipped. Both stages may inject crossover distortion and distortion due to transformer saturation or ringing.

4—Tube amp uses triode configured as a cathode follower, the tube equivalent of an emitter follower. Does not invert; loses some voltage, but has low output impedance to drive the next stage. Solid state amp uses transistor configured as emitter follower with identical function.

6—3rd triode gain stage; inverts; skewed squashing is practically certain in this stage because signal measures tens of volts$_{p-p}$. 3rd transistor gain stage inverts; transfer remains linear unless driven to clipping.

2—Volume control with 'bright switch' permanently engaged; identical function in tube and solid state.

1—Triode gain stage, biased class A; inverts; gives linear gain unless input exceeds squashing threshold. Transistor gain stage, biased class A; inverts, gives linear gain unless driven to clipping.

3—2nd triode gain stage, biased class A; inverts. Skewed squashing is likely here, because signal amplitude is likely to exceed the squashing threshold. 2nd transistor gain stage, biased class A; inverts; linear transfer unless driven to clipping.

5—Passive tone network that predated the ladder circuit; results in voltage loss. Function is identical in tube and solid state amps.

7—Triode configured as phase splitter; no voltage gain, but being a triode, may additionally squash the signal at one or both ends. Single transistor configured as phase splitter; linear unless driven to clipping.

Fig. A118. Two guitar amp schematics, one a tube amp—a simplified version of a mid-fifties Fender 'Pro'—the other a discrete-transistor amp. The amps have identical topologies, yet change the signal in vastly different ways.

and heat were hard to ignore. Solid state promised clean delivery in trim kits, shrouded by an unspoken pledge of better sound.

The flaw in this logic was to assume that simple circuits mirrored simple transfer functions. The premise proved all too true in solid state, but tubes' innocence hid modes whose magic remains a mystery. Pickers found out how good they had it when they compared transistor amps to their old Titans and Twins. A burst of 'improved' tube amps after '65 exposed the folly of fixing something not broken, and dispelled remaining doubt that Leo had it right all along. Once tube sound flowered, the worth of vintage amps rose in kind, with a tacit bonanza for those who could wring that sound from solid state. Three decades and many patents later the quest remains on track, but so far no cigar.

Each tube-sound emulation patent forms a piecewise analogue. The subtext of those patents reflects an awareness of tube sound as a whole that, if not patentable, yet must guide the fusion of analogues. An emulation succeeds by capturing the sum of the tube amp transfer function, which splits naturally into preamp, power amp, and miscellaneous modes.

Preamp Modes

Most guitar preamps consist of cascaded triode gain stages. A triode is a tube with three electrodes: plate, grid, and cathode (analogous to collector, base, and emitter of a bipolar transistor; or to drain, gate, and source of a field-effect transistor). Resistors tied to the three electrodes set the tube's *operating point* (OP), the DC bias at the output electrode, in this case the plate. A key quirk is that the OP shifts upward as the input level rises (Fig. A121). Most sources ascribe this to the signal driving the grid positive relative to the cathode. Shift causes the signal to hit the wall at the positive excursion. Short of massive overdrive, the negative excursion never runs out of room. A second crucial aspect is that the signal lands softly, appearing rounded off rather than clipped. Unipolar squashing yields even harmonics of low order, compared to the high-order odd harmonics of bipolar clipping.

Because a triode's first departure from linearity manifests as skewed squashing, tube sound seems
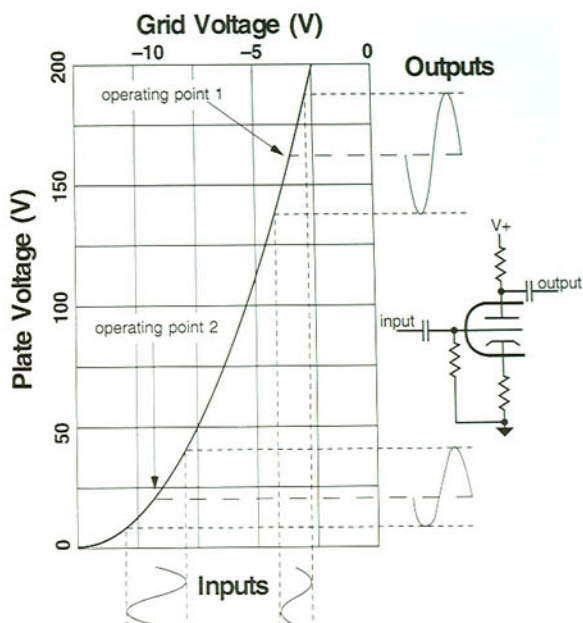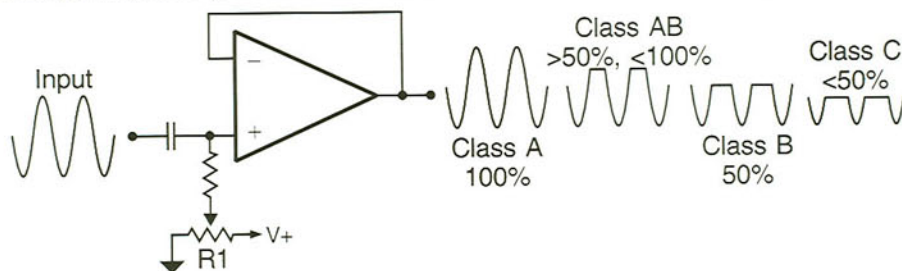

Fig. A119. Each vacuum tube exhibits characteristic input vs. output traits, typified by graph. Curve is substantially linear at high operating point (1), nonlinear at low operating point (2), where gain also drops. Curve is one reason most guitar amp triodes bias the plate at or above $\frac{1}{2}$V+; but doing so results in squashing as the operating point shifts upward when signal level becomes large enough to drive grid positive relative to cathode. Nonlinearity lives at both ends of the transfer function.

to require that the preamp run out of room. And since one stage fails to flesh out the sound, squashing needs to occur serially, the more stages the better, at least by accounts of richer sound in amps flush with triodes (Ref. 10; also, the original Ampeg SVT placed 11 triodes between the input and the power stage).

Each plate-output triode stage inverts. Constraint in the next stage gets the uncramped halfwave from the prior stage. After four stages, both ends may have been cramped twice: distortion of distortion; "harmonic sweetening," as some fancy. Because squashing is detectable below the level of audible distortion, and because transistor amps don't squash, this feature makes a likely explanation for the aspect of tube sound known as warmth. Analog tape, another warm

Fig. A120. Op amp with variable input bias illustrates various classes of operation. Class A bias means that the stage conducts 100% of the cycle. Class AB, less than 100% of the cycle, but more than 50%. Class B conducts 50% of the cycle, Class C less than 50%. Each tube or transistor of a push-pull stage is biased so it conducts during only half the cycle, and so that the polarity of the halves are opposite; the outputs of both devices combine to reconstitute the input signal.
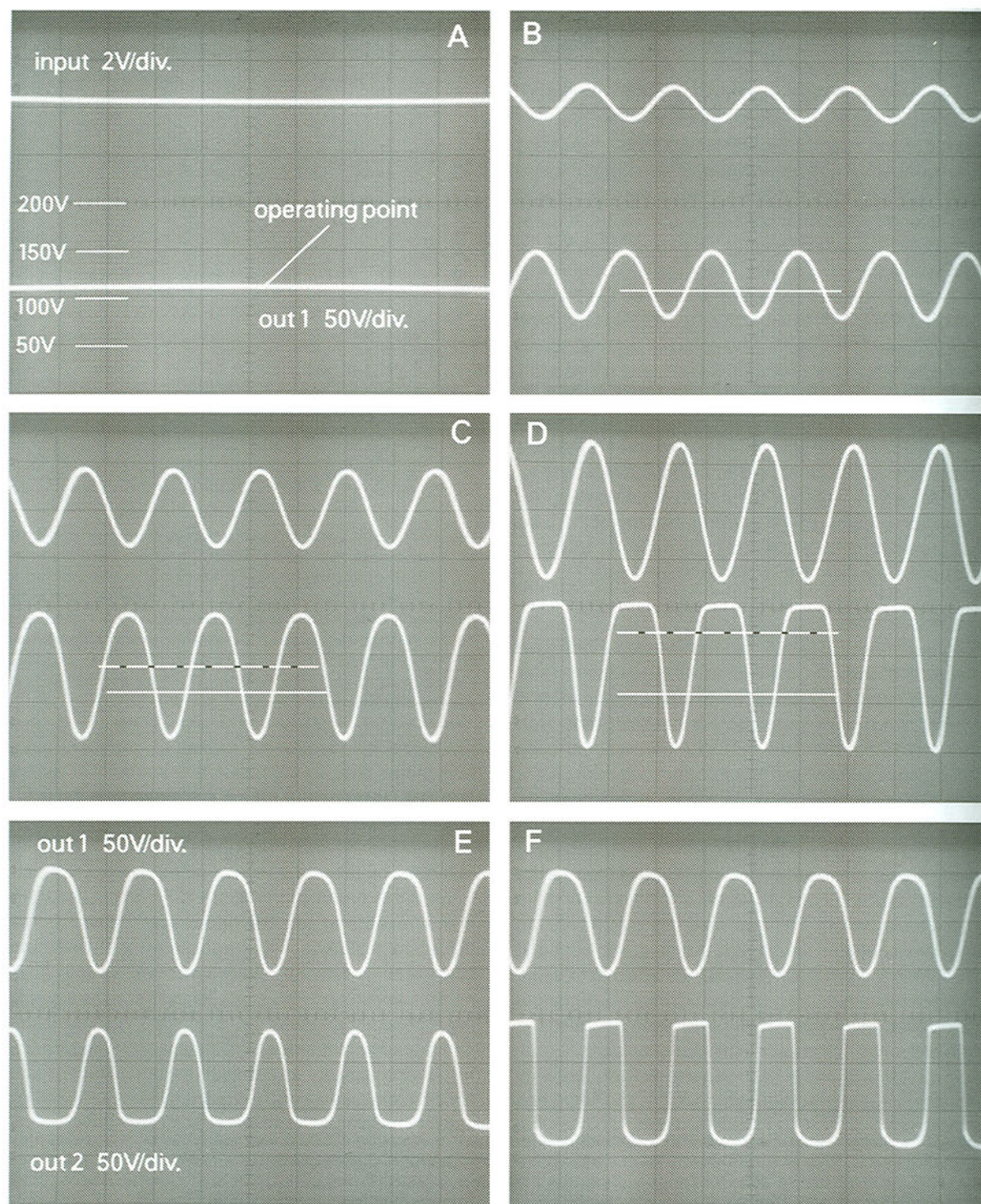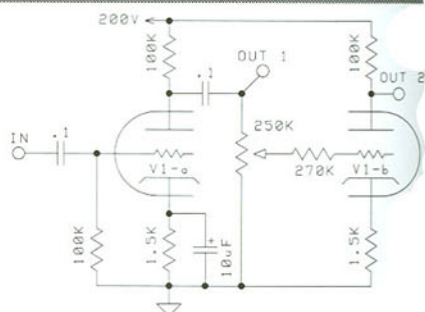
Fig. A121. I/O of schematic @ right (Sovtek® 7025/12AX7). A—No input; plate of V1-a measures 110V, the operating point (OP) on 200V supply. B—1-KHz sinewave is applied. Even at this level the OP has begun to rise. C—As level increases, two phenomena occur. First, the top of the wave rounds off: asymmetrical triode squashing. Second, the OP shifts further upward; white line at 110V shows original OP, dotted line shows likely new OP. Asymmetrical squashing is exacerbated in amps whose operating point exists well above ½V+ to start. D—Squashed end of wave squares up as drive level & OP rise. E—V1-a output feeds V1-b, whose output appears on bottom trace; 250K pot adjusted for roughly equal levels. F—With slightly more signal fed to V1-b, rounding-off at one end of the signal combines with clipping at the other end. Despite massive overdrive of V1-b, the lower excursion does not clip; in fact, it still has ~70V of headroom. This indicates extreme OP shift.

Fig. A122. Tube amp power output stages. 1—Class A; the transformer's primary winding makes the plate load; 6V6 is biased so that it's ON all the time. Input signal alternately turns it ON harder, or partly OFF. Negative feedback path is typical, but not essential. 2—Pu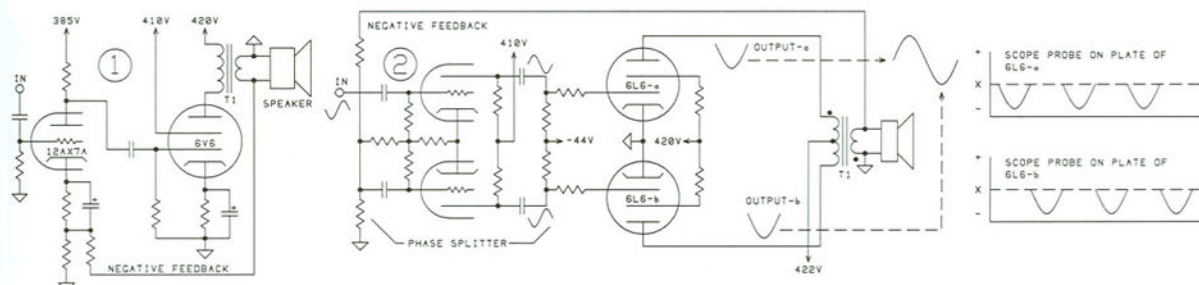sh-pull output stage, preceded by a phase splitter. A transformer forms the plate load of a pair of 6L6 pentodes. Both grids are biased @ −44V, placing them at the conduction threshold. Each grid receives a version of the signal inverted compared to the other. When one tube sees the positive swing and conducts, the other sees the negative swing and does not conduct. A scope probe tied to the plate of each 6L6 shows a train of negative halfwaves (relative to X, each plate's resting bias), because each tube turns ON only during positive input half-cycles, and the output at each plate is inverted. The plate of 6L6-a ties to the inverting end of T1 (note the phase dots). Inversion makes the negative halfwave positive, so all positive pulses in the secondary winding of T1 originate from 6L6-a. The plate of 6L6-b ties to the noninverting end of T1; the signal does not invert, so 6L6-b creates all the negative pulses in the secondary of T1. This mode is known as push-pull. As with class A amp, negative feedback path is typical, but not essential.

medium, saturates with squashing (Fig. A123).

Compression and distortion rise with drive level. Warmth breaks over into overt distortion, coupled with growing sustain. Clipping of half the wave may occur in later gain stages (Fig. A121–F). Fortuitous overdrive gave some amps their singing quality. Manipulation of where and how overdrive happens forms a key aspect of tailoring tube sound.

Passive tone circuits appear to play a dual role in the preamp. Their voltage loss requires an additional triode stage; their midrange dip emphasizes harmonics to subsequent stages.

Power-Amp Modes

Guitar amps use pentode power tubes. A pentode is a tube with five electrodes, of which only four may be accessible; the fifth electrode is internally wired to the cathode in "beam pentodes." Common pentodes come one to a tube, a much bigger bulb than a dual triode. Their sound depends greatly on bias and drive level. Class A bias means the tube is 'on' all the time; significant current flows through the plate load even with no signal applied. The plate load is a transformer winding with an impedance of 2500–10,000 ohms, compared to the preamp's 100K+ plate loads. The audio signal alternately causes more and less current to flow through the transformer. The approach offers poor efficiency and short tube life, but delivers a sound distinct from push-pull stages'. Several tube amp boutiques specialize in the class A sound.

Most classic amps use paired pentodes biased for class B or class AB operation. This *push-pull* arrangement means both tubes are turned barely off in the absence of a signal. They conduct upon receipt of a positive input; but since the signal feeding each tube is antiphase relative to the other, each tube turns on for half a cycle while the other tube is off. Both tubes' plate

loads consist of a transformer winding, where the signal halves reconstitute the whole. The design is more efficient than class A, allowing higher power and longer tube life. While the push-pull approach cancels even harmonics (and power-supply ripple), mismatch between the tubes is said to generate those harmonics. Mismatch may also exist in coupling caps and bias resistors in the phase splitter driving the push-pull stage.

Bias level affects the sound of both bias classes. Tubes biased such that too much current flows at rest wear out quickly and lack power. Tubes biased for too little resting current suffer distortion and weak sound. Improper bias of push-pull stages also heightens crossover distortion.

Irrespective of operating mode, power tubes feed the speaker through a transformer. The transformer has a sound, often unnoticed until a unit gone bad is

Fig. A123. Top trace input, bottom trace output of open-reel tape deck @ 3.75 ips, 1 KHz sinewave. Tape saturation manifests as symmetrical squashing. Composite photo necessitated by deck's wow & flutter. Scale both 500 mv/div.
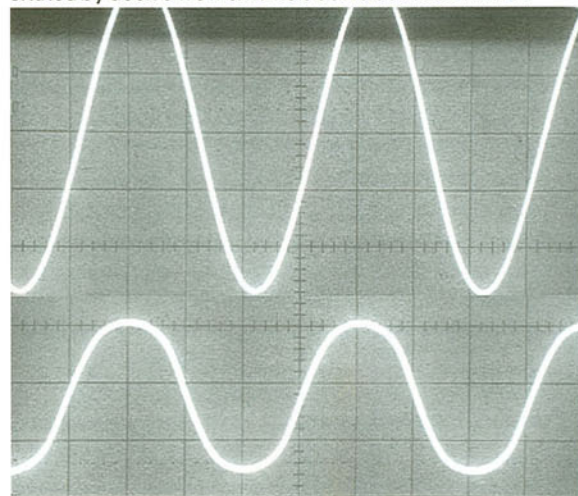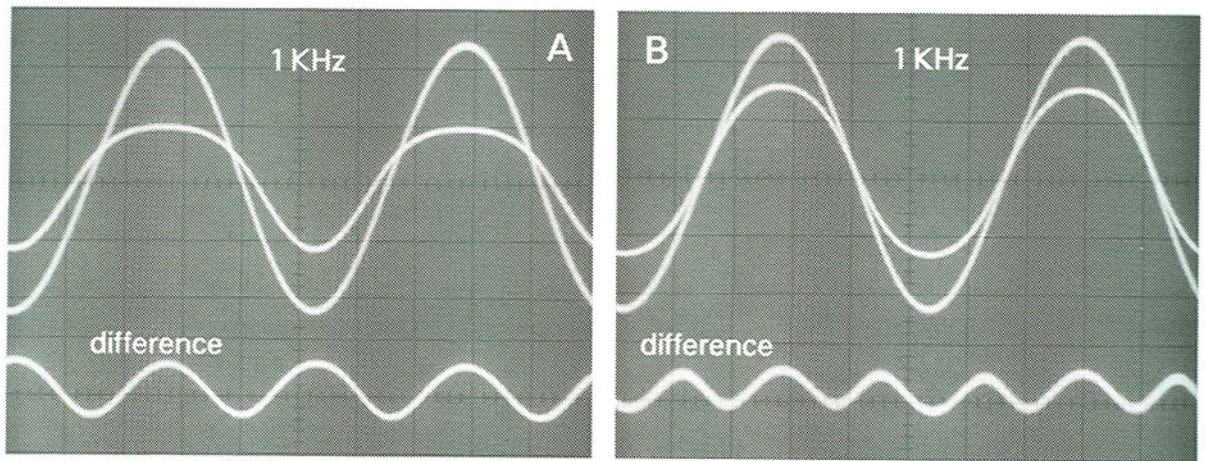
Fig. A124. A—Composite photo demonstrates that unipolar squashing generates predominantly second harmonics. B—Composite of symmetrical squashing demonstrates mainly third harmonics. In both photos, bottom trace was obtained by normalizing the two top traces for amplitude, then subtracting them.

replaced with one giving a different sound. Sonic differences have been traced to core material, winding arrangement, potting, and proximity of turns to core. Two transformers rated 10,000:8 ohms but built differently are likely to sound different. As with triode squashing, core saturation doesn't have to be blatant to affect the sound.

Most power stages contain a negative feedback path from the transformer output winding to the phase splitter in push-pull amps, or to the pentode driver in class A amps. Without negative feedback, the transformer acts as a bandpass filter that rolls off below ~200 Hz and above ~4500 Hz. Negative feedback widens bandwidth and lowers distortion.

Damping factor tells the ratio of speaker impedance to output stage impedance. Transistor output stages typically show an impedance well under one

Fig. A125. Damping factor, the ratio of speaker impedance to amp output impedance, affects frequency response because speaker's impedance is not constant. In this example, peak at resonance results in 30% increase in voltage in the parallel amp/speaker circuit of tube amp; less than 2% change in solid state circuit. This simplified analysis ignores current, but conveys the nature of the phenomenon.

ohm, giving a high damping factor. A tube amp's net output impedance often equals the speaker's nominal impedance, giving a damping factor of one. The speaker's true impedance varies with frequency. A peak of several times nominal impedance occurs at the speaker's free-air resonance; a gradual impedance rise above ~1 KHz is due to voice coil inductance. So long as the drive stage has a much lower impedance than the speaker's, these nonlinearities bear little on sound. But if output impedance is high relative to speaker impedance, Ohm's law dictates a rise in voltage transfer when speaker impedance rises. The result is a hump at the speaker's free-air resonance, followed by a dip, followed by a gradual rise above 1 KHz (Fig. A125).
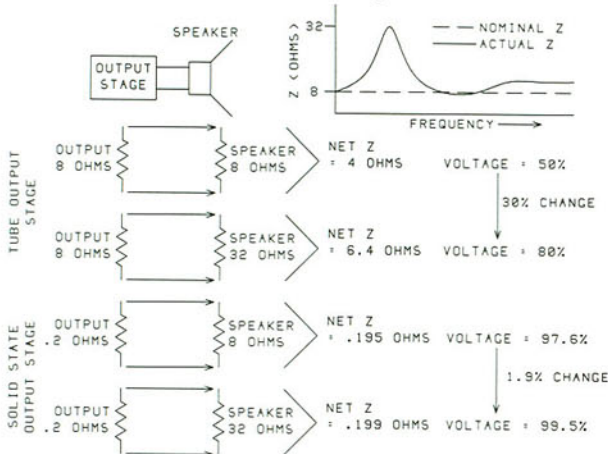
Despite old and simple topologies, power circuits are functionally more complex than preamps, especially when overdriven. Interdependent factors include ringing and saturation modes in the transformer, drive-dependent crossover distortion, and a negative feedback path. Sound also depends on whether the cathodes tie directly to ground ("grid bias"), or tie to ground through resistors ("cathode bias").

Miscellaneous Modes & Related Factors

All tubes are to some degree microphonic. The text cannot commend the exercise, but notes that thumping a preamp tube in a Fender Twin cranked up to 10 causes an audible pop in some amps. Chassis vibration affects sound by this mode. These days purists root out microphonic tubes, a luxury few indulged during the rise of rock 'n' roll.

A minor factor, but one cited often enough to mention, is a fall in volume coincident with overdrive, due to a fall in power supply voltage under heavy current demand. The phenomenon is called *sag*; but before the voltage sags, the filter caps impart their charge. The audible burst of energy preceding sag is called
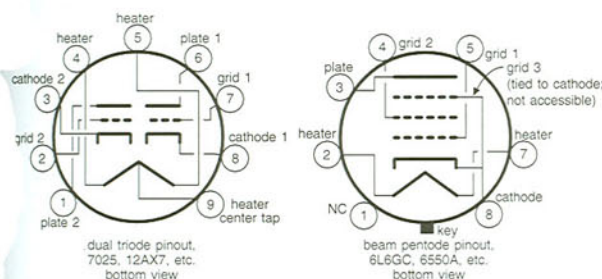
Fig. A126. Pinouts of the two most common types of tubes used in guitar amps. Left shows dual triode of the 7025/12AX7 series. Right shows pentode of the 6L6 series, whose third grid is not externally accessible.

punch.

Classic tube amps matured in tandem with specific accoutrement. Speakers with flimsy cones give classic sound better, or at least at lower drive levels, than speakers resistant to breakup. The cone must be made of paper, rather than Kevlar, polypropylene, or aluminum.

Enclosure resonance bears on the sound of any amp. Whacked particleboard goes *thwok*, plywood *thwoop*, solid pine *thwunng*. Amps built from resonant lumber—comparatively thin pine, for example—predominated in tube amps' golden era, and yield a sound some players prefer.

Old tube amps compare to snowflakes: no two alike. Part of the explanation lies in the 20% tolerance of parts from which the amps were built. Besides that, resistors and capacitors shift with age. After three decades their values may depart from spec by tens of percent. Shifted resistor values alter gain and bias; shifted cap values alter frequency response, and power-supply ripple and reserve.

Tube Sound Emulations

Tube sound is big business, even when it flows from solid state. Flip through any gear directory and count the transistor products that claim to generate the sound of a tube amp.

To recap, tube sound—at least the current understanding of it—flows from:

- ▶ asymmetrical, serial squashing in the triode cascade
- ▶ a large ratio of headroom to input voltage
- ▶ punch & sag
- ▶ the sound contributed by a complex reactive output system, including transformer ringing/saturation, crossover distortion, negative feedback, and spectral shift due to low output damping
- ▶ sonic alteration due to microphonics

The composite function is complex, for tube amps deliver more than the sum of their parts. Within the scope of realistic expectations, the experimenter

might approach emulation by comparing tube amp functions to solid state circuits that pervert signals in a similar way. For instance, several solid state circuits squash, including linearly biased CMOS inverting buffers and overdriven OTAs, along with analog multipliers configured to approximate the sine function, and transistor/diode networks built to convert triangle waves to sinewaves. Squashing in the power stage has been emulated using MOSFET power transistors driven by BPT transistor stages that run at higher voltage. The BPT stage never runs out of headroom and thus does not clip; the MOSFET stage runs out of room, but tends to squash rather than hard-clip.

Punch and sag can be realized many ways, such as a compressor whose attack delay allows a peak to precede the fall in volume due to compression.

Crossover distortion is realized in many diode and transistor circuits, several reviewed in the prior Appendix.

Damping-related spectral shift can be produced by a simple active filter network, and also by a solid state amp that actually alters damping for specific frequencies by selective feedback paths.

With most functions nailed, it seems prudent order them as they occur in a tube amp: asymmetrical squashing precedes crossover distortion, which precedes spectral skewing. Tone and volume controls oc-
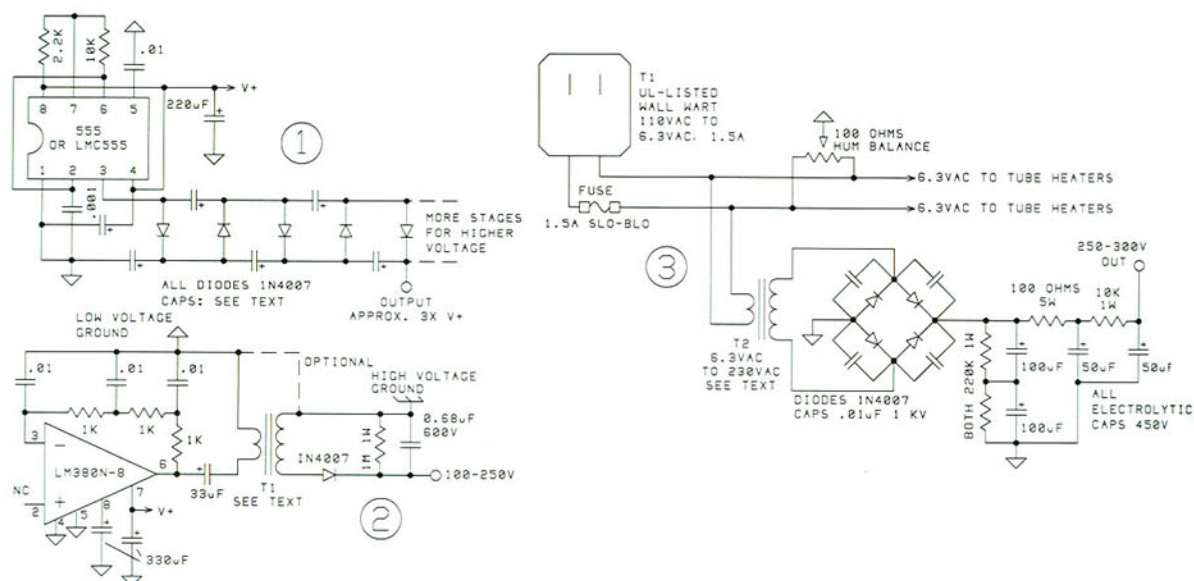
Fig. A127. High voltage supplies suitable for experimenting with triodes without having to deal directly with the AC line voltage or buy a full-size tube-amp transformer. 1—Ultrasonic oscillator feeding voltage multiplier made up of diodes and capacitors. Caps are 0.1–1 μF ceramic or electrolytic, with voltage rating at least twice that of the final output voltage. Supply voltage limit to 555 is 15V. 2—Author's breadboard supply for triodes, an LM380 configured as an ultrasonic phase-shift oscillator driving step-up transformer T1, whose output is rectified and filtered. Output depends on T1 and supply voltage to the 380. With the 380 running @ 17.5V, a Mouser 42TL013 8:1K transformer gives a solid 100V; a Magnetek 12:7500 transformer gives more than 250V unloaded, about 200V driving several triodes with 100K plate resistors. Neither transformer is designed for this application, and could fail. The 380 (8-pin version) needs a stout supply, 10–22V @ 500 ma, and requires a heatsink. 3—Generating full tube-amp voltages. A 6.3VAC wall wart drives tube filaments directly; 100-ohm pot reduces hum by pseudo-balancing the filament drive voltage. T2 is a low-wattage unit with a dual 110/230VAC primary winding. T1 drives the 6.3V winding; output is taken off the 230VAC winding; rectified in bridge; filtered by a standard RC network. Additional RC stages may be needed to decouple more than two triode gain stages in series. Filament supply should easily handle two 12AX7s. All three circuits, especially #3, generate potentially lethal voltage. Using these circuits is for experts only.

cupy the same relative place in the emulator as they do in the amp.

Patented means of achieving these functions are summarized at the end of this Appendix.

Tube vs. Tube

Like most cults, tube sound grew orthodoxy out of zeal; spun ritual off orthodoxy. The rituals clustered under a new rubric, gauging life in terms of tone, in gradations so fine that not everyone hears them. But what looks at first to be hairsplitting resolves into a palette of tube amp shadings; of differences not just audible, but measurable. Individual tubes of the same type, made in the same plant from the same batch of materials on the same day, exhibit measurable differences. These include microphonics, gain, and noise, plus the tube's quotient of gas and the strength of its vacuum. Microphonics and gas justify discard; noise identifies tubes best used outside the signal path, as in tremolo oscillators. Screening for distortion—particularly, the distortion threshold and the gain/distortion ratio—gives an index said to allow groupings of tubes so as to maintain the sound of amps equipped with similarly grouped tubes. It also lets the player alter the sound of the amp by changing to tubes having

a higher or lower gain/distortion ratio. (Ref. 25).

The Mod Cabal

Tube amps have been under the microscope now for a generation. In the course of that scrutiny they have given up many empiric rules for good sound. Couple this lore with a plenitude of amps that could sing but don't, and you have a thriving trade modifying tube amps. Common mods include:

▶ power stage bias & balance; this affects distortion, efficiency, and tube life
▶ adapting amps originally built around now-defunct tubes (6146B, 7027, 12DW7, etc.) to tubes still made
▶ preamp bias; this usually requires changing the plate resistor and/or the cathode resistor, which shifts operating point, and sometimes gain
▶ decreasing negative feedback; since tube sound is distortion, some players prefer the sound they get by raising the value of the feedback resistor, or removing it
▶ various measures to increase preamp gain: use of higher-gain tubes, decreased loading to

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transfer more signal, disconnection of reverb/tremolo mixing circuits, removal of stability caps to extend treble response; occasionally, addition of one or two gain stages to the preamp

- ▶ addition of new and different preamp stages running in parallel with the existing stage, and selectable by a switch; some mods run two different-sounding preamps in parallel, and mix their outputs
- ▶ altering frequency response by changing the values of coupling capacitors, tone control caps, or cathode bypass caps
- ▶ convenience & safety mods, e.g., an effect send/return loop; a resistor across the transformer output so the power tubes are never completely unloaded; addition of earth ground to amps with two-prong plugs
- ▶ configuring a modern amp to give the sound of a vintage amp, either as the vintage amp sounded when new, or as it sounds now; the mod is called *backdating*

Tube Stomp Boxes

Since tubes work much of their magic at low plate voltage, the market has seen a proliferation of tube-based boxes. Most are preamps; some add tone networks. For the stomp-box builder, the transition from transistors to tubes comes as naturally as the prior generation's move the other way. Tubes have slightly different biasing requirements than transistors, radically different safety requirements. Their basic topologies resemble transistor circuits', except that no tube analogues exist of PNP bipolar transistors, or P-channel field effect transistors.

A tube needs two power supplies, one to heat the cathode, one for the plate. Triode filaments require 12.6VAC @ 150 ma if wired in series, or 6.3VAC @ 300 ma if wired in parallel. The latter mode is most common, because pentodes' filaments also run at 6.3VAC. Running all tubes at the same filament voltage simplifies the power supply. Proximity of a powerful AC signal to the electrodes causes hum, reduced by pseudo-balancing the heater supply with fixed resistors or a pot. Running the filaments off filtered DC gives quiet-

er results and is common in audiophile products, rare in guitar amps.

Triodes in a typical tube amp see 200–375VDC on the hot end of the plate resistor. The fact that they run at a fraction of that voltage puts them within reach. The tube version of the Aphex Aural Exciter ran at 16V. This limits headroom and lowers the distortion threshold, which is exactly what many players want.

Tubes pose the problem of deriving what is for a stomp box high voltage—say, 50–100V—from a low voltage supply. Common methods to achieve this include the well-known charge pump/voltage multiplier, and a step-up transformer driven by an ultrasonic oscillator. Filament-voltage AC can be stepped up to 250–300V by reversing a small step-down transformer (Fig. A127–3). Heater and plate supplies equip the expert builder to remake the entire preamp chain of classic amps, including their vibrato, tremolo, and reverb circuits.

Experimenting with tube circuits calls for parts whose voltage rating exceeds the supply voltage, and for strict safety procedures to avoid potentially lethal electric shock. To repeat the warning at the beginning of this Appendix: **There is no 100% safe way to work with lethal voltage. Anyone working with high voltage assumes all risks. The author presumes that anyone choosing to experiment with tube circuits is an expert.**
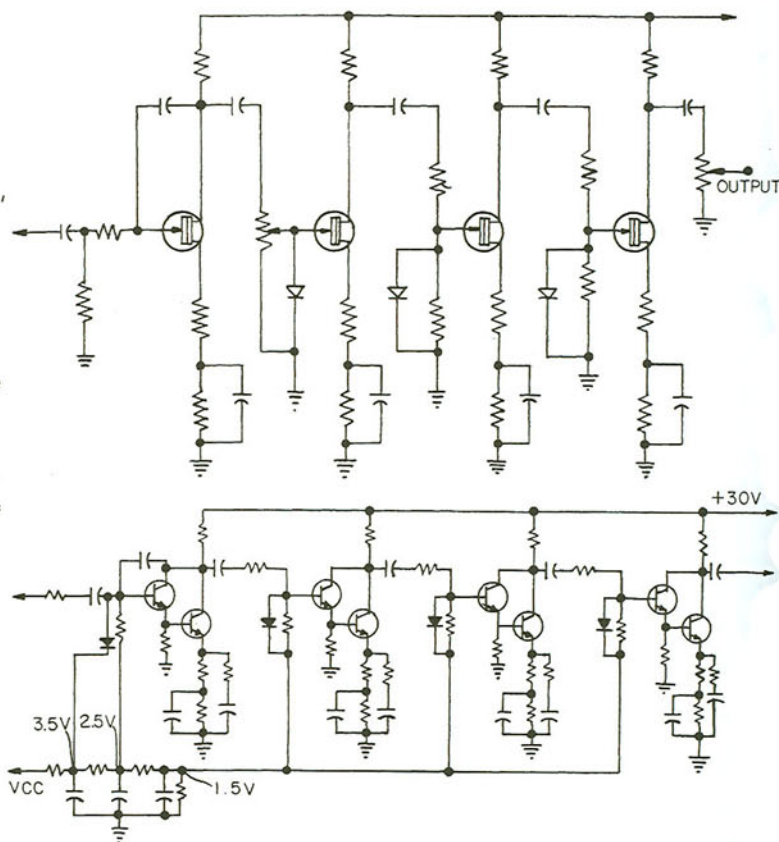
The Nature of Tube Sound

Understanding what tubes do sheds little light on why people dig the sound. Tube amps, as others have observed, are musical instruments. Massive evidence paints them as distortion engines. Tube sound emulations have succeeded in the manner of floral wallpaper or wood-grain Formica that captures look but not essence. The difficulty of recreating tube sound argues for an open mind when assessing these systems. Tube amp schematics mirror simple, almost primitive beasts that prove fussy enough to postulate interactive, if not chaotic modes. Everything, it sometimes seems, affects the sound of a tube amp, in ways neither expected nor observed in transistor amps of comparable topology.
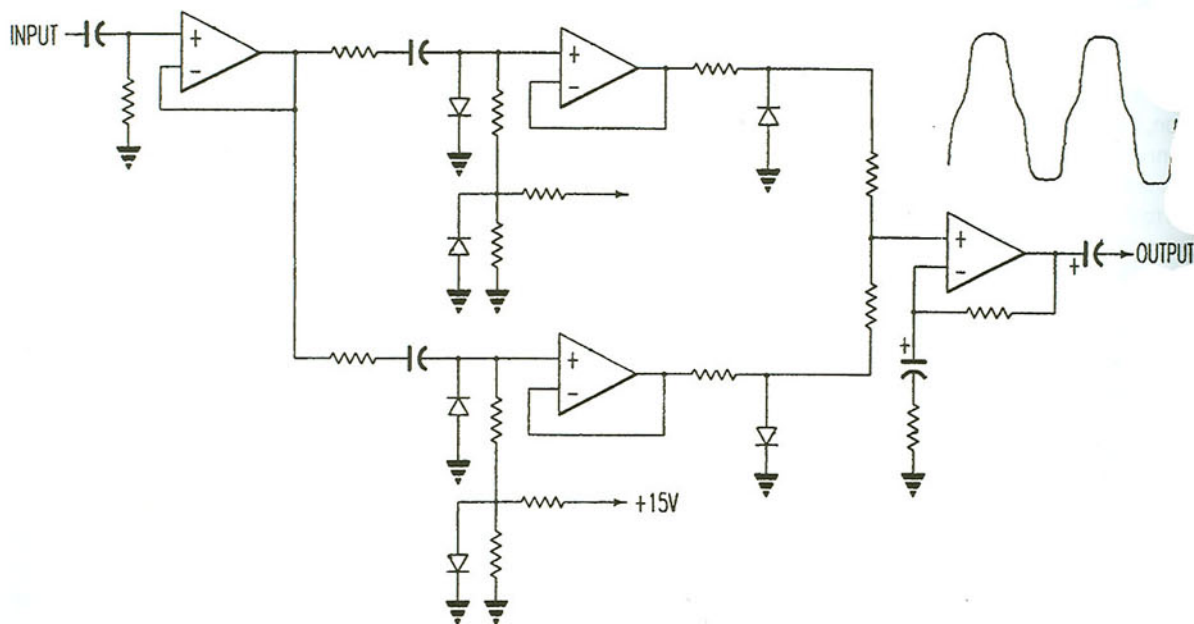
Tube-Sound Emulation Patent Summaries

All diagrams shown with summaries were scanned from their respective patents. The many identifying numbers and lines that normally appear on patent schematics have been removed for the sake of clarity.

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No. 5,619,578 "Multi-Stage Solid State Amplifier that Emulates Tube Distortion" Recreates the asymmetrical squashing of serial triode stages. Uses resistor-diode networks as nonlinear function blocks; uses FETs or Darlington-connected BPTs as gain/inversion stages. Specifies relatively high supply voltage, 40V for FETs, to get closer to the high positive-headroom-to-squashing-threshold ratio found in tube amps. In a typically biased triode, this comes to 100V/1.5V = 67; in the solid state circuit, if diodes turn on @ 0.5V, ratio = 18V÷0.5V = 36; patent indicates that this is high enough to give the desired quotient of second harmonics; patent cites this as a key factor in the sound of a tube amp; finds biasing less critical to proper performance in Darlington circuit than in FET version.

No. 5,524,055 "Solid State Circuit Emulating Tube Compression Effect" Recreates phenomena associated with overdrive of tube amp push-pull power stage: crossover distortion, and diode-like clipping; achieved with resistors, caps, diodes, and op amps. Waveform below right contains deadband due to of crossover distortion, flattened tips due to diode clipping.
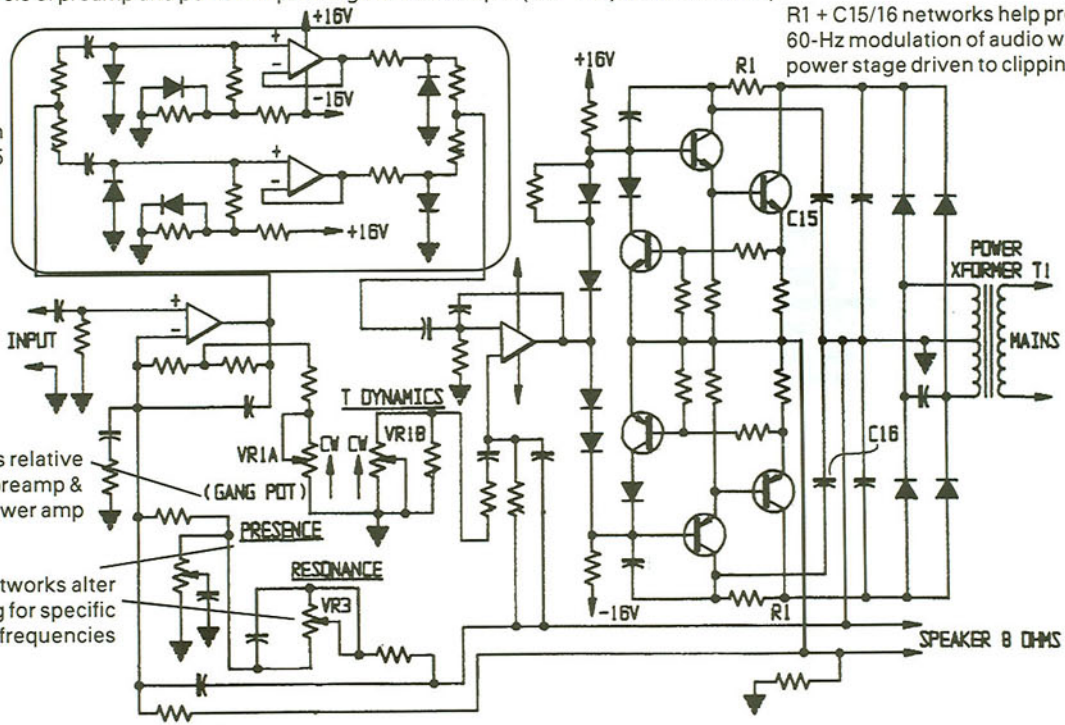
No. 5,675,656 "Power Amplifier with Clipping Level Control" Details solid-state amp that emulates several tube-amp functions; incorporates the crossover distortion/diode clipping network of patent No. 5,524,055; emulates damping-related spectral shift by altering feedback for pertinent frequencies (also a patented technique, No. 5,197,102); minimizes 60-Hz modulation in the power amp during clipping by supplemental RC decoupling networks; also allows variation of relative drive levels of preamp and power amp through use of dual pot (the 'T. Dynamics' control).

R1 + C15/16 networks help prevent 60-Hz modulation of audio with power stage driven to clipping

circuit from pat. 5,524,055
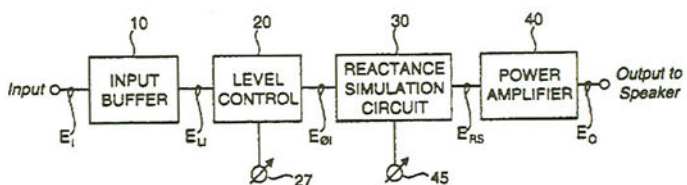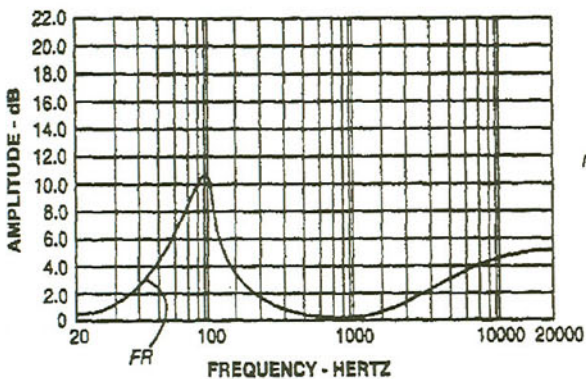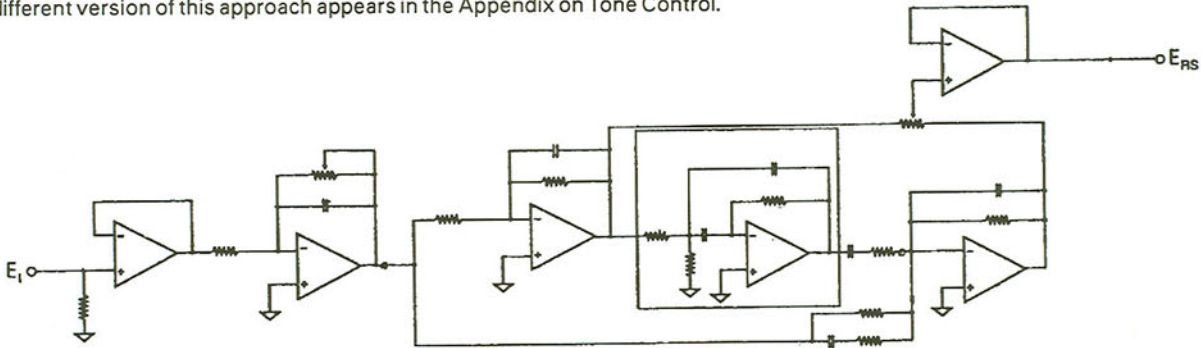
dual pot alters relative drive level of preamp & power amp

RC networks alter damping for specific frequencies

INPUT

T DYNAMICS

VR1A   VR1B
CW CW
(GANG POT)

PRESENCE

RESONANCE
VR3

+16V

R1

C15

POWER XFORMER T1
MAINS

C16

-16V   R1

SPEAKER 8 OHMS
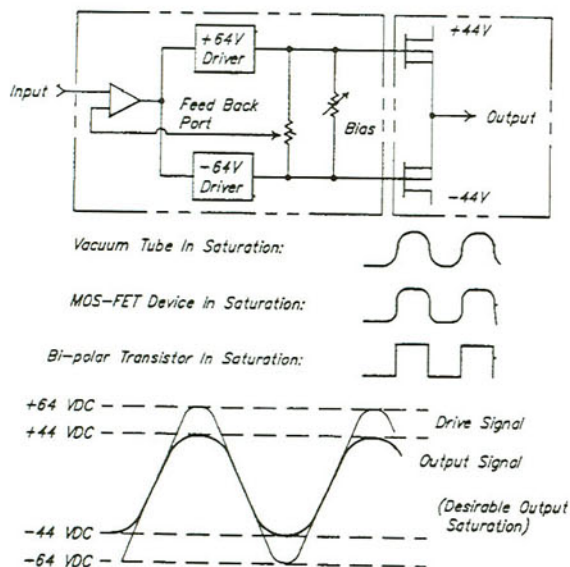
No. 5,268,527 "Audio Power Amplifier with Reactance Simulation" Uses standard active filter techniques to produce the tone curve associated with low tube amp output damping. Schematic shows one embodiment of the invention; a slightly different version of this approach appears in the Appendix on Tone Control.

E_{RS}

E_I

FR

FREQUENCY - HERTZ

AMPLITUDE - dB

| Input | INPUT BUFFER | LEVEL CONTROL | REACTANCE SIMULATION CIRCUIT | POWER AMPLIFIER | Output to Speaker |

10   20   30   40

E_i E_{LI} E_{OI} E_{RS} E_O

27   45

No. 4,987,381 "Tube Sound Solid State Amplifier"
Bascially, consists of BPT preamp stage running @ ±64V,
driving a MOSFET push-pull power stage running at ±44V.
The MOSFETs run out of headroom before the BPT drive
stage does; the MOSFETs tend to round off the signal
rather than hard-clip.
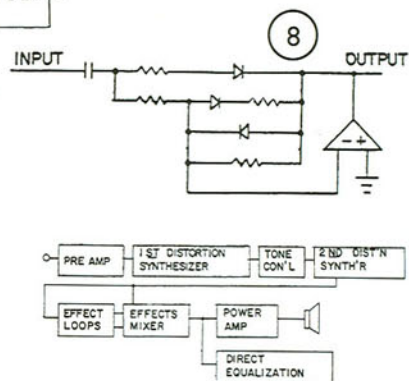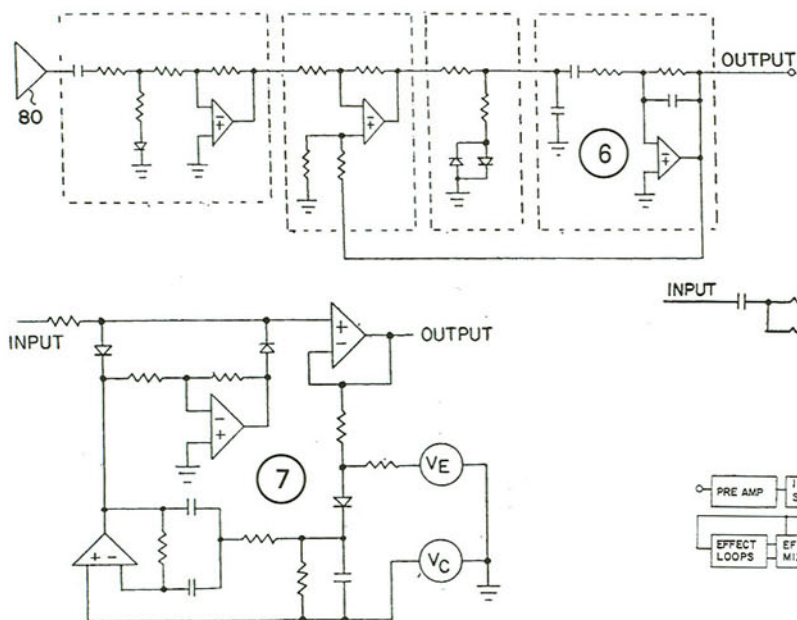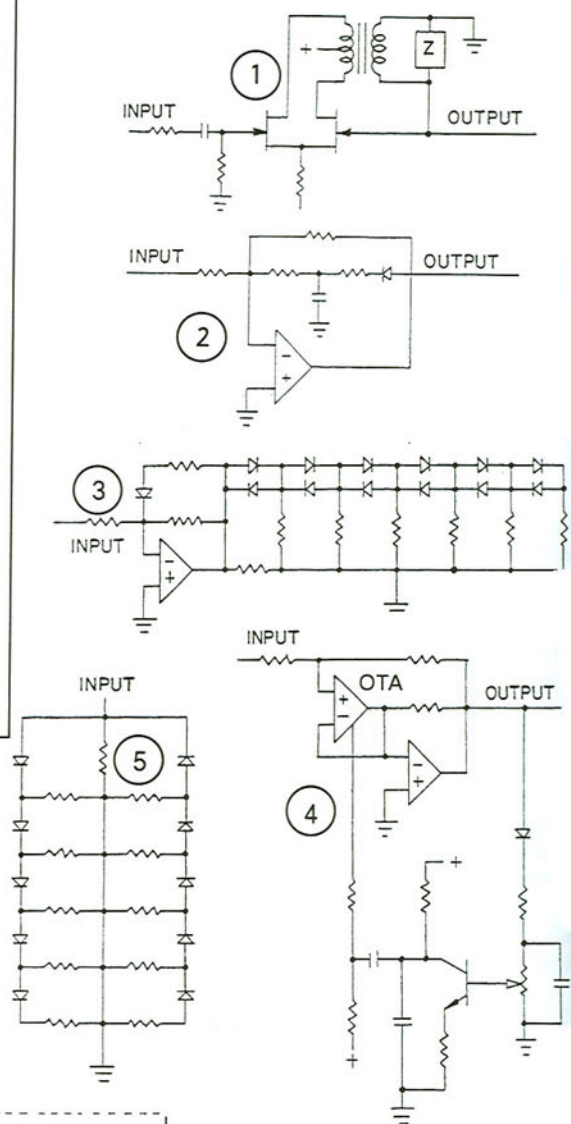
Vacuum Tube In Saturation:

MOS-FET Device In Saturation:

Bi-polar Transistor In Saturation:

+64 VDC ———
+44 VDC ———   Drive Signal
                               Output Signal
                               (Desirable Output
-44 VDC ———   Saturation)
-64 VDC ———

No. 4,995,084 "Semiconductor Emulation of Tube
Amplifiers" A fairly comprehensive tube amp emulation
that includes sag, punch, operating point shift, asymmetri-
cal squashing, and crossover distortion, achieved in a
number of novel circuits; includes sample algorithms to
realize functions through digital signal processing. 1—
distortion circuit; 2—bias shifter; 3—harmonic distortion
generator; harmonic predominance even/odd can be
determined by choice of resistors; 4—punch emulator;
5—variable gain circuit; 6—distortion circuit; 7—sag
simulator; 8—operating point shifter.

Noise Reduction

Noise remains a fact of most players' lives, and one exacerbated by daisy-chained pedals. Fortunately, many options exist to reduce this noise.

Single-Ended Noise Reduction (SENR)
Single ended denotes noise reduction applicable to any feed. Downward expansion describes the process of passing the signal through a VCA that reduces volume when the level falls below a threshold. An infinite expansion ratio equals gating. The process affects all frequencies equally.

A related approach exploits the fact that high-frequency audio masks hiss, which happens to be the most intrusive noise. Substituting a variable lowpass filter for the VCA confines the effect to treble. The filter is designed to open when a weighted level detector senses treble in the program, since treble masks hiss. The filter closes when treble subsides, reducing noise without affecting bass and minimally affecting midrange. The filter could use any slope, but usually accepts a mild slope in trade for inconspicuous operation. An incarnation of this approach known as Dynamic Noise Reduction (DNR®) surfaced in add-on processors and Delco radios in the eighties. DNR remains available in the LM1894 integrated circuit.

Transparent operation requires that expander or filter open in ~500 μs, close in 50–100 ms. Slower attack blunts the leading edge of treble transients; delayed decay causes hiss to trail each peak.

The nature of SENR forces a compromise between noise reduction and loss of information. Too low a threshold lets noise keep both elements fully or partly open; too high a threshold leads to program loss when the system fails to open.

SENR can reduce noise greatly, but artifacts of the process parallel the degree of noise reduction. Combination filter/expander systems operate more discreetly than single-mode systems giving the same amount

> ## Noise Reduction Techniques
> Single Ended
>   downward expansion/gating
>   voltage controlled LP filtering (e.g., DNR)
>   combinations (e.g., HUSH)
> Complementary
>   static systems
>   quasi-companding
>   treble emphasis/de-emphasis
>   dynamic systems
>   treble only (e.g., Dolby B, C)
>   wideband companding (e.g., dbx)

of noise reduction. Fifteen dB of expansion with 10 dB of filtering seems a common mix, and one embodied in the SSM2000 HUSH® integrated circuit. Though based on several patents, this chip stands apart by virtue of an automatic threshold mechanism, explained in Fig. A131.

The pursuit of better SENR has spawned other improvements. For example, deriving the control voltage from a compressed replica of the signal expands the control range. An internal compander with mistracking between compressor and expander automatically gates the signal with less pumping than is usually associated with low-level expansion (Ref. 45; technique is incorporated as an option in Project No. 23).

Complementary Noise Reduction
Complementary systems encode a signal prior to transmission or storage, decode by applying the inverse change upon reception or playback. Static versions use fixed changes, dynamic versions apply level-dependent changes.

Static Systems
Quasi-companding is a static system that boosts the signal before processing; reduces it by the same amount after processing. Noise acquired in transit
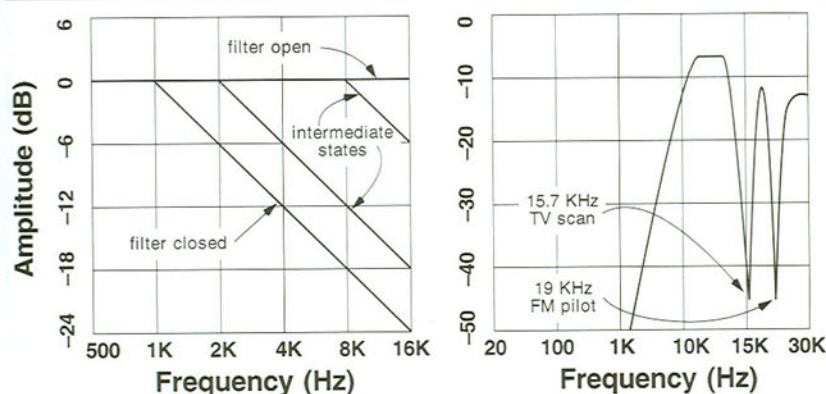
Fig. A129. Left graph illustrates response of voltage controlled lowpass filter; corner frequency just above 1 KHz, slope 6 dB/octave. Right graph shows response typical of highpass filter placed ahead of level detector that controls the lowpass filter. Sharp 36 dB/octave slope prevents filter from responding to bass or most of midrange; rackmount products may include notch filters for television scan frequency and FM radio pilot tone; these are not usually needed in a stomp box.
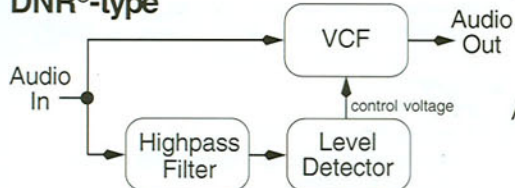
Downward Expansion
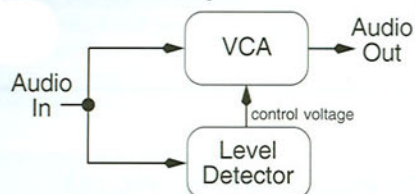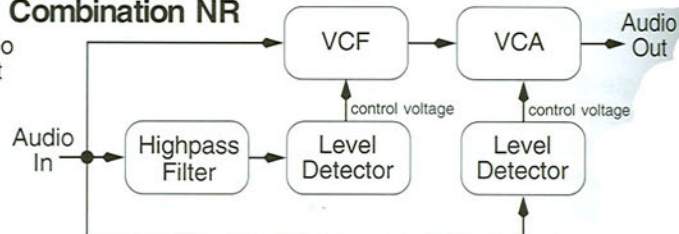
Improved DE
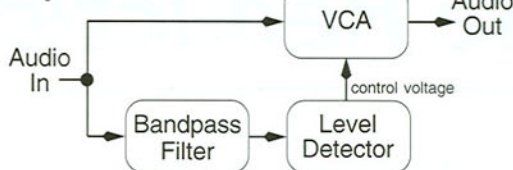
DNR®-type
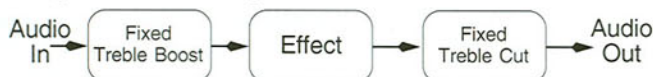
Combination NR

Quasi-Companding

Emphasis/De-emphasis

Dolby® B & C

Companding (e.g., dbx®)
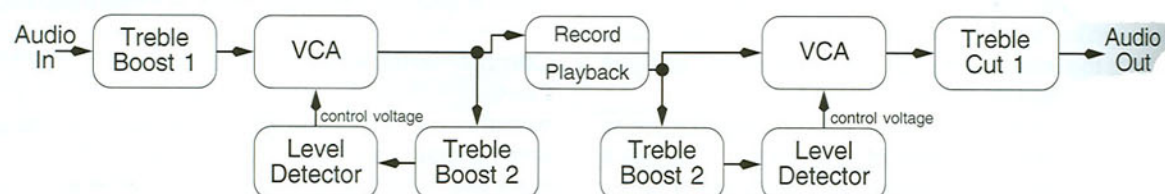
Fig. A130. Block diagrams illustrate several noise reduction modes. Downward expansion is a reactive mode that requires a level detector, driving a VCA. All frequencies are affected. Improved downward expander precedes level detector with bandpass filter that reduces pumping, which occurs when system reacts to subsonic energy. VCF-based NR, such as DNR, affects only treble, mainly above 1.5 KHz. Sharp highpass filter precedes level detector, to prevent response to midrange and bass. Combination systems use moderate amounts of downward expansion with lowpass filtering to achieve net noise reduction on the order of 25 dB. The self-adjusting HUSH version is discussed in more detail on the following page. Quasi-companding is a complementary system that manipulates overall level; emphasis/de-emphasis type acts only on treble. Dynamic complementary systems Dolby B & C apply level-dependent treble emphasis on recording, level-dependent de-emphasis on playback. Bottom diagram shows the workings of a typical companding system, such as dbx. Treble emphasis precedes compression, whose feedback control loop contains a second emphasis network to compress treble more than bass. On expansion, the parallel control path contains the same treble emphasis to expand treble more than bass, and a de-emphasis network following expansion to match the treble emphasis applied before compression. Overall noise reduction typically exceeds 30 dB.

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falls by the amount of gain reduction. Players can apply this to most effects by running them at as high an internal level as possible, such that gain reduction is needed to tailor the signal to the final output.

A more common approach limits emphasis to the treble. Hiss introduced by the effect falls by the

amount of de-emphasis. This method capitalizes on the fact that the band above 2 KHz is the main repository of noise, but contains less musical energy than the band below 2 KHz. This allows greater emphasis compared to quasi-companding. Static noise reduction is built into phono preamps and analog tape decks
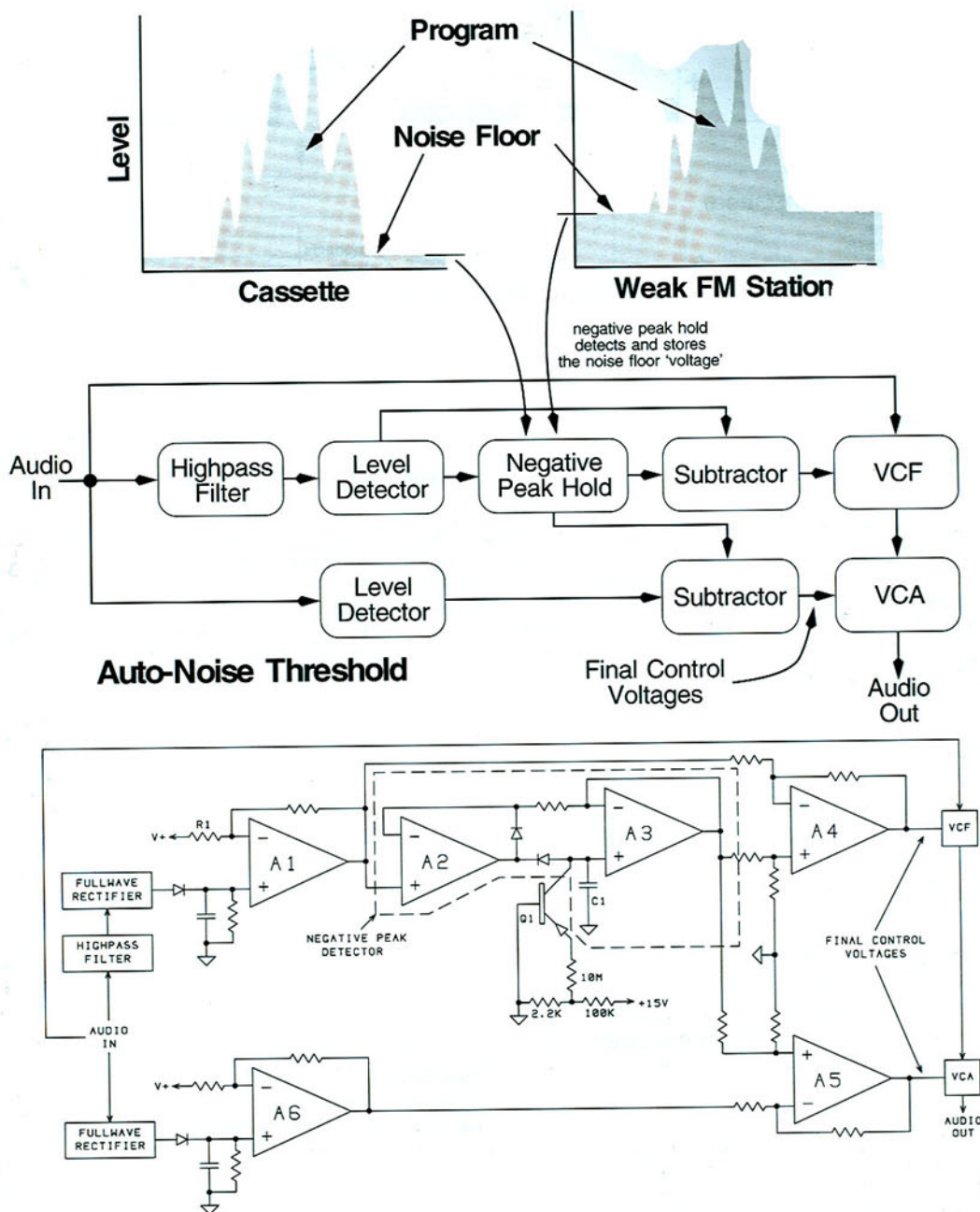
Fig. A131. Top graphs illustrate the concept of the noise floor, the minimum noise present in an audio source in the absence of a signal. Middle diagram illustrates function of the auto-noise threshold in the SSM2000. The circuit assumes that the minimum amount of hiss detected over a period of time registers the noise floor. The circuit derives a voltage proportional to the hiss and stores it in a peak detector. The circuit uses this changeable value instead of a manually set threshold to derive the final control voltages that open the lowpass filter and downward expander on the chip.

Schematic explains the circuit in greater detail. Audio input feeds two level detectors. A sharp highpass filter precedes the VCF detector to keep it from sensing bass and midrange. System is based on the assumption that the lowest level of hiss represents the noise floor. In the absence of an input signal, the output of A1 is driven to its negative limit by positive voltage applied through R1. Signal or hiss generates a positive voltage that keeps A1's output above the negative limit. The maximum negative excursion of A1's output is detected and held in a negative peak detector. This voltage is subtracted from raw VCF and VCA level detector signals to yield the final control voltages. The peak detector ignores any voltage more positive than the one stored in C1, but instantly adopts a voltage more negative than the one stored in C1. Thus, the negative peak detector tracks down, but not up. It can sense a lower noise threshold, but not a higher one. Adaptation to a higher noise threshold is allowed by constant-current source Q1, which provides some tens of nanoamps. This neutralizes the negative charge in C1 over some tens of seconds, raising the threshold; but the threshold instantly resets to a lower value on receipt of a potential more negative than the one stored in C1. The circuit's upper and lower limits prevent threshold from migrating to useless extremes.

Quasi-Companding

Pre-emphasis/De-emphasis

Companding

Fig. A132. Top—Quasi-companding involves boosting the signal as much as possible, as quietly as possible prior to processing; reducing it by as much as possible after processing. Noise introduced by processing is reduced by the amount of gain reduction. System headroom limits the amount of noise reduction. Middle—Treble pre-emphasis applied to signal prior to processing reduces system noise by the amount of de-emphasis applied after processing. Phonograph recordings and analog tape incorporate standardized emphasis/de-emphasis. Bottom—True companding systems usually include treble pre-emphasis. Process is more complicated than the first two, but eliminates analog tape noise, or noise produced by the effect when used internally in a stomp box.

as the RIAA phono and NAB tape EQ curves, respectively (Fig. A133).

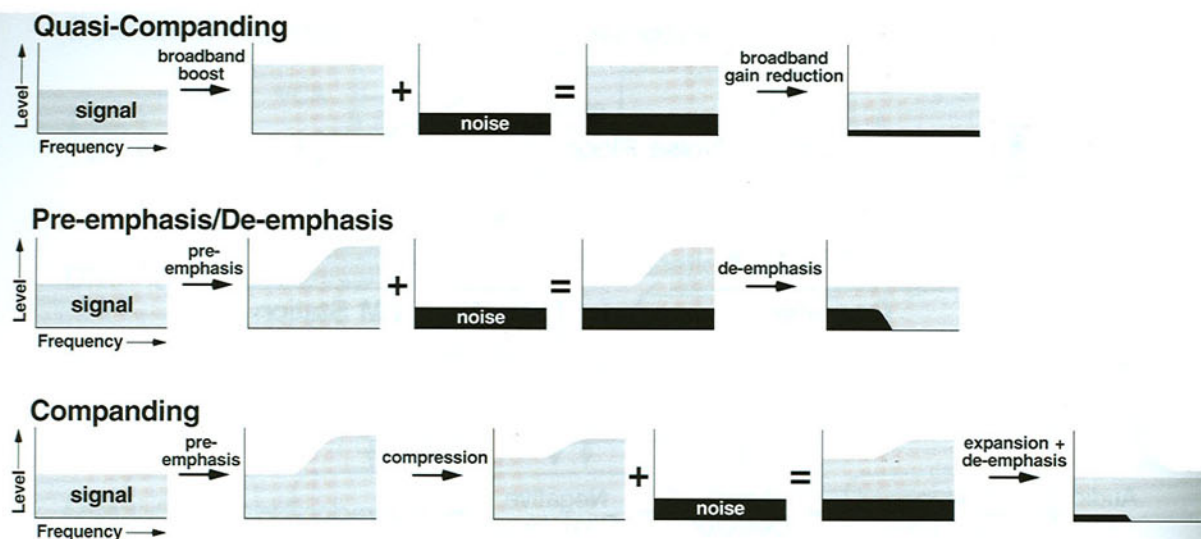As a nonreactive system, static noise reduction involves no thresholds, suffers no attack/decay lag. It uses simple circuitry and avoids the tracking hurdles of companding. Installation can be as easy as adding two caps and two resistors to a stock effect. On the downside, the technique pushes headroom to the limit. A 9V box whose preamp output peaks at 2V might get away with boosting the signal by a factor of 3, achieving ~9 dB of noise reduction. A ±15V box could boost peak output to $24V_{p-p}$ pretty comfortably, reducing noise nearly 22 dB.

Dynamic Systems

Probably the most pervasive complementary dynamic system, Dolby B first appeared in cassette decks in 1971. The system encodes using level-dependent treble emphasis, decodes by level-dependent de-emphasis. Maximum noise reduction approaches 10 dB. The emphasis applies only to medium- and low-level signals, leaving loud passages alone. Dolby C arrived in 1977 and extended noise reduction to 20 dB. Both systems exist in integrated circuits available only to licensees of the product. Dolby A is a more powerful, multi-band system that was used to make studio recordings and film soundtracks in the sixties. The recent Dolby S system varies emphasis according to the spectral content of the program.

Companding describes the process of compressing a signal prior to recording or transmission, expanding it upon playback or reception. Two benefits attend the function. First, compression cuts the signal's dB amplitude by half. This lets media of limited dynamic range accommodate the signal. Second, whatever noise the compressed signal picks up in transit ex-
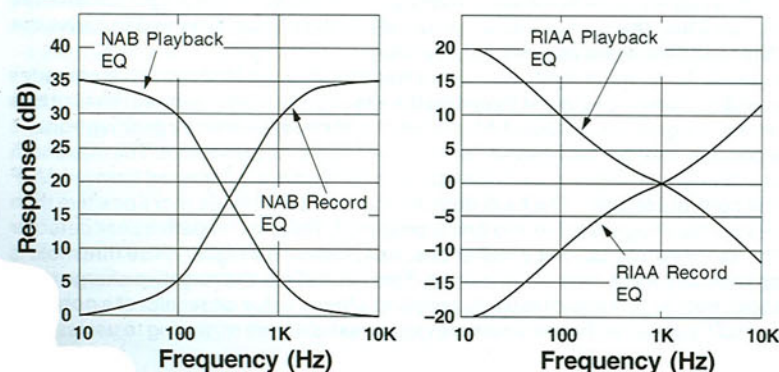
Fig. A133. Approximations of the two most common static emphasis/de-emphasis EQ systems. Left graph shows extremely sharp treble pre-emphasis applied to analog tape recordings; matching de-emphasis applied on playback. Different EQ curves are used for 1.875–3.75 ips and 7.5–15 ips recordings. Tape noise reduction systems, such as Dolby® or dbx®, apply on top of this EQ. Right graph shows RIAA emphasis/de-emphasis applied to vinyl phonograph records. Twenty dB of bass boost explains why phono playback is subject to rumble. Curves approximated from several sources.
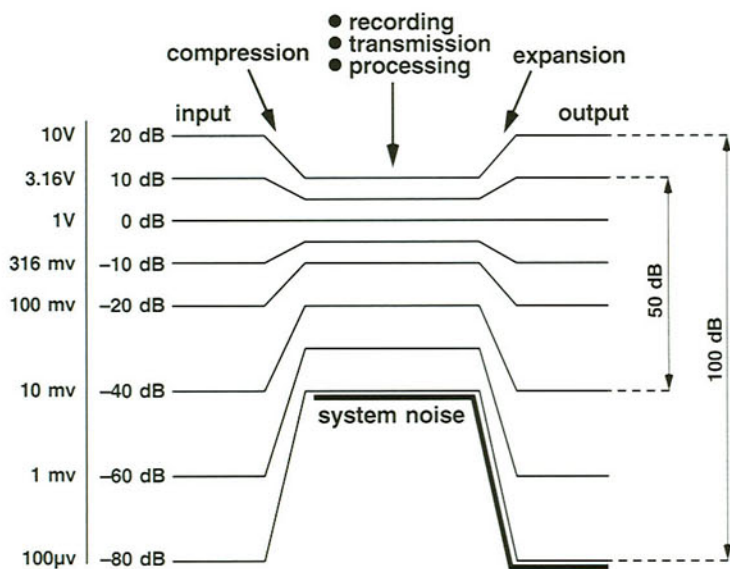
Fig. A134. Graph illustrates the dynamics of companding. In this example, 2:1 dB compression squeezes 100 dB of dynamic range into 50 dB. As a side benefit, whatever noise the signal picks up in the system expands downward, usually into inaudibility. '0 dB' is an arbitrarily set point at which compressor input = output; 1V here.
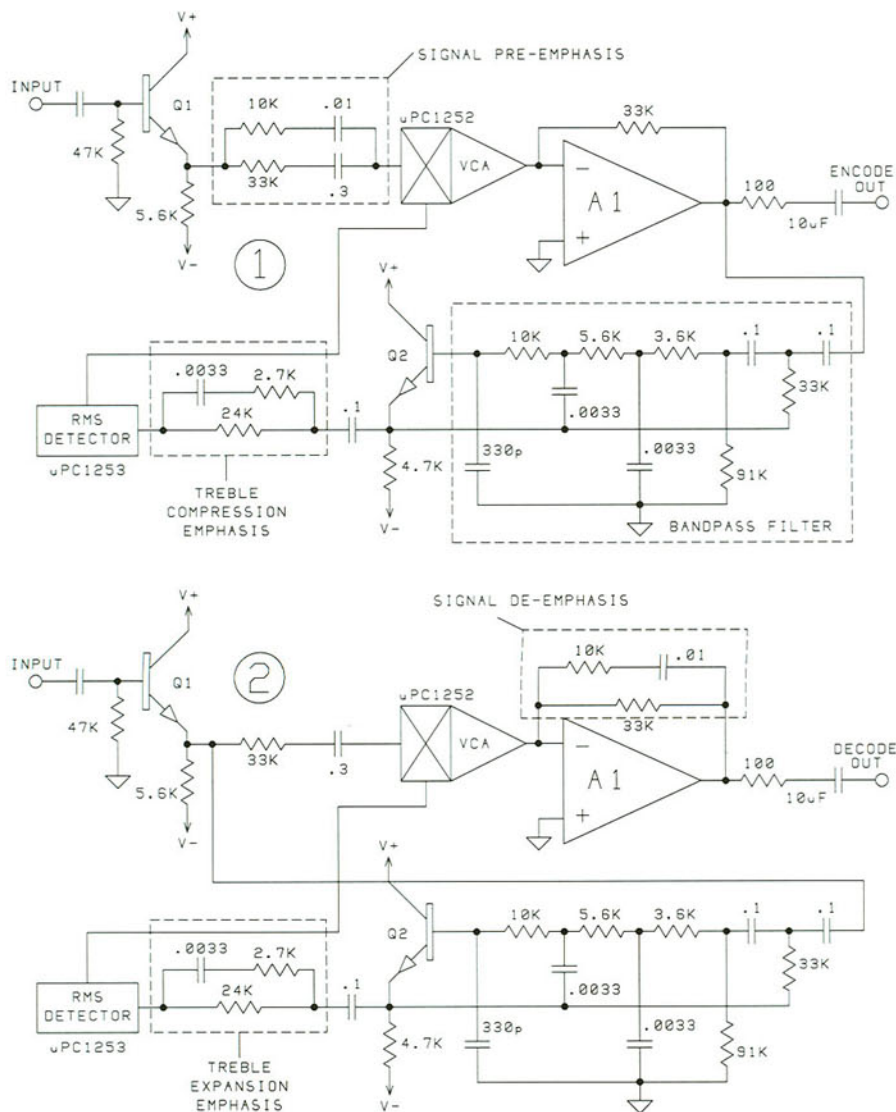
Fig. A135. Core elements of the dbx® Type I companding system. 1—Compression applies about 10 dB of treble pre-emphasis to the signal, kicking in above ~500 Hz. Treble emphasis is also applied ahead of level detector, to compress the boosted frequencies more than unboosted ones. Bandpass filter emphasizes frequencies to which the ear is most sensitive and prevents the level detector from reacting to subsonic and ultrasonic energy. The 0-dB point is about $2V_{p-p}$; control path is feedback. 2—On expansion, signal receives complementary treble de-emphasis. Treble emphasis ahead of RMS detector expands treble frequencies more than bass. Control path is parallel. Type I noise reduction is used on systems with inherently good performance, such as open-reel tape decks. Type II, used on cassette decks, applies greater treble compression to help compensate for cassettes' poor high-frequency response. The same functional blocks are present in most high-performance companding systems. Schematics do not show bias and trims needed on VCA and RMS detector, which are available as NTE1794 and NTE1795, respectively.

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Fig. A136. Basic fixed emphasis/de-emphasis noise reduction. 1—Transistor model. Emphasis corner of R2 with C1 is ~1600 Hz. R5 limits boost to 20 dB. Matching de-emphasis is applied after effect by C2-R6. 2—Functionally identical version using op amps. Corner frequency and boost limit can be changed by manipulating capacitance and resistance values. Boost limit is determined mainly by system headroom.

pands downward upon decoding, usually below the level of audibility. Most companding systems add treble pre- and de-emphasis to enhance noise reduction and prevent bass from audibly modulating treble. Net noise reduction typically exceeds 30 dB.

Successful companding systems exhibit remarkable similarity: About 10 dB of pre-emphasis kicks in above 500 Hz. Further treble emphasis ahead of the level detector causes treble to be compressed more than bass. Bandpass filtering ahead of the level detector emphasizes frequencies to which the ear is most sensitive. Compression uses a feedback control path; expansion, a parallel path. A schematic of the dbx® Type I companding system illustrates these features (Fig. A135).
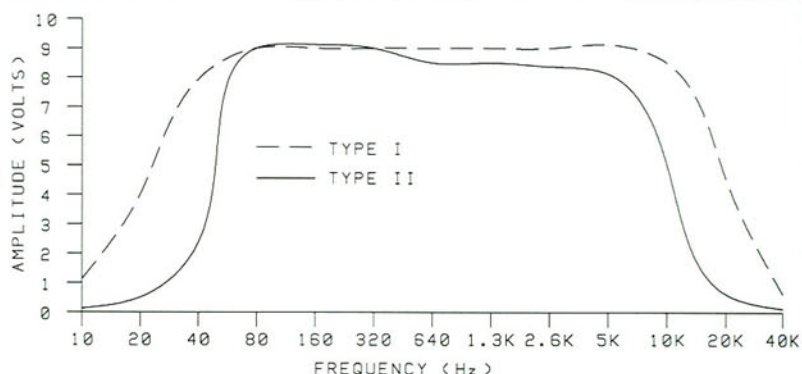
The stock NE570 compander and dbx Type I operate at a threshold of about $2V_{p-p}$. This *0-dB point*, at which compressor input equals output, exceeds the maximum level of most recording and transmission media. Thus, companding systems apply mainly upward compression, downward expansion. The 0-dB point is easily altered to meet the needs of other systems.

Companding is a portable process in that compres-

sor and expander function autonomously. A signal stored on tape can later be expanded on another deck equipped with the same companding system. Even so, some degree of mistracking always exists. Stomp boxes present the opportunity for simultaneous or *internal* companding, 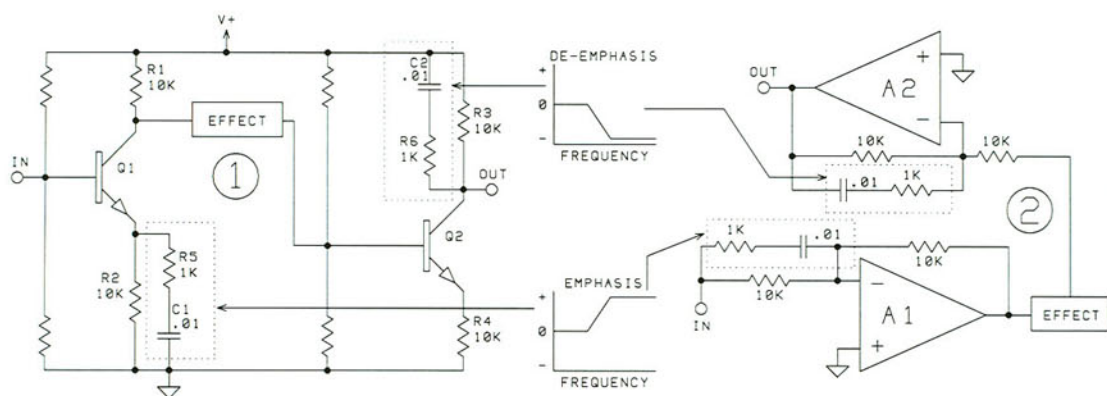in which the compressor's control voltage also drives the expander. This simplifies the circuit and practically eradicates mistracking, but it cannot be used with effects that involve delay, or that greatly alter amplitude or spectral profile. Those latter cases require the expander to run off its own level detector. Other problems associated with companding relate to upward compression. The compressor applies maximum gain to very low-level signals. The next transient may clip unless the compressor attacks instantly; but fast attack heightens ripple in the control feed, raising bass distortion. Luckily, expansion reverses this distortion. The better the tracking between compression and expansion, the greater the distortion the system can neutralize.

In trade for a few quirks companding offers greater noise reduction than other approaches. Its proven circuitry makes a cheap and practical solution for rad but rowdy effects.

Fig. A137. I/O curves of bandpass filters used in dbx noise reduction systems to condition the signal feeding the level detector. The Type I filter, whose schematic appears in Fig. A135, has fairly wide response, rolling off only at the extremes of bass and treble. Type I is meant for recording media whose response covers the full audio spectrum. Type II was used on cassette decks. Its response rolls off sharply below 60 Hz and above 10 KHz. Output of both networks plotted relative to constant $10V_{p-p}$ input.
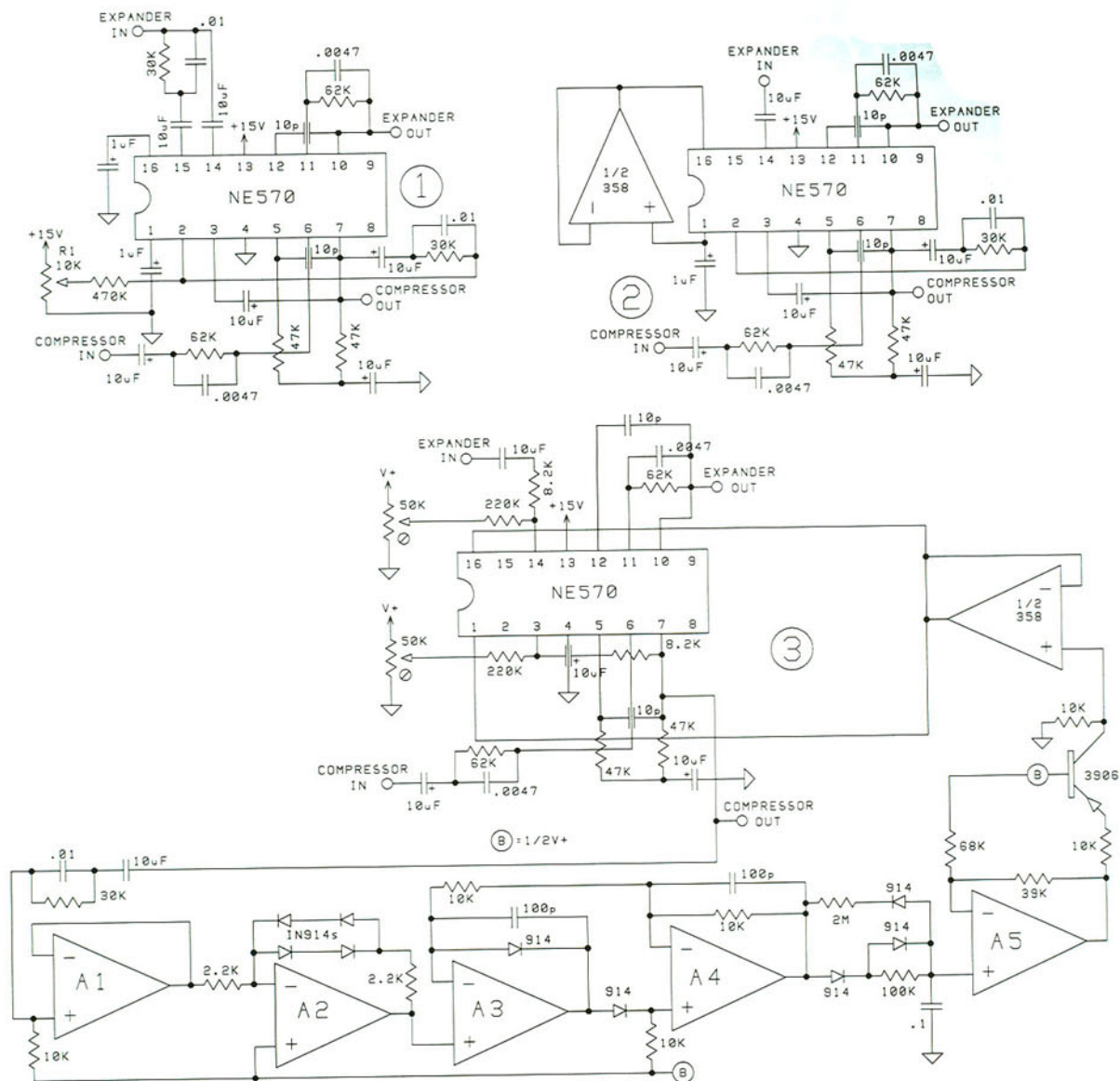
Fig. A138. 1—Truncated version of Signetics' Databook 570-based companding system; 62K/0.0047µF networks apply treble emphasis on compression, matching de-emphasis on expansion; 30K/0.01µF networks cause treble to be compressed and expanded more than bass. R1 lets user apply variable mistracking, such that upward compression wanes at low levels; corresponding expansion does not wane, automatically gating the signal; technique is described further in Ref. 45. 2—Companding system identical to #1, but control voltage generated by compressor also drives expander. Impedances at pins 1 & 16 are high enough that loading effects emerge if the two tie together; the LM358 buffer relieves loading. System tracking is excellent but, obviously, compression and expansion must occur simultaneously. 3—External level detector improves transient performance. Treble emphasis network feeds buffer A1, in turn feeding modified log amp A2, whose gain approximates that found in the 570's level detector. A2 output feeds fullwave rectifier A3/A4, feeding attack/decay network whose output is boosted and buffered by A5. A5's output is level-shifted in Q1, buffered by half an LM358 (or other amp whose output can swing to ground). In all three of these companding systems, gain between compressor output and expander input is assumed to be one. A level-matching buffer is necessary otherwise. The 50K trimpots can be used to minimize envelope bounce in critical applications; adjust for minimum envelope bounce with bi-level tone burst input. Trim is sensitive to variations in supply voltage, requiring local 5V regulation if the supply is not regulated.

Voltage Control Techniques

Perhaps half of all stomp-box effects demand that functions normally varied by changing resistance vary with voltage. Tremolo, vibrato, phase, sibilance reduction, compression, and expansion prove awkward otherwise. The process occupies static, dynamic, and rhythmic classes. Static devices use voltage control as a matter of convenience, or to yoke several segments to a single feed. Dynamically controlled effects drive the VCB off a voltage proportional to the audio input. Rhythmic effects drive the VCB off an oscillator.

Voltage control occurs in dedicated chips, or in conventional circuits adapted to the task. Getting the most from each approach demands attention to detail. Often the choice is dictated by how fast the effect needs to respond, and how much feedthrough and distortion it will tolerate.

Approaches to Voltage Control

Optocouplers

One of the oldest and most versatile approaches to voltage control consists of a light source potted with a light-responsive element, a combination called an *optocoupler*. Optocouplers come in many flavors. The one that has lived in guitar amps since Leo Fender's heyday uses a photocell as a light-dependent resistor (LDR). LDRs mate to incandescent bulbs, neon lamps, or LEDs. Cadmium sulfide photocells exhibit peak sensitivity to green light, around 565 nm; cadmium selenide photocells respond more toward the red end of the spectrum. The type continues to reign as the premiere voltage-variable resistance in stomp boxes because audio sees the LDR as pure resistance, and because floating voltage-variable resistance is awkward to realize in ICs and transistors.

An LDR's range spans <200 ohms under full illumination to >20 megohms in darkness. Because few circuits need or necessarily tolerate such a wide range, common practice parallels the LDR with a resistor to limit maximum resistance, and/or inserts a series resistor to limit minimum resistance. Since replacing resistors with LDRs puts any resistance-varied function under voltage control, LDRs can be considered the universal VCB.

Wide range and negligible feedthrough should make LDRs perfect VCRs, but the breed suffers several quirks. Given a stiff control input, LDRs respond in less than 5 ms, but may take several hundred milliseconds to return to maximum resistance. Sluggish decay distorts the control envelope and limits depth in

Abbreviations:

BPT	bipolar transistor
FET	field effect transistor
HPF	highpass filter
I–V	current-to-voltage
LDR	light-dependent resistor
LED	light-emitting diode
LPF	lowpass filter
BPF	bandpass filter
BRF	band reject (notch) filter
OTA	operational transconductance amplifier
SVF	state variable filter
V–I	voltage-to-current
VCB	voltage controlled block, generic for any of:
VCA	voltage controlled amplifier
VCAt	voltage controlled attack
VCD	voltage controlled delay
VCDy	voltage controlled decay
VCE	voltage controlled equalizer
VCF	voltage controlled filter
VCI	voltage controlled inductor
VCM	voltage controlled mixer
VCO	voltage controlled oscillator
VCP	voltage controlled panner
VCPh	voltage controlled phase
VCR	voltage controlled resistor
VCV	voltage controlled voltage

rhythmic effects above 1 Hz. It also integrates the control feed such that some LDRs use the raw output of a fullwave rectifier as a control feed. External integration of the control feed reduces distortion on signals below 60 Hz. Other drawbacks include high cost, limited selection, scarcity, and poor intersample matching and tracking. Battery life falls if the effect spends much time with the LED fully lit due to the 10 ma or more required.

Optocouplers can be had as the Clairex CLM6000 and the Vactec VTL2C2; each currently retails for about $3. The two are functionally interchangeable but come in distinct, irregular cases.

An LDR-based optocoupler is easily built by mating a light source to one or more photocells. Durability demands strain relief for the leads, achieved by soldering the parts to a small circuit board or potting them in a light-tight tube. The builder should expect higher ON-resistance from this jury-rigged device compared to commercial pieces, but can compensate by changing resistances elsewhere in the circuit. (A selection of photocells having different intrinsic attack and decay recently became available [see Parts Sources], making dynamic response somewhat selectable in homemade optocouplers.)

Phototransistor-based optocouplers attack and decay in a few microseconds. They're cheap and widely available; bipolar examples include 4N25 and 4N36; H11F3 is an FET type. They work fine as saturated switches for audio, or for continuous variation of direct current. They do not show AC signals a pure resistance, which results in distortion when they're used as variable resistors in the signal path. BPT-based optocouplers exhibit the same polarity sensitivity as conventional BPTs acting under direct voltage control; FET-based optocouplers conduct bidirectionally. Optocouplers broaden the voltage control range over that readily achieved using conventional transistors, especially FETs, which tend to pinch off abruptly.

LDR and phototransistor optocouplers use identical drive circuits. Many common op amps will supply the current needed for full illumination. Low-power op amps may have to drive a transistor used as a current controller. Series wiring gives best results when multiple optocouplers run off a single feed. Wiring LEDs in parallel usually results in most of the current flowing through the LED having the least resistance. If parallel wiring proves necessary—say, because the drop across the LEDs wired in series exceeds the available voltage—a resistor in series with each LED helps equalize the distribution of current.

Conventional Transistors in Voltage Control
Circuits simplify, cost falls, and options expand if the effect tolerates the distortion that often accompanies use of transistors as though they were VCRs in the signal path. Transistors make good audio switches, whose uses are discussed in a separate Appendix. The occasional circuit gives low enough distortion to enable transistors to act as signal-path VCRs, one example being FETs used as shunt elements to tune a state variable filter (Fig. A148–4).

Operational Transconductance Amplifiers
While tubes and transistors qualify as transconductance devices, it proves handy to limit this category to operational transconductance amplifiers (OTAs). Transconductance, simply, means the flow of current through an electrical channel, per unit of current or voltage affecting that channel. The trait is what enables transistors and tubes to amplify. OTAs behave much as conventional op amps, but provide access to a gain-varying transconductance channel. The commonest and cheapest types are LM3080 single and LM13600 dual. Both chips process current; most circuits condition the input by a network that converts voltage to current. Conversion of current back into voltage takes place at the output.

OTAs place a remarkable array of functions under voltage control. The LM13600 data sheet presents more than two dozen examples (Ref. 19). Some of those useful in stomp boxes are detailed below.

Special-Function Integrated Circuits
The need for voltage control of certain functions led to series of chips built to perform those functions. Several types are discussed below.

Voltage Control Functions & Methods

Voltage Controlled Resistance
  LDR-based optocouplers
  phototransistor-based optocouplers
    bipolar
    FET
  OTAs (LM13600/13700)
    single-ended
    floating
  Transistors (distortion)
    FET
    MOSFET
    BPT
Voltage Controlled Amplification
  Dedicated Chips
    NE570/571 configured as VCA
    SSM2120 (w/level detector; current output)
    SSM2122 (current output)
    SSM2018 (voltage output)
    SSM2118 (current output)
  OTAs Configured as VCAs
    LM3080
    LM13600/13700
  Discrete-Component VCAs
  Op-Amps using external gain-control elements
    FETs
    optocouplers
      transistor-based
      LDR-based
  Analog Multipliers
    AD534, AD634, etc.
  Passive (Simple Divider)
    optocoupler-based
    transistor-based
Voltage Controlled Phase
  substitute VCRs for fixed resistors
    optocoupler
      LDR
      FET (distortion)
    OTA
Voltage Controlled Panning
  2 VCAs mated to inverse control voltage
  SSM2018
Voltage Controlled Mixing
  same as voltage controlled panning, but both
    VCAs feed summing amp
Voltage Controlled Filters
  conventional active filter/EQ substituting
    VCR for fixed filter resistances
  substitute analog multiplier for fixed resistor
  OTA-based filters
  switched-capacitor filters (not discussed)
Voltage Controlled Inductance
  simulated using op amps
Voltage Controlled Oscillators
  dedicated function generators
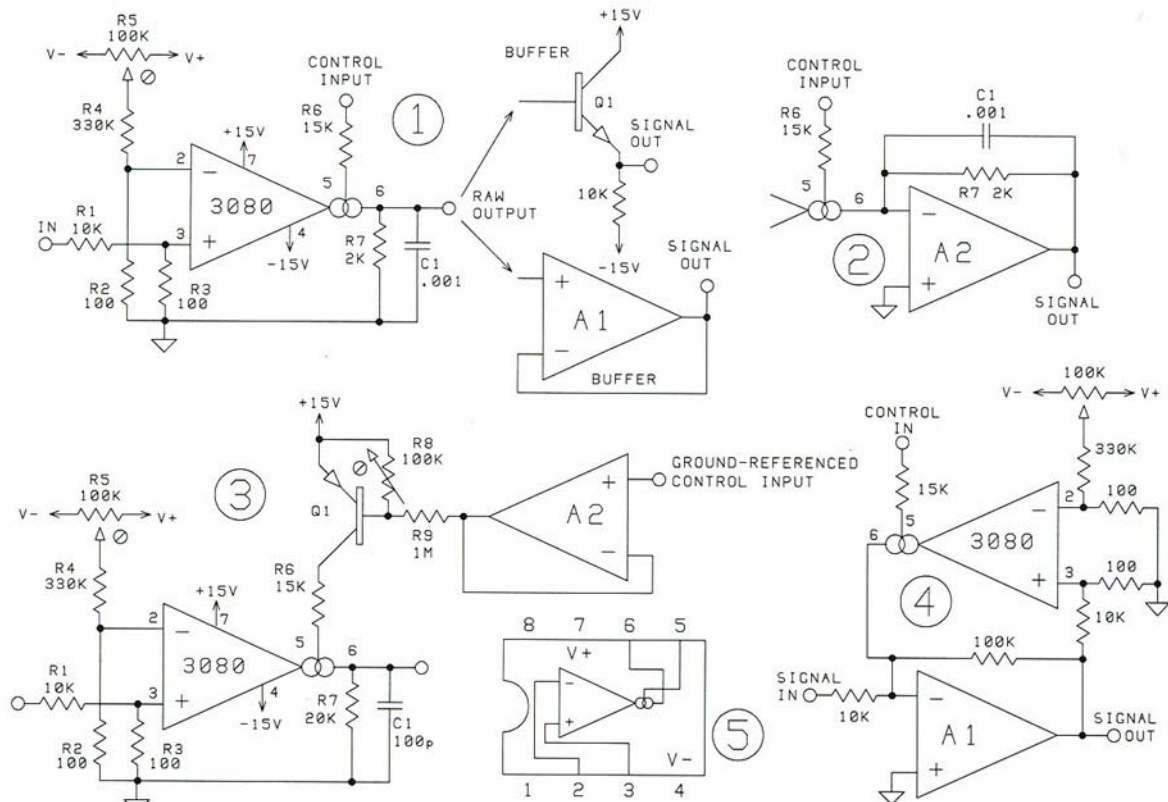 conventional VCO mated to VCR

Fig. A139. Getting the most from an LM3080 used as a stomp-box VCA. 1—Chip is configured as a VCA whose gain varies 0–1 over the control range −14.4V to +15V. Divider R1-R2 drops signal voltage by a factor of 101; equals conversion to current on which OTA acts. Signal emerges from pin 6 as a current, dropping a voltage across R7. C1 is needed to limit bandwidth; the 3080 is a high-speed device that functions well into the megahertz. Raw output is typically buffered, as by emitter follower Q1 or voltage follower A1. 2—A common alternative ties pin 6 to an op amp configured as a current-to-voltage converter which also serves as a buffer; now R7 and C1 appear in the feedback loop. The op-amp I–V converter inverts; to retain a net noninverting signal path, the input would have to shift to the 3080's inverting input. Signal can tie to either input, as can feedthrough trim from R4-R5. Gain is directly proportional to value of I–V conversion resistance. Raising the value of R7 raises the maximum gain, but degrades noise performance because a high noise current exists at the OTA's output; this drops a noise voltage across the I–V resistance. The 1:101 input voltage reduction lets this VCA respond cleanly up to about 3V$_{p-p}$. Above that point, signal becomes squashed, reaching a diode-like limit at 10V$_{p-p}$. This form of distortion is not always unwanted in a pedal. The distortion threshold can be raised by raising the value of R1, or lowering the values of R2-R3, which lower the input voltage level. This gain loss must be made up by raising the value of R7, increasing noise as previously described. Control voltage is applied to pin 5 through R5. Because pin 5 exists close to V−, the maximum differential across R5 approaches [15V − (−15)] = 30V; (30,000 mv ÷ 15,000 ohms) = 2 milliamps, the maximum allowable into pin 5; exceeding this value could damage the chip. If supply were changed to the ±7.5V typical of 9V nicads, R5 scales proportionally, to ~7.5K, to keep the same control range. 3—One means to convert 3080 VCA to a ground-referenced VCA: ground the input of A2; trim R9 to give unity gain of VCA. Note that R7 has been increased to 20K; this VCA can vary gain from 0 to 10, suiting, say, a dual-mode sustainer. Negative control voltage to point A turns Q1 on harder, increasing gain above 1; positive control voltage tends to turn Q1 off, decreasing gain. Response is nonlinear, but a feedback control loop compensates for nonlinearity. This allows the 3080 to use 2120-type (ground-referenced) level detectors. 4—LM3080 controls gain of op amp A1 by acting as a voltage controlled feedback path. A1 has static gain of 10, set by ratio of 100K/10K. When the 3080 VCA is off, gain is 10. As VCA turns on, signal fed back to inverting input increases negative feedback, lowering gain. 5—Pinout of LM3080.

Voltage Controlled Function Blocks

Voltage Controlled Amplifiers

A prevalent need met by a profusion of hardware makes this the commonest VCB, realized in a host of dedicated chips, and almost as easily in conventional circuits adapted to the task. Of several dozen options, the text will consider four chips, the LM3080 and LM13600 OTAs, and the NE570 and SSM2120 dynamic controllers; plus VCAs built from LDR-based optocouplers. These parts share reasonable cost, availability, high performance, and a long and fruitful heritage in the pedal scene.

Getting the most out of chip VCAs is something of an art. The builder will find them noisier than conventional op amps supplying the same gain. In the stompbox domain, all four chips peak when configured to vary gain from zero to one. This happens to be the most useful arrangement, serving compressors, downward expanders, and tremolos; and the one al-
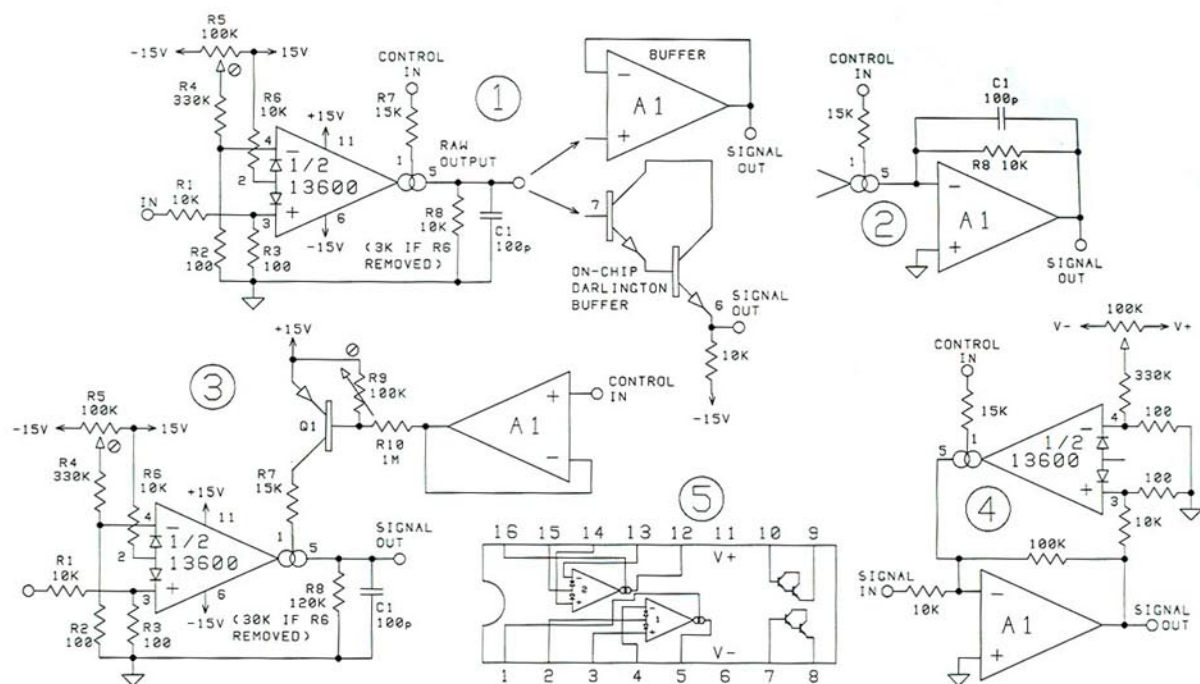
Fig A140. The LM13600 (or 13700) functions almost identically to the LM3080, with two exceptions. Schematic resembles that for a 3080 VCA, but a new pin has been added for diode bias (pins 2 & 15). This reduces the input-level–dependent distortion; the 13600 acts pretty much like a 3080 when diode bias is omitted. 1—This VCA varies gain from 0–1 over the control range −14.5V to +15V, but the I–V conversion resistance R8 is five times larger than that needed for the 3080. This results in greater noise and greater control feedthrough. A second difference is the presence of on-chip Darlington transistor buffers. The collectors are tied to V+ inside the chip. Darlington emitter exists two diode drops below ground; output of op-amp voltage follower exists close to ground. 2—Use of op amp I–V converter, which inverts. Output exists close to ground. 3—VCA whose gain range spans 0 to ~8; normalizes to a ground-referenced control feed as for the 3080, but noise of this system will be significant due to very high value of R8. By omitting R6, R8 can drop to 30K, giving a VCA functionally interchangeable with a 3080. 4—Using 13600 to moderate gain of an op amp by providing voltage controlled feedback. 5—Pinout of LM13600/13700.

lowing least feedthrough and noise.

LM3080

Introduced early in the seventies, the 3080 gave hobbyists one of the first chip VCAs. The OTA manipulates current: input voltage is converted to current by resistive dividers; the chip responds to control current, obtained by passing the control voltage through a resistor; and the output is furnished as a current, converted back to voltage in the drop across a resistor.

Fig. A139–1 shows a typical VCA whose gain varies 0–1 over the control range −14.5V to +15V. If the control range shrank or expanded, R7 would have to change proportionately to avoid a shift in response. Feedthrough trim is sensitive to the supply, such that applications needing trimmed feedthrough require a regulated supply or a stable trim reference.

The I–V resistance should be kept as small as possible, because the OTA output terminal sources a large noise current whose value rises with control current. An output resistance greater than ~5K results in audible noise in stomp boxes.

Some circuits allow use of the VCA as if it were a VCR. Examples include Envelo-Matic's SVF, and circuits where the OTA acts as a voltage controlled channel in the feedback loop of an op amp (Fig. A139–4).

An op-amp I–V block inverts, so a net noninverting path must use the OTA's inverting input.

Distortion rises rapidly as the voltage at the OTA input terminal exceeds ~30 mv$_{p-p}$. Distortion manifests as rounding-off in the manner of triode overload, gradually flattening to resemble diode clipping. A DC bias applied to one input skews the symmetry of the squashing. These features make OTAs useful as distortion generators.

LM13600

This dual OTA differs from the 3080 in two respects. First, each OTA contains a diode bias port (pin 2 or pin 15) whose use greatly improves linearity. However, reduced distortion incurs reduced gain, requiring a larger I–V conversion resistance, which magnifies feedthrough and noise. By leaving the diode bias pins open, the OTAs duplicate 3080-types. Also, each channel contains an NPN Darlington transistor whose collector ties to V+ inside the chip. The manufacturer refers to the transistors as "controlled-impedance
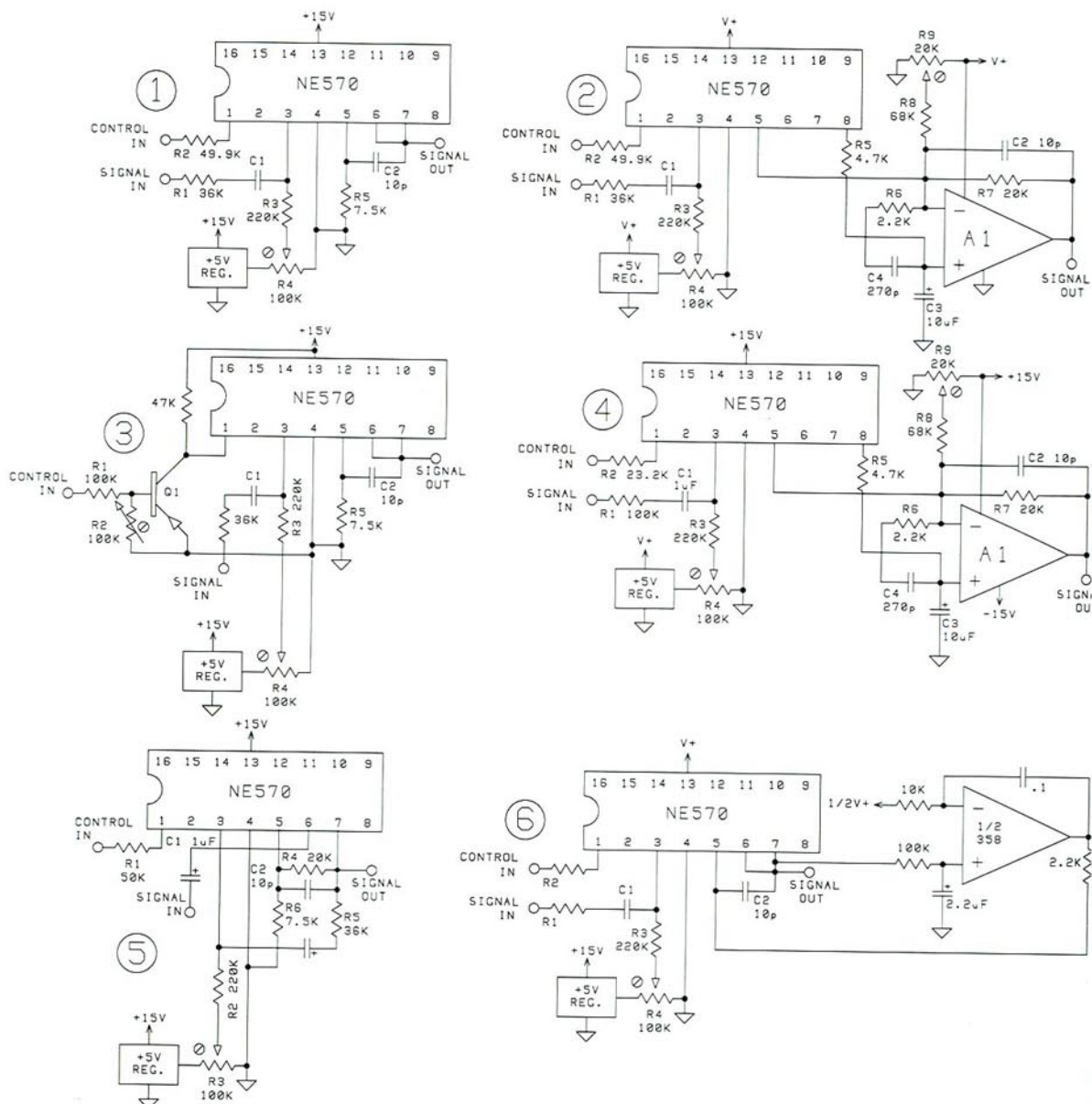
Fig A141. 1—NE570 configured as VCA whose gain varies 0–1 over the control range 0.5–14.5V, which corresponds to the output swing of a 3317X-type op amp. The value of R1 is chosen to allow the chip to handle the 15V system's headroom limit of about 12V$_{p-p}$. If control voltage range shifts, R2 may also have to be changed to avoid shifting the 570's voltage vs. gain response. Feedthrough trim approaches nil, the best of all chip VCAs discussed in this Appendix, but is exquisitely sensitive to voltage, so trim network runs off locally regulated 5V supply. R5 is needed to keep output DC offset near ½V+ for maximum headroom. If supply changes, R5 will require change also. C2 is usually needed for stability and low noise. Low slew rate of internal op amp limits clean headroom to about 8V$_{p-p}$ @ 20 KHz. 2—Basically the same setup, but using an external op amp. R5 supplies bias voltage for A1; R6 & C4 are necessary for stability. R9 trims DC offset for supply other than 15V. 3— One means to refer control to ½V+; tie end of R1 to ½V+; trim R2 to give unity gain. A negative voltage (i.e., less than ½V+) into R1 turns Q1 further ON, reducing gain; positive voltage turns Q2 OFF, increasing gain. 4—Similar to #2, but A1 runs off ±15V while 570 runs off 15V single supply; input signal range up to 30V$_{p-p}$; output swings nearly 28V$_{p-p}$; control range GND to +15V for gain of 0–1. Trim R9 for maximum headroom. Using 5532-type for A1, this VCA exhibits very low distortion from 20 Hz to 20 KHz. 5—In this VCA, the gain cell has been wired in the feedback loop of the internal op amp; this inverts the control response in that positive voltage into pin 1 reduces gain. Feedthrough trim is essentially as good as that for circuit #1. 6— Similar to circuit #1, but 570's output offset is now servo-controlled, eliminating need for DC offset trim. Keeps output at ½V+ over the supply range 6–22V. Changing supply voltage usually changes appropriate values of R1 & R2, however.
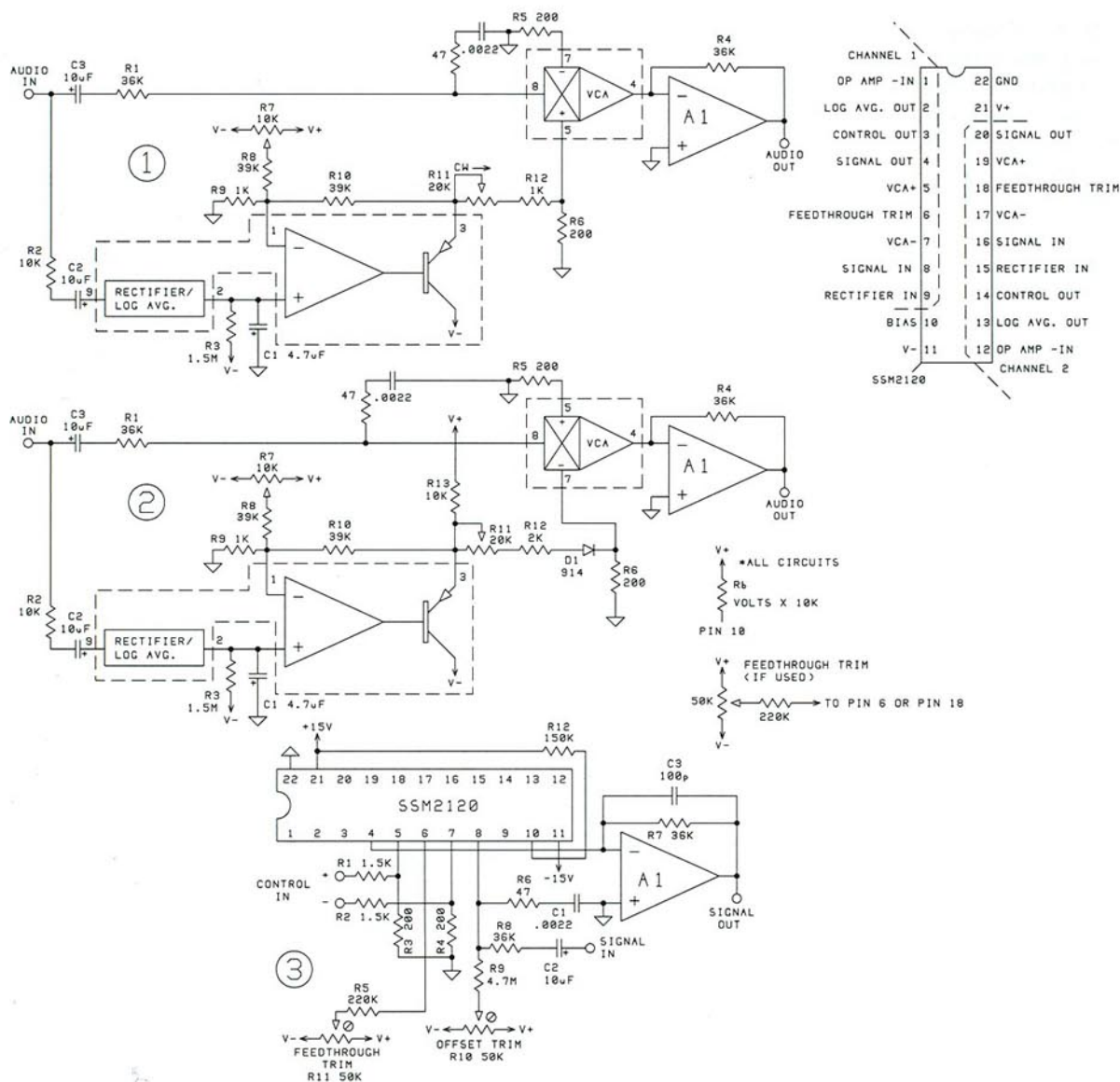
Fig. A142. Applying the SSM2120. Components inside dotted lines are inside the chip; only channel 1 is described. 1—Signal Path: Signal input couples through C3 to R1, whose purpose is to reduce current level, because VCA controls current. (The 47-ohm resistor & 0.0022μF cap form a snubber, vital to stable operation; use the snubber in all applications.) VCA output, as a current, comes off pin 4, ties to external op amp configured as current-to-voltage converter. Matching values of R1 & R4 give unity gain when both control ports, pins 5 & 7, exist at ground potential. Control Path: Audio couples through R2-C2 to rectifier input; value of R2 is chosen to suit expected input range; 10K suits up to 30V$_{p-p}$. Note polarity of C2; a positive DC offset exists at rectifier inputs. Raw log amp output is available at pin 2, which requires R3 for bias; value of R3 = negative supply in volts × 100K; the ±15V supply specified dictates 1.5M; ±6V would dictate 600K; etc. C1 is the averaging cap, controlling attack and, especially, decay. Averaged control voltage ties to noninverting input of internal op amp, configured for gain of 40 by R9-10. Amplified control voltage is available at pin 3. With no signal input, the voltage at pin 3 is negative; audio input generates voltage that can rise as high as ground, but not above. R7 varies the DC offset present at pin 3. Both VCA control ports are biased to ground through 200-ohm resistors. Control voltage couples to pin 5 through R11-12. When no signal is present, control voltage swings negative, causing gain reduction whose degree depends on threshold setting and setting of R11. Action describes a downward expander. 2—Circuit similar to #1, with three changes. First, a pull-up resistor, R13 has been added. This enables control voltage to swing positive and negative. Second, the control voltage now ties to pin 7, the '−' port; positive voltage into this port causes gain reduction. Finally, addition of D1 lets only positive voltage into the control port. Result is a downward compressor. Reversal of D1 gives upward compression; removal of D1 gives dual-mode compression. Control path is parallel. ALL 2120 CIRCUITS require a bias resistor on pin 10, tied to V+. Value = positive supply in volts × 10K; ±15V supply = 150K; ±10V supply = 100K; etc. Omit control feedthrough trim unless circuit needs it. A tremolo needs trim, most compressors do not. 3—Pure VCA based on 2120. Supplies gain reduction or boost, depending on control voltage polarity and VCA control pin selected. Net signal path is noninverting. VCA driven to >20 dB of gain may develop a large DC offset at signal output pin, which can be nulled by offset trim network tied to pin 8.

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buffers." They're normally configured as emitter followers that buffer the output I–V conversion. The buffer's emitter exists at two diode drops below ground, or below ½V+ in single-supply systems. This offset may become significant in headroom-critical applications.

Circuits based on this OTA resemble those for the 3080, but having two buffered OTAs in one package allows easier realization of a greater variety of functions.

Breadboarding OTAs demands care, for chip damage may result if more than 2 ma enters the control port or the diode bias port. Before applying a control voltage, always place a resistance in series with pins 1, 2, 15, & 16 in the 13600; pin 5 in the 3080. The value of the resistance should be such as to limit maximum current into the terminal to ~2 ma.

NE570

Signetics designed their 570 for long-distance telephone companding circuits in the mid-seventies, but the chip found a home in plenty of pedals. Rumors of its demise are premature, for it remains widely available, inexpensive, and capable of feats not apparent
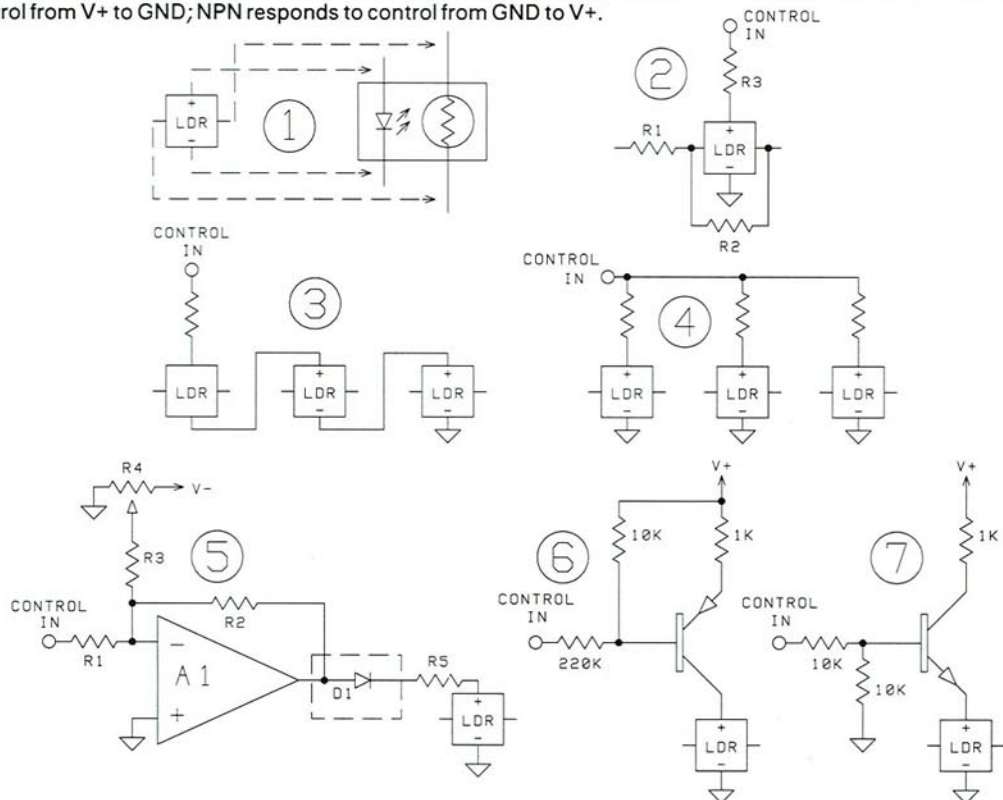
from databook schematics.

Each channel contains a current-controlled gain cell, a rectifier, and a 741-like op amp. As with OTAs, the input is converted to current; the gain cell responds to control current; and the output exists as a current, converted to voltage by an internal or external op amp configured as an I–V converter. The gain cell does not invert; the I–V op amp does, so the 570's net signal path inverts. The noninverting gain cell can be placed in the op amp's feedback loop to realize a second type of VCA.

Like other special-function ICs of the same vintage, the 570's output does not automatically bias at ½V+. Proper biasing can be achieved by selection of feedback resistance. Often the easier approach is to use a fixed, standard feedback resistance and apply a trim to the op amp's inverting input. It's also possible to servo the 570's output to track changes in the supply (Fig. A141–6).

The 570 allows several VCA configurations. The one most useful also requires the fewest parts, and gives least feedthrough and lowest noise. Trimmed feedthrough approaches nil, but is extremely sensi-

Fig A143. 1—Pinout of text's LDR-based optocoupler symbol. 2—Many applications require series resistance R1 to define minimum resistance, and/or parallel resistance R2 to limit maximum resistance. R3 limits current to LED. Value depends on available voltage and desired response. 3—LDRs are easily ganged to respond to single control voltage. Series wiring is preferred, but may be impractical in low-voltage circuits, since none of the LEDs light until voltage exceeds their aggregate forward drop. 4—If circuit demands parallel wiring, place a resistor in series with each LED. Otherwise, the bulk of current flows through one LED. 5—Many common op amps source enough current to drive the LED directly. R4 controls static state of LED; control voltage varies LED brightness with respect to this initial point. D1 may be needed if A1's output swings negative. 6 & 7—Micropower op amps can drive suitably biased transistors, which feed current to LED. PNP transistor responds to control from V+ to GND; NPN responds to control from GND to V+.
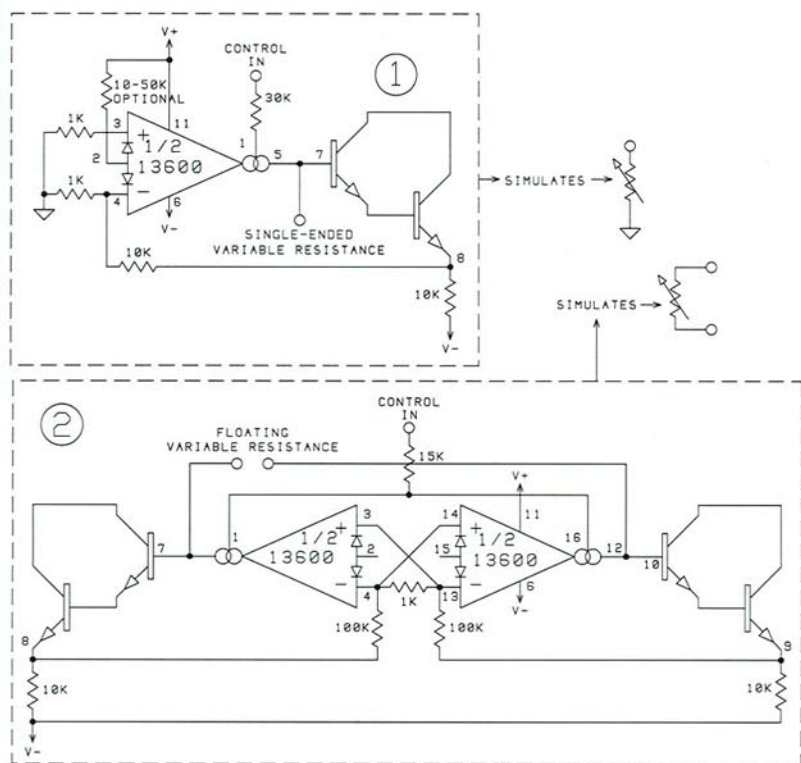
Fig A144. 1—OTA simulates variable resistance referred to ground. Diode bias resistor affects linearity of audio signal, but also affects feedthrough and simulated resistance range. Try to use largest diode bias resistor possible. 2—Two OTAs simulate floating variable resistance. Significant feedthrough occurs in both circuits. Worthwhile to test in circuit, rather than using estimating performance based on predicted resistance. Despite seeming complexity, both circuits are practical, and cheaper than LDR-based optocouplers. They enable rhythmic effects whose depth does not wane with modulation rate. Circuits affected by DC offset at simulated resistance terminals can couple through capacitor.

tive to variations in supply voltage. Unless the supply is well-regulated, the feedthrough trim network should run off a locally regulated supply.

The 570 contains a 741-like op amp, whose low slew rate manifests in distortion of 20-KHz tones above $8V_{p-p}$. This is not a problem in most stomp-box circuits. An external op amp, such as 5532/4 or 07X, takes undistorted response well past 20 KHz. The external op amp does not dramatically improve noise performance, though, because most of the noise originates in the gain cell.

The 570's rectifier is a modified log amp, giving a log curve at low levels, changing to a linear curve at high levels. This curve suits companding, which needs high gain at low levels, low gain at high levels.

The NE571 is a lower grade of the 570; the 572 and 575 are variations on the same theme, but are a lot harder to find than the 570.

Circuits shown in Fig. A141 find use in stomp-box apps, such a tremolos, compressors, and downward expanders.

Referring the Control Voltage to Ground

VCAs based on the 3080, 13600, and 570 suffer a drawback compared to 2120 types in that their control ports do not naturally exist at ground (or $\frac{1}{2}V+$ in single-supply systems) in the unity-gain state. The builder may find it handy to convert the OTAs to give unity gain at ground. This simple technique requires a PNP

transistor and a trimming network. A feedback control loop compensates partly for the nonlinearity that accompanies this approach.

SSM2120

This 22-pin DIP is currently popular on the homebrew scene, offering studio-quality noise and distortion specs. Each chip contains two channels; each channel contains a level detector and a VCA. In stock configuration, the chip realizes compression and expansion with no external active devices but an op amp.

Each VCA has two control ports with mirror-image response: positive voltage into the '+' port boosts gain, positive voltage into the '−' port reduces gain; the reverse applies to negative control voltage. This obviates inverting buffers in the level detector and allows simultaneous response to separate control feeds. The fact that the ports exist at ground in the unity-gain state makes threshold control a snap.

Tapping this chip's potential demands externalizing most of the level detector, because the stock configurations do not allow variable attack and decay. Projects in this book take the chip's raw level detector output; buffer and boost it; add a variable DC offset to vary threshold; feed the signal to variable decay and attack networks, then feed a buffered divider for ratio control.

The 2120's gain cell and the external I–V op amp both invert, giving a net noninverting signal path.
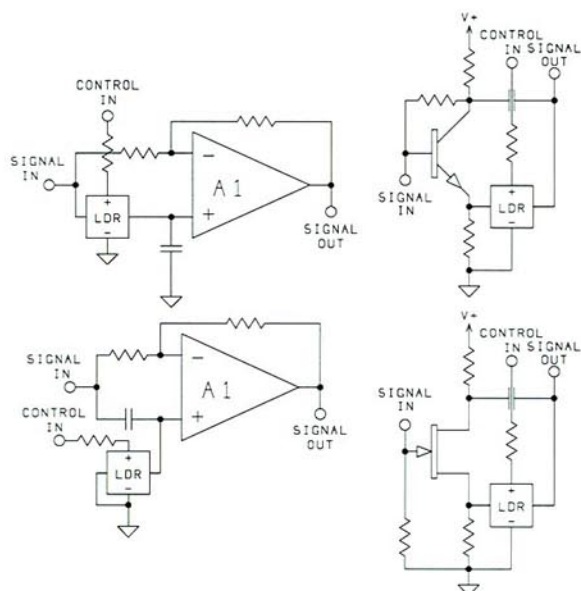
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Fig. A145. Voltage controlled phase using optocouplers as VCRs.

LDR-Based VCAs

Aside from built-in attack and decay, an LDR-based VCA's claim to fame is its suitability to gain greater than one. Chip VCAs excel at gain reduction. They're usually noisier than conventional circuits when applying boost. Fig. A147 shows a few common LDR-based VCAs.

Other VCAs

The devices just discussed typify three broad VCA categories: adaptable workhorses, upscale chips optimized for audio, and optocouplers. The stomp-box-meister will find many other VCAs and VCA-adaptable devices available. They include dedicated VCAs, such as SSM2013 and SSM2018; analog multipliers, such as AD534; and the ancient MC3340 voltage-controlled attenuator.

Voltage Controlled Filters

VCFs underlie several stomp-box functions, including sibilance reduction and some forms of single-ended noise reduction; auto-wah and some types of pedal-wah. These several functions are accessible in a handful of filter types. Figs. A148 and A149 demonstrate practical examples.

Voltage Controlled Resistance

Substitution of VCR for one or more fixed resistances puts just about any function under voltage control. If VCRs came in floating, instantly responsive blocks devoid of feedthrough, they might have obviated other approaches. LDR-based optocouplers predominate in this role because they approach floating variable resistance closer than anything else.

LDRs' chief alternative is the LM13600, which simulates single-ended or floating resistance (Fig. A144). The type suffers significant feedthrough, but offers no response lag.

In stomp boxes, VCR is commonly used to achieve VCPh and VCO (subsonic, sonic, and ultrasonic).

Voltage Controlled Panning & Mixing

The conventional approach to VCP and VCM combines separate VCAs driven by inverse control feeds; each VCA feeds an output for VCP; both VCAs feed a mixer for VCM. The recently released SSM2018 VCA offers dual inverse outputs that simplify the function (Fig. A150–2).

Effects based on VCP and VCM include stereo pantremolo, in which one channel waxes as the other wanes; tremo-mixing, in which the wet/dry mix shifts rhythmically; and level-dependent panners, in which dynamics drive a crossfader of a lone instrument, or of two separate sources.

Fig. A145B. 1—Equivalent voltage controlled circuit inside TDA1074; A3's '+' input ties to Vr, a ½V+ reference generated inside the chip. Control voltage that varies ± relative to ½V+ takes wiper of equivalent pot R1 to output of A1 or A2. Each chip contains two pair of these blocks; each pair is tied to one control path. The circuit performs a number of functions useful in stomp boxes, such as voltage controlled bass (2), voltage controlled treble (3), and voltage controlled mixing (4). The uncommitted nature of the four blocks allows adaptation to many other functions.
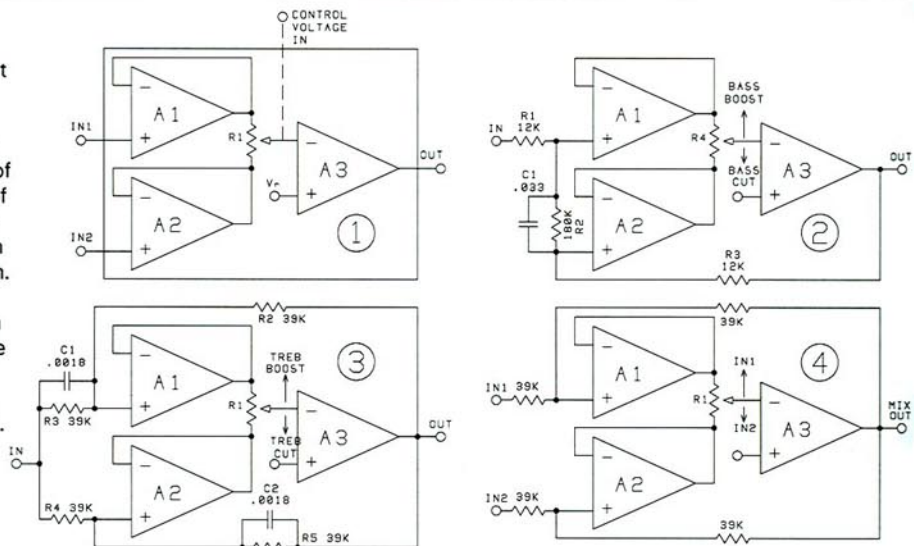
Fig. A146. Illustration of the level-dependent attack & decay built into the VTL2C2, a common LDR-based optocoupler. Schematic shows connection of LDR to positive pulse generator. Top trace 5V/div., bottom trace 1V/div.; sweep speed 200 ms/div. (2 seconds/frame). A—Control pulse of less than 2V barely gets the LDR to light; attack and decay take hundreds of milliseconds. B—A 6V pulse speeds attack; expanded sweep showed attack on the order of 5 ms; decay still takes several hundred ms; in this case, decay takes so long that LDR fails to return to fully OFF resistance between pulses. This has the effect of integrating the control feed, such that no external integration is needed down to ~60 Hz.
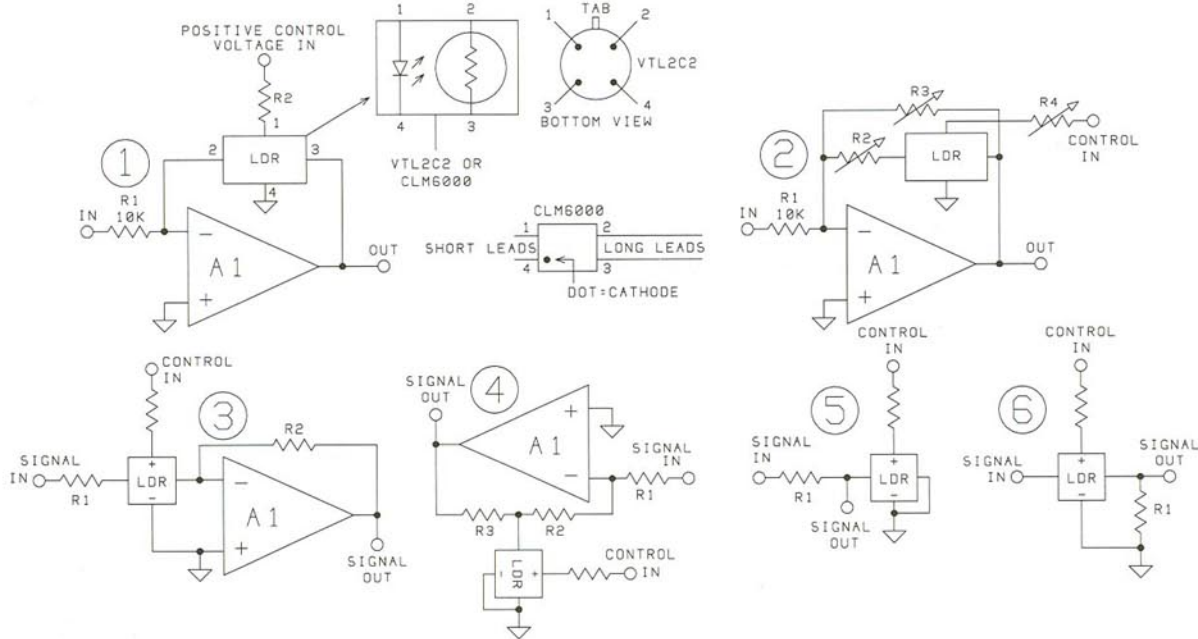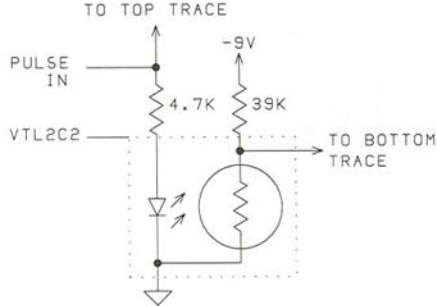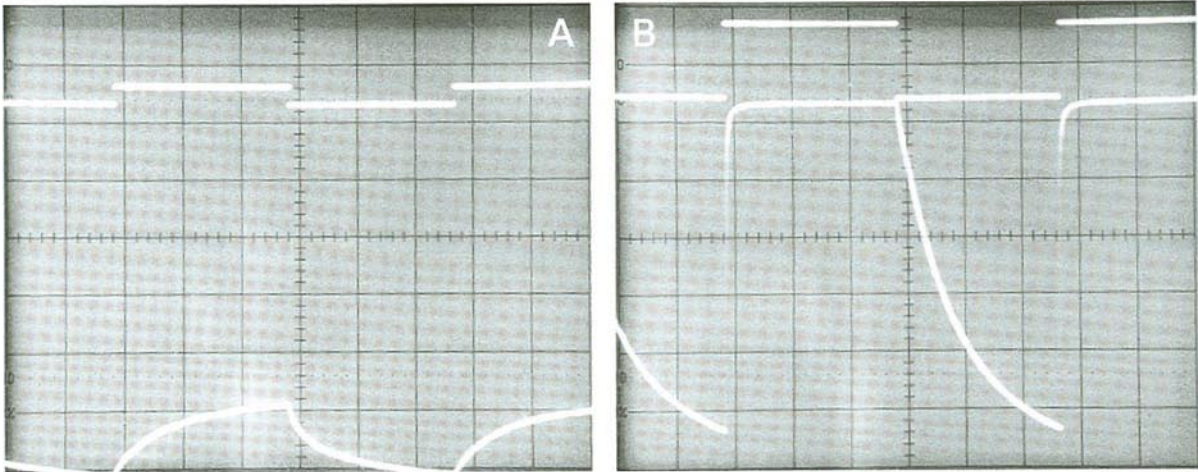

Fig. A147. 1—VCA consisting of an op amp with an LDR in the feedback loop. When LED is fully ON, LDR resistance approximates 125 ohms; gain = (125÷10K) = −38 dB. With LED fully OFF, LDR resistance exceeds 20 megohms, so high that some op amps cease to function; short of maximum LDR resistance, op amp gain can top 60 dB. 2—Practical embodiments add R3, whose value limits maximum feedback resistance and thus maximum gain; and sometimes R2, which limits minimum feedback resistance, thereby limiting maximum gain reduction. R4 alters LED sensitivity to control voltage. 3—LDR can control gain elsewhere in the signal path, in this case as part of R1; positive control voltage raises gain. 4—LDR acts as voltage controlled shunt to ground in middle of op-amp feedback loop. When LDR is OFF, it has no effect on gain. When it turns ON, it acts as a shunt to ground for all frequencies, giving the same effect as raising the values of the feedback resistors, raising gain. 5 & 6—LDR controls gain passively in simple divider circuits. In #5, LDR acts as shunt that lowers gain when LED lights; in #6, LDR acts as series element that lets more voltage through when LED lights.

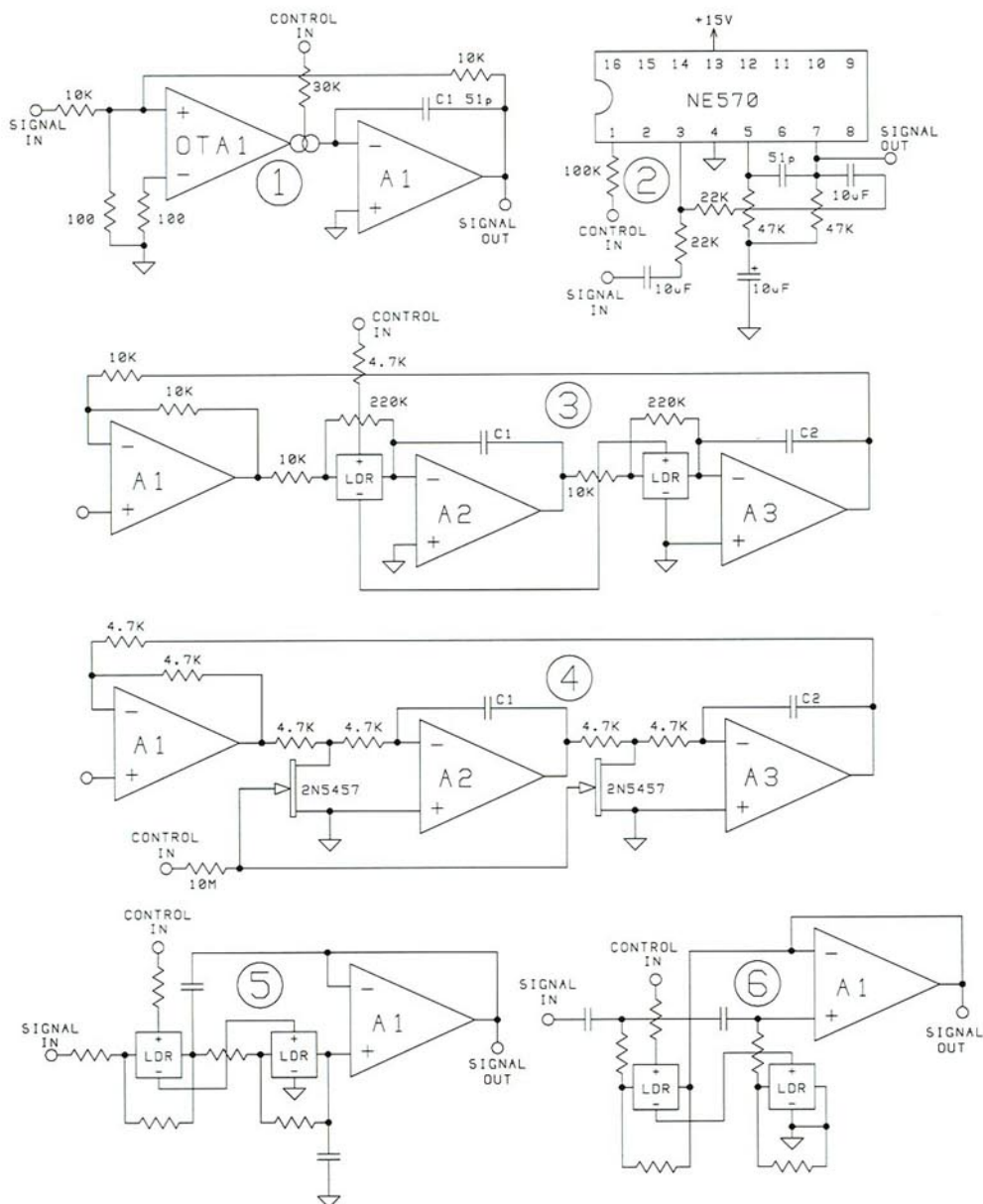Fig. A148. Practical stomp-box voltage controlled filters. 1—VC lowpass filter, 6 dB/octave based on LM3080 or LM13600 OTA. On a ±15V supply, control range extends from +15V (filter open to ~20 KHz) to −14.25V, at which point 20 Hz is ~12 dB down; filter cuts off completely if control is taken below −14.25V. Feedthrough trim same as for 13600-based VCA; do not allow control voltage to drop below −14.25V. 2—VC lowpass filter based on NE570. Control range +15V to 0.95V, at which point filter is ~3 dB down @ 100 Hz. Feedthrough trim uses same technique as VCA and is very effective, but only if control voltage does not swing below 1.2V. 3—State-variable filter core achieves voltage control by substituting LDR-based optocouplers for resistors. 4—Another SVF, but uses FETs as shunt elements. Feedthrough is about 90 mv. Control range spans pinch-off (around −2V) to V+. Advantage over LDR-based circuit in that depth of response does not fall with modulation rate. 5 & 6—LDR-based optocouplers sub for pots in standard active filters; #5 lowpass, #6 highpass.

Voltage Controlled Attack & Decay

These functions find a home in program-responsive level detectors, and in some types of distortion generators. Most use transistors to alter the rate of capacitor charge or discharge. Practical circuits appear in the Appendix on Dynamic Effects.

Consumer Electronics ICs

Integrated circuits meant to enable voltage control of common home/car stereo functions (volume, balance, bass & treble) offer considerable potential in the pedal realm. Fig. A145B shows the basic function and a few applications of the TDA1074A.
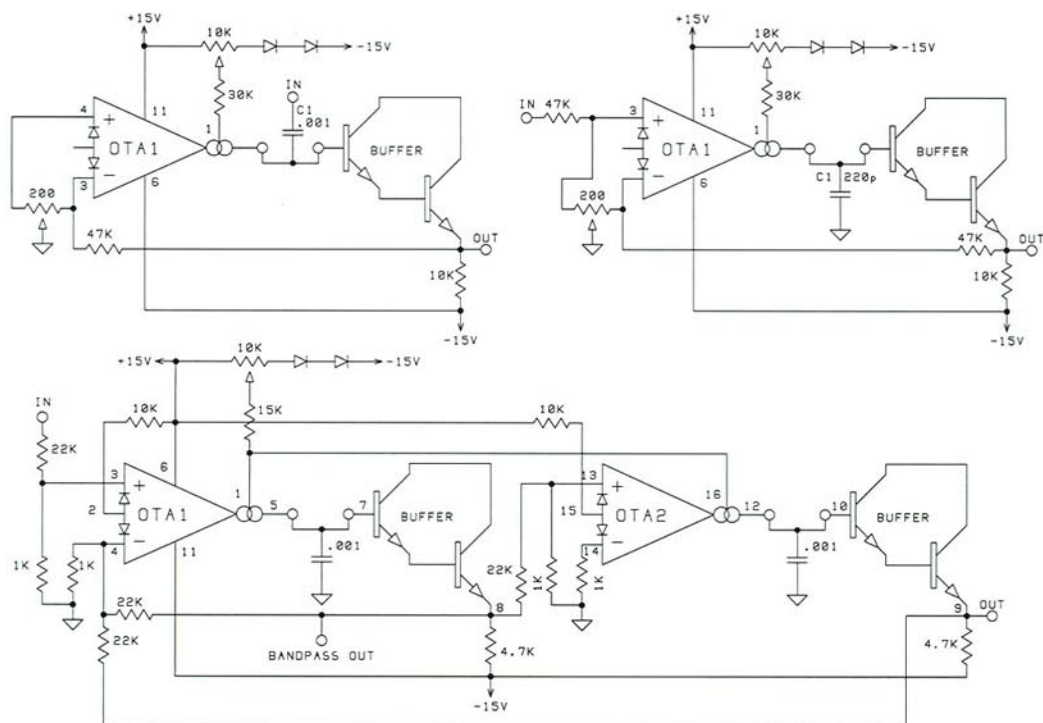
Fig. A149. 13600-based VCFs. 1—VC highpass filter, 6 dB/octave, range ~40 Hz to 4 KHz 2—VC lowpass filter, 6 dB/octave, range ~ 200 Hz to 10 KHz. 3—Unity-gain voltage controlled state variable filter; component values shown allow tuning over the range ~40 Hz – 5 KHz. All three filters suffer significant feedthrough; all three cut off completely if control voltage falls much below –13.8V. Cascade stages for sharper slopes. Diodes 1N914 or similar.
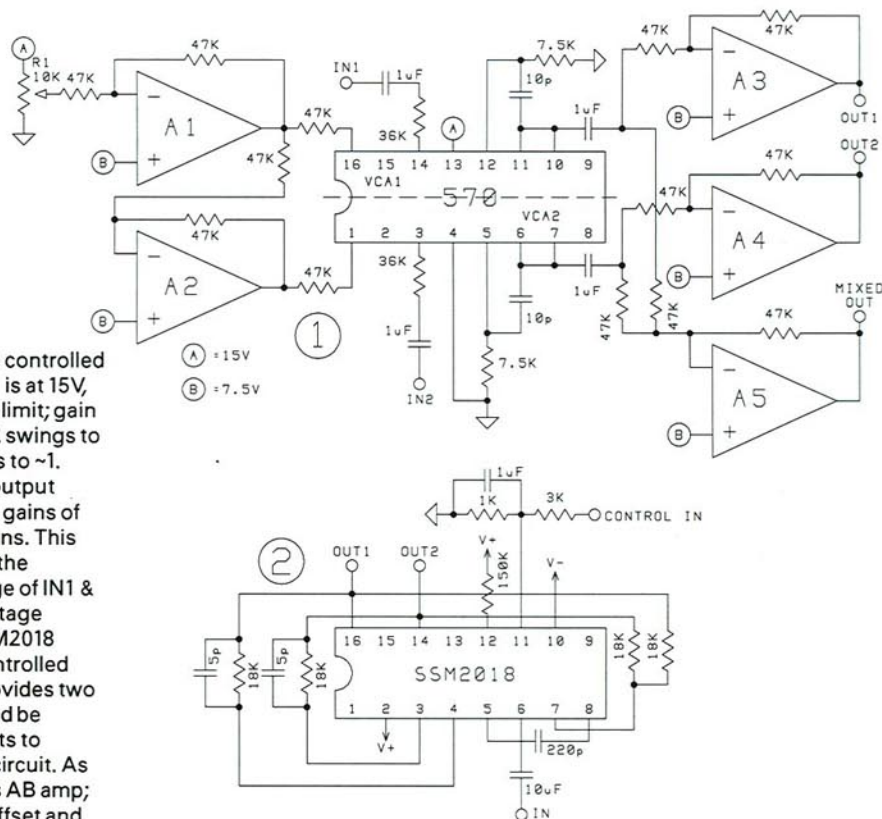
Fig. A150. NE570-based voltage controlled panner/mixer. When wiper of R1 is at 15V, output of A1 swings to negative limit; gain of VCA1 drops to 0; output of A2 swings to positive limit, gain of VCA2 rises to ~1. When wiper of R1 is at ground, output states of A1 & A2 invert; relative gains of VCAs move in opposite directions. This action pans the signal between the outputs, or varies the percentage of IN1 & IN2 in the mixed output. 2—Voltage controlled panner based on SSM2018 'OVCE' (operational voltage controlled element). Chip is a VCA that provides two inverse-gain outputs. VCM could be achieved by feeding both outputs to summing amp, as with the 570 circuit. As shown, chip functions as a class AB amp; can be configured for class A. Offset and symmetry trims not shown.

Fig. A151. Coupling the control voltage to the VCB commonly requires manipulation of amplitude and DC offset. Op amps do the job in many cases. 1—Circuit typical of tremolos, phase effects, etc. R1 varies signal level applied to R2, subjected to gain in A1 equal to R3÷R2. R5 controls DC offset present at A1's output. Signal inversion is irrelevant to most rhythmic effects. 2—Noninverting stage commonly seen in dynamically responsive effects, such as compressors and de-essers. Choice of A1 matters, because op-amp types differ in their maximum output voltage swing. When working up an effect, use the op amp that will be use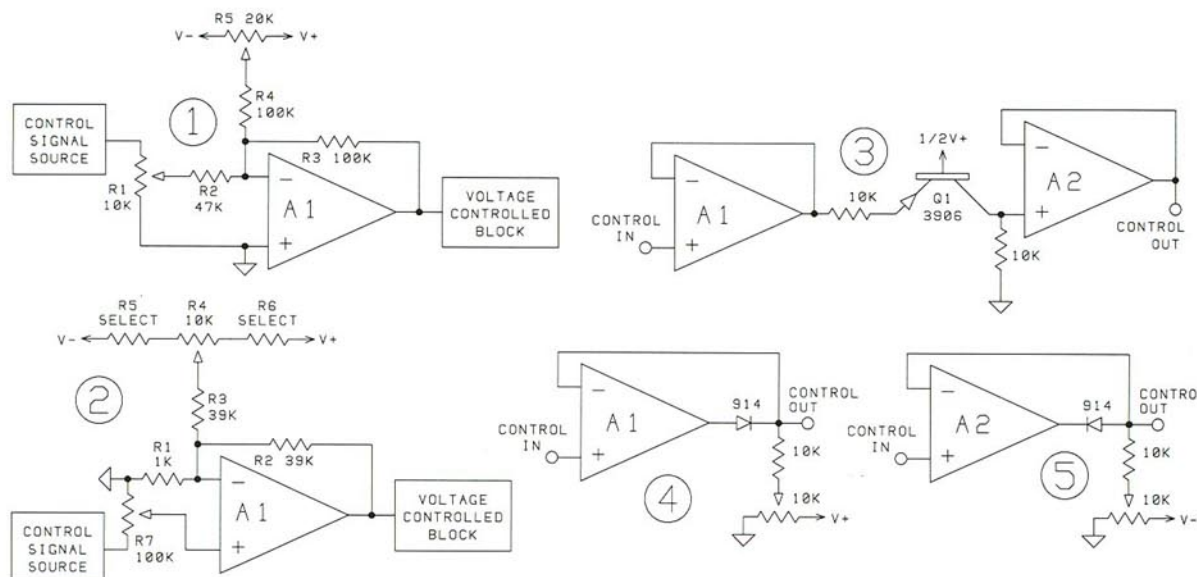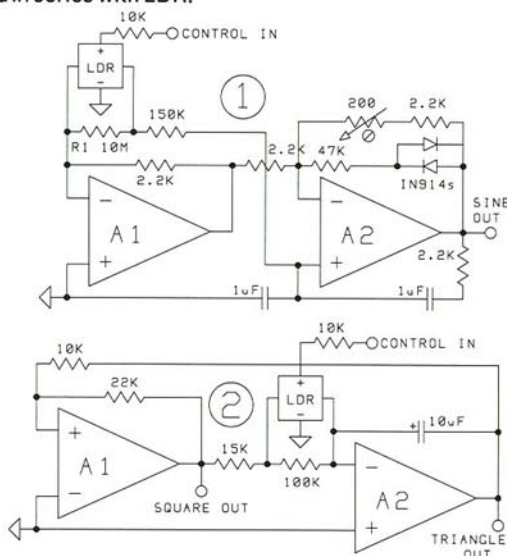d in the final circuit. 3—A control voltage that varies between $\frac{1}{2}$V+ and V+ can be level-shifted to vary between GND and $\frac{1}{2}$V+, by coupling through a PNP transistor. This is useful in, for instance, an NE570 using an external level detector which is referenced to $\frac{1}{2}$V+. A2 should be 324/358 or other type whose input and output embrace the negative supply. 4—Means to define an arbitrary lower limit of a control voltage that varies between GND and V+; setting of 10K pot defines limit. 5—Means to define an arbitrary upper limit in a control voltage that varies between GND and V−; setting of 10K pot defines limit.

Coupling Control Voltage to VCB

The coupling to the VCB provides a convenient point to manage control voltage amplitude, and, often, the magnitude of a DC offset impressed upon the control feed. The DC offset sets the VCB's state in the ab-sence of a control input; the control voltage alters that state rhythmically or dynamically.

Coupling through an op amp neatly meets the need. Op amps allow simple management of AC signal gain, as well as impression of a DC offset upon the AC. A single op amp acts as a complex summing block by utilizing its inverting and noninverting inputs.

Each class of op amp offers a slightly different positive and negative output swing which the builder must reckon during design. With some exceptions, the output won't swing rail to rail. Pertinent upshots arise in TL06X/7X/8X op amps, whose outputs don't swing negative enough to cut off a 13600-based VCA; MC3317X and LT1097 types swing within ~0.5V of the negative supply, and will cut off the VCA. LM324/358 outputs swing to ground in single-supply circuits, but only if the noninverting input is biased to ground. Many CMOS op amps swing rail-to-rail. These observations make it prudent, when working up a voltage controlled effect, to drive the VCB off the op amp likely to perform this function in the circuit. Level shifting is readily achieved using an external transistor (Fig. A151–3). Some circuits require the control voltage to observe positive or negative limits, an easily realized feat (Fig. A151–4, 5).

Fig. A152. 1—Sinewave oscillator adapted to voltage control by placing LDR in parallel with R1. To change minimum rate, change value of R1; to change maximum rate, change 150K series resistor. 2—Square/triangle oscillator similarly adapted; similarly tuned by altering resistance in parallel and in series with LDR.

Feedthrough

VCBs suffer varying degrees of leakage of the control

230

voltage into the signal path. Blocking feedthrough of a static voltage is as easy as coupling through a capacitor; rhythmic and dynamic artifacts will penetrate. Rhythmic feedthrough manifests as pulsation or clicking. Dynamic feedthrough manifests as distortion; sometimes as a click or pop coincident with a musical peak. Audibility of feedthrough depends on the amplitude and shape of the modulating waveform, and, if regular, the rate, faster rates tending to intrude. The problem is reduced by these measures:

- trimming
- configuring VCAs to vary gain from 0 to 1
- placing a highpass filter after the VCB to attenuate subsonic energy
- avoiding sharp edges in the control feed
- gating
- quasi-companding or a pre-/de-emphasis system
- internal companding

A click sometimes accompanies each fresh transient in dynamic effects that use instantaneous attack. This does not represent feedthrough, but the result of impressing a sharp edge on the audio. The problem sometimes abates when attack is delayed a few milliseconds.

Squarewave sources are notorious for polluting audio circuits with clicks, even when they don't connect to the signal path. Sub-microsecond rise and fall make powerful high-frequency harmonics that defeat bypass measures. Effects using true squarewave control should mount the squarewave source close to ground, and give analog and 'digital' circuits independent return paths to ground and separate ties to the power bus. The occasional circuit benefits from isolating the squarewave source from the power bus through an LC network. Because few effects need squarewave control, the builder should at least test alternatives, such as a soft-clipped sinewave.

Miscellaneous Adjuncts

Panning & Mixing

Pedal effects present the need to blend signals continuously from wet to dry. A pot tied between low-impedance sources acts as a variable voltage divider/mixer. When the wiper is at one or the other extreme, the dominant divider action takes place between the selected op amp's output impedance and the full pot impedance. Wet/dry isolation is not complete but typically exceeds 60 dB, adequate for most effects. Proportional mixing takes place as the wiper moves from either extreme toward the center.

Applications requiring complete isolation can use Fig. A160–2, a variation of which makes a panning circuit in which the square of the sum of the output voltages remains constant (Fig. A160–3; Ref. 6).

Altering Potentiometer Taper

A linear pot used in three-terminal (voltage-divider)

Fig. A160. 1—Simple mixing circuit useful in most effects. 2—More elaborate circuit ensures 100% wet or dry at full extreme. 3—Panning circuit in which sum of squares of left and right outputs is constant throughout panning range.
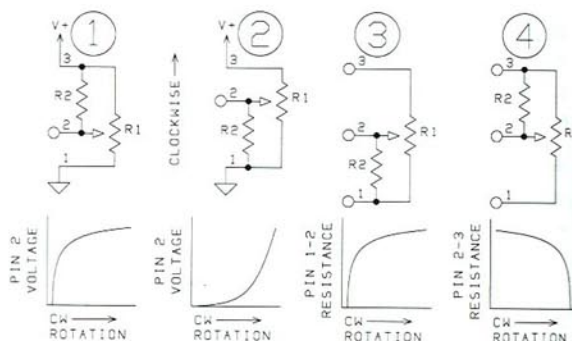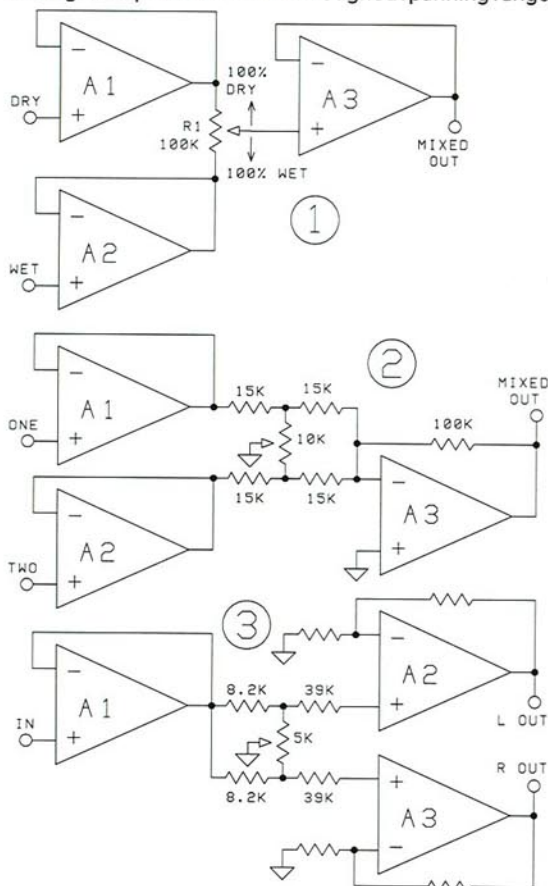

Fig. A161. Linear-taper pot R1 can simulate a reverse-log–taper pot by wiring R2 between terminals 2 and 3, but only when pot acts as a three-terminal voltage divider; the voltage output of such a pot follows a log curve. Severity of taper depends on ratio of R1:R2; the higher the ratio, the more severe the taper. 2—R1 simulates a log-taper pot in voltage-divider mode by wiring R2 between terminals 1 and 2; voltage output of this pot follows a reverse-log curve. 3 & 4—Taper of linear pot used as variable series resistance (two-terminal mode) can be altered, but alteration becomes sensitive to direction of pot rotation, and is useful for applications requiring fine control as pot resistance approaches maximum.

mode can be given a log or a reverse-log taper by paralleling a resistor with the appropriate arm (Fig. A161–1, 2). A linear pot used as a series variable resistance can be altered such that control becomes finer as resistance approaches maximum (Fig. A161–3, 4).

Level Indicators

Level indicators have countless uses in the music scene, displaying input or output level, amount of gain reduction, and so forth. LED bargraphs usually run off one of the ubiquitous LM391X chips. Each contains a 10-comparator ladder, a voltage reference, and constant-current LED drivers. The 3914's ladder is linear, the 3915's logarithmic; the 3916's approximates the VU meter scale. Basic use calls for matching control voltage to the chip, or tuning the chip's response to a fixed control voltage (Fig. A162–1).

New analog meters typically sell for $20 and up, but used or surplus meters often surface for $10 or less. Infinite resolution coupled with a cool/retro look make them attractive options. Adapting the meter to the available drive usually requires trimming (Fig. A162–2). Meters 100 μa or less may require a shunt resistor to avoid very high trim resistance, which drops a significant voltage from small induced currents.

Antidotes to Deadening

Heavy compression, or moderate compression propagated in dubbing, deadens sound. The greater the ag-
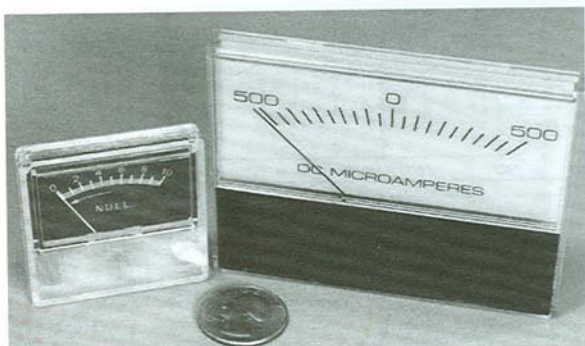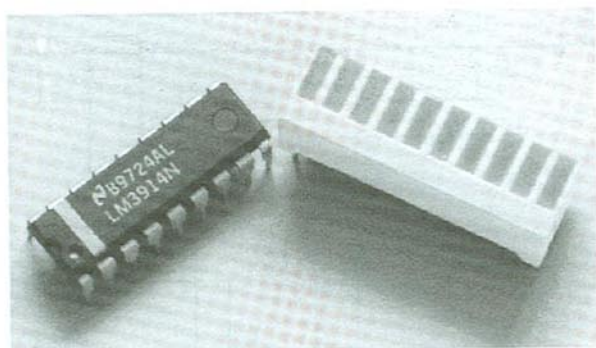
Fig. A162. 1—Basic LM3914 bargraph driver circuit suiting many applications in stompbox & rackmount projects. Circuit shown adjusts to display input voltage by lighting 10 LEDs in sequence. For example, if compression control voltage varies 0V to 3.2V, apply 3.2V to pin 5; trim R1 so LED10 barely lights. Internal resistor ladder divides 3.2V into 10 segments, so a new LED lights every 0.32V of input voltage (the log-taper of the 3915's resistor ladder better suits audio applications). Chip runs off 3–18V. Value of R2 controls LED brightness; 5K shown is fine for breadboarding; change to 10K pot in series with 1K and trim to desired brightness. 2—Adapting a meter to existing control voltage calls for knowledge of meter's nominal sensitivity; assume 1 ma full scale. If the voltage being measured varies 0V to 5.0V, choose R1 to deliver a current of 0–1 ma to the meter. The meter has internal resistance which forms a divider with R1. While the right value of R1 could be calculated, trimming is virtually always required, so make R1 a trimpot. Set R1 to maximum resistance; apply 5.0V; trim so meter reads full scale. R2 represents a shunt resistor that might be needed with, say, a 50 μa meter working off a 10K trimpot. Meters come with linear and log-tapers, easily ascertained in unmarked surplus samples by manually applying a voltage and noting meter response.
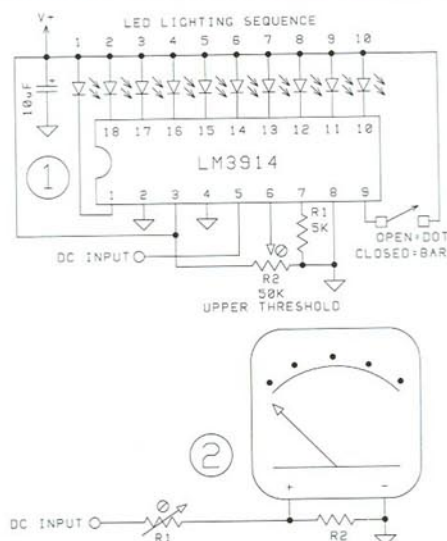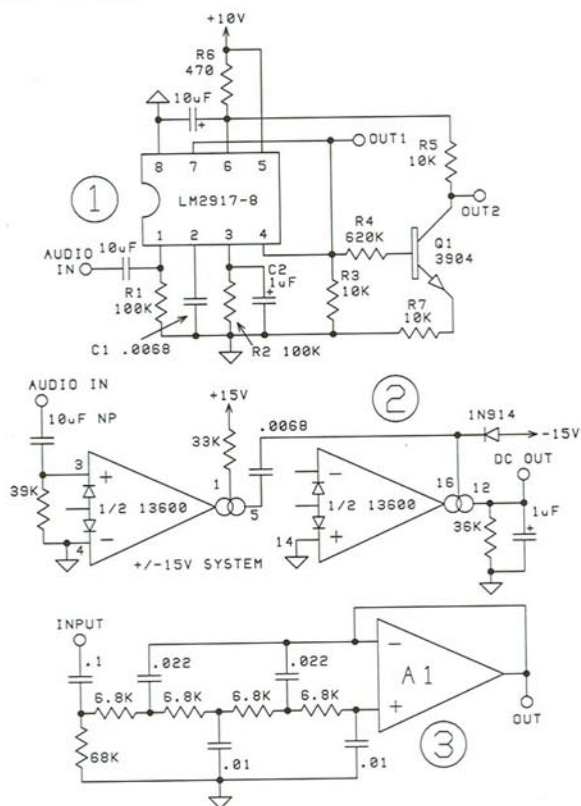

Fig. A163. (left) 1—F–V converter useful for guitar. OUT1 voltage ranges 0.1V @ 80 Hz to 6V @ 2400 Hz or above. OUT2 gives the inverse control voltage. Circuit applies only to the 8-pin version of the LM2917. 2—Functionally similar circuit based on LM13600. In both circuits, the frequency at which maximum voltage is obtained can be lowered by raising the value of the 0.0068μF cap; the frequency can be raised by decreasing this cap's value. 3—To avoid having maximum output triggered by high-frequency distortion products, many applications need a sharp lowpass filter ahead of the input; cutoff above ~1500 Hz, such as this quasi-24 dB/octave design.

gregate compression, the deeper sparkle recedes. One answer to this problem is to apply compression-dependent treble boost. The technique dovetails with meter drive circuits because the compression control voltage also controls the treble boost. A second approach applies treble emphasis to the signal feeding the compressor, but not to the signal feeding the level detector. Matching de-emphasis occurs in a VCF after the compressor, such that net frequency response is flat below the compression threshold. Above threshold, the compression control voltage reduces the de-emphasis of the VCF. This method is quieter than compression-dependent treble boost.

The notion of vitalizing sound goes back at least half a century. Approaches include delay, phase shift,
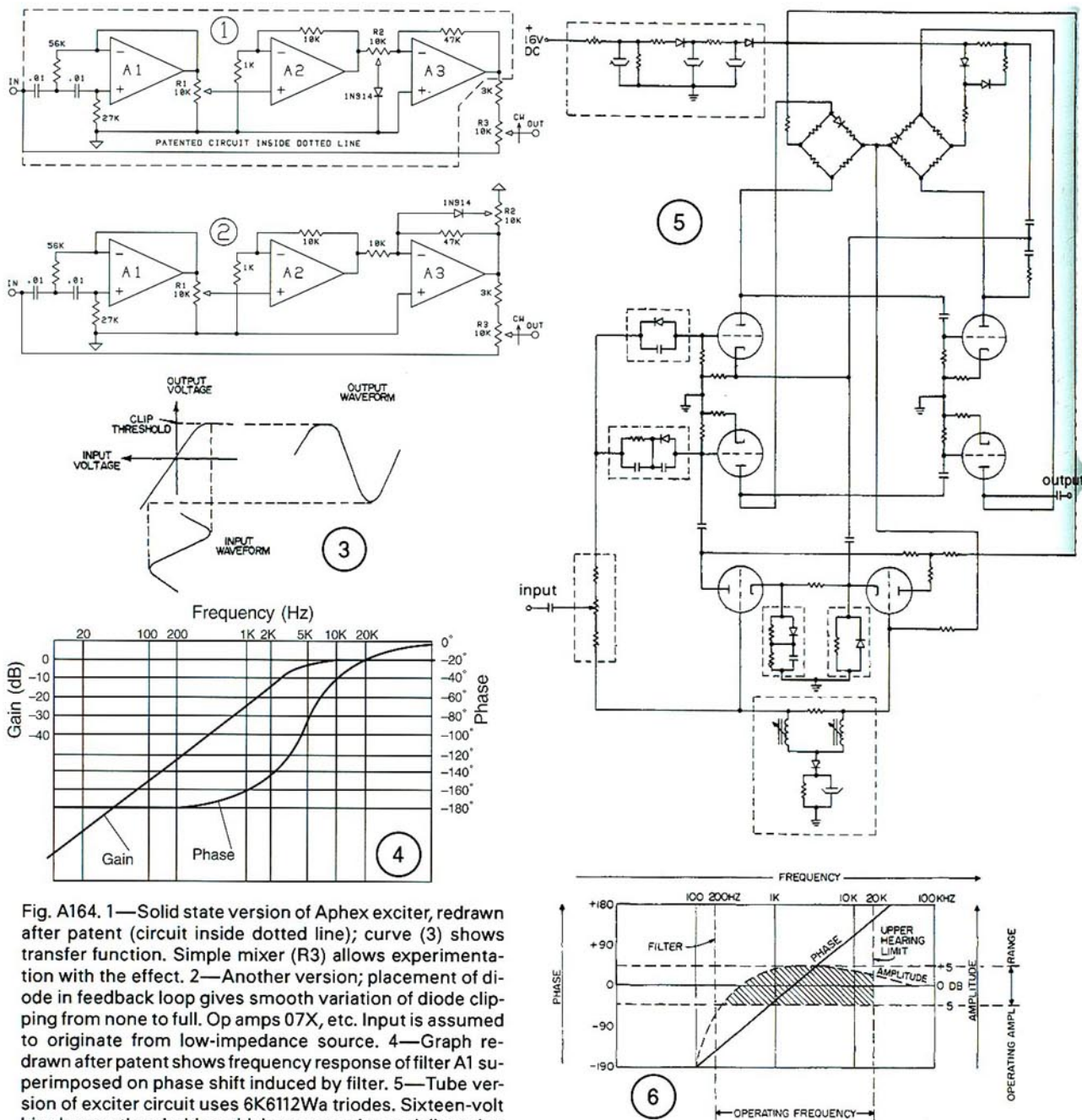
Fig. A164. 1—Solid state version of Aphex exciter, redrawn after patent (circuit inside dotted line); curve (3) shows transfer function. Simple mixer (R3) allows experimentation with the effect. 2—Another version; placement of diode in feedback loop gives smooth variation of diode clipping from none to full. Op amps 07X, etc. Input is assumed to originate from low-impedance source. 4—Graph redrawn after patent shows frequency response of filter A1 superimposed on phase shift induced by filter. 5—Tube version of exciter circuit uses 6K6112Wa triodes. Sixteen-volt bias lowers threshold at which compression and distortion kick in. 6—Graph shows several aspects of operation. Figs. 3, 5, & 6 copied from Ref. 37.

spectral shift, and distortion; and combinations of these changes, fleshed out in products like the BBE Sonic Maximizer® and the Aphex Aural Exciter®.

The original Aphex patent describes highpass filtering—specifically including the resultant phase shift—followed by a specific type of distortion. Wet signal mixes with dry in varying amounts. While operation of the solid-state exciter looks straightforward, the patent itself questions what's happening in the tube version (Fig. A164–5; the patent is intriguing to consider in the broader context of tube sound).

Frequency-Driven Effects (FDEs)

Frequency is to FDEs as loudness is to dynamic effects. FDEs derive a control voltage proportional to the frequency of the input, ignoring amplitude above a trigger threshold. Effects made possible by this approach include:

► FD tremolo (rate, depth, waveform shape)
► FD wah
► FD distortion (e.g., a distortion multiplexer driven off a VCO whose rate varies with input frequency)
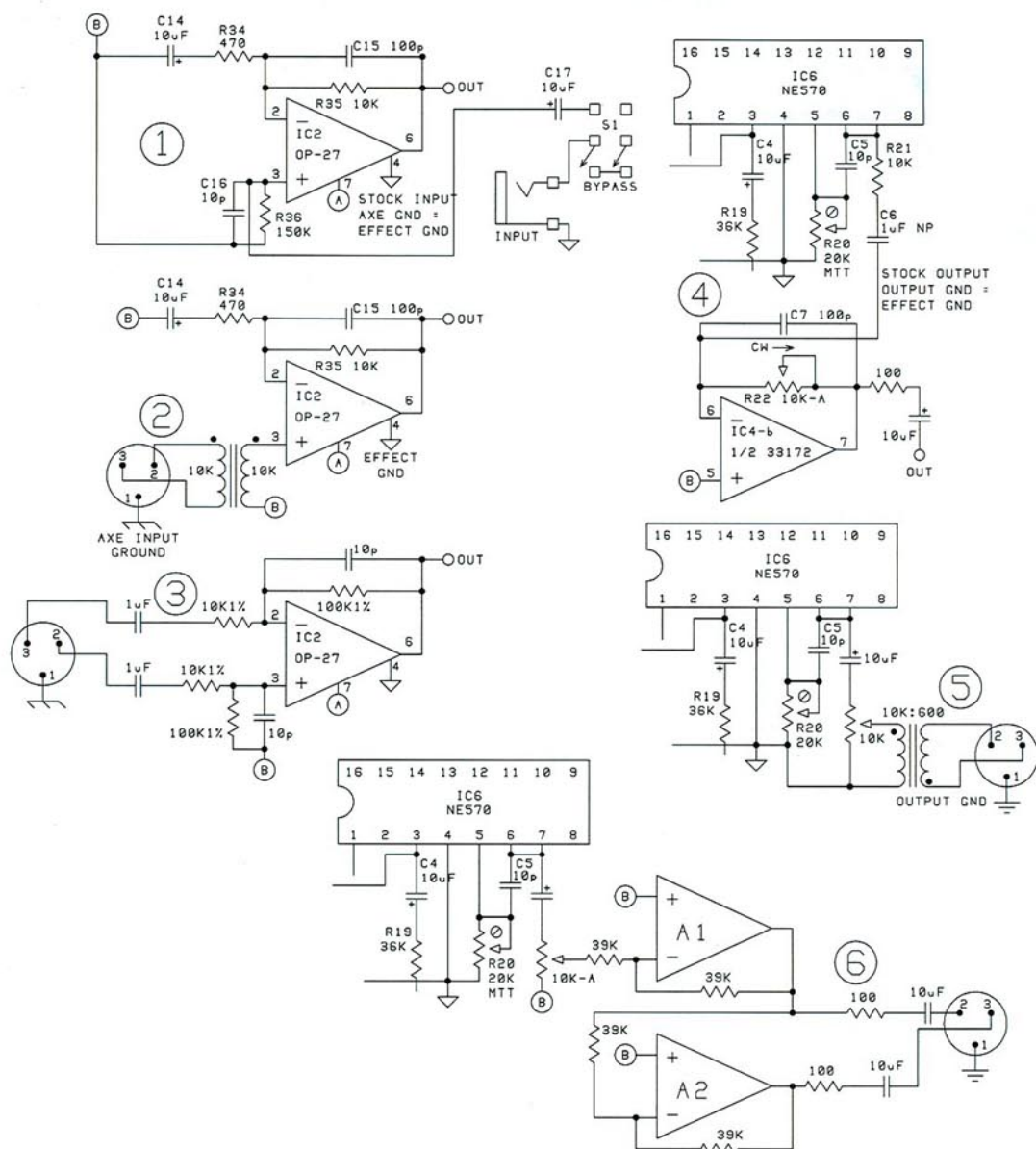► FD mixing

Fig. A165. Typical options for adapting unbalanced stomp-box I/O to balanced. 1—Tremolo-Matic II's single-ended preamp. Axe ground ties to effect ground. 2—Same circuit adapted to balanced input using 10K:10K transformer; axe ground does not contact effect ground. 3—Balanced preamp with gain of 10; separate grounds. 4—TM2's stock single-ended output; effect ground ties to output ground. 5—Balanced output achieved with transformer; effect ground now isolated from output ground. System must account for ~12 dB of voltage loss from 10K:600 transformer; not usually a problem. 6—Active balanced output.

▶ FD phase (best in stereo)
▶ FD delay/reverb/echo

The possibilities expand if the FD voltage is summed with or subtracted from that of a level detector, making the control voltage a function of two variables.

FDEs derive a control voltage from a frequency-to-voltage (F–V) converter, sometimes called a tachometer. The common need for such a circuit spawned a number of integrated circuits, such as the LM2907/2917 series; the LM13600 also makes a useful F–V converter (Fig. A163–2).

FDEs have to be treated as if they were frequency dividers, meant to process one note at a time. The output becomes unpredictable when the converter tries to process a chord. A six-string's fundamental tones cover ~80–1280 Hz, but much of the amplitude is harmonic. Predominance fluctuates between fundamental and harmonic as notes decay. Pure tone sources,

such as keyboards or synthesizers, give more predictable response from F–V converters.

Going Balanced

Most stomp boxes provide unbalanced input and output stages, a convention followed by most projects in this book. Going balanced lets the box interface with rackmount gear equipped only with balanced inputs and outputs, and prevents ground loops by isolating axe ground from effect ground or (more typically) effect ground from amp ground. Balanced input stages' ability to reject common-mode noise plays a minor role in the grand scheme of hum.

Transformers make a simple solution. Mating them to input or output is a formality. Their simplicity is offset partly by susceptibility to magnetically coupled hum, and by limited selection in the hobby price range.

Active stages deliver greater versatility at the cost of greater noise. A balanced input stage is known as a differential amplifier. As preamps, balanced stages are unavoidably noisier than comparable single-ended stages, because they contain two transistors, each with its own E_n and I_n. Quadrature addition results in ~41% more noise than an identical single stage. Typically, this noise is negligible when dealing with axe-level signals.

The SSM2141 and SSM2142 make single-chip alternatives to discrete-component balanced inputs and outputs. A single transistor can supply a balanced output in a manner reminiscent of tube amps' catho-dyne phase inverter.

Ground lift makes a key option in balanced gear. Providing for this may demand a plastic jack in a metal panel, or a plastic panel to mount a metal jack. The shield lead of twisted-pair cable needs to contact only one of the ground systems.

New Uses for Pedal Pots

Rather than merely switching in or out of the signal path, pedal effects vary under control the player's foot. Common treadle mechanisms consist of a rotary pot whose shaft is fitted with a gear, driven by a short, toothed strip hinged to the toe-end of the treadle. The shaft turns when the pedal rocks (Fig. A168). To avoid damage at the stops, the pot's wiper does not traverse its rotational limit.
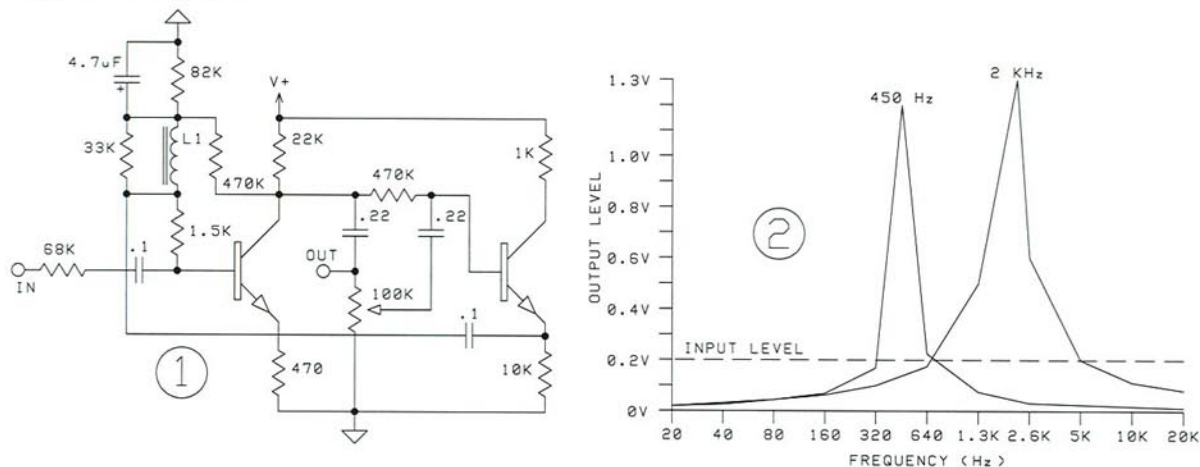
The commonest pedal effect is the wah-wah, a variable bandpass filter that moves over the approximate range 450–2000 Hz. Fig. A167 shows the core schematic and rough response curve from a wah pedal made late in the eighties. Unlike fuzz, wah has a musical antecedent in the sound of a rubber cup flapped at the mouth of a cornet. It is apparent from the schematic and the nature of the effect that wah is readily customized by changing frequency range, Q, and gain.

Beyond wah, pedal pots' potential remains unclaimed. A treadle pot opens the door to effects such as:

▶ pedal-controlled panner
▶ pedal-controlled mixer
▶ pedal-controlled non-wah filter
▶ pedal-controlled tremolo (rate, depth, waveform shape)
▶ pedal-controlled delay effects (phase, chorus, 'verb, echo)
▶ pedal-variant VCO tied to ring modulator
▶ pedal-variant distortion percentage wet/dry, or switching rate of a distortion multiplexer
▶ multiple effects slaved to one mechanism (e.g., multichannel crossfader)

The builder need not buy a new pedal to experiment. Temporarily desolder the pot leads from a wah or volume pedal, wire the pot to the power supply. Fig.

Fig. A167. 1—Basic schematic of one type of wah pedal. 2—Crudely plotted I/O curve of circuit reveals bandpass that shifts between ~450 and ~2000 Hz.

toothed strip

pot

pot

gear

toothed strip

Fig. A166. Basic ring modulator circuit. Instrument feed multiplies against constant tone in 4QM; output consists of sum and difference products. Use a sine oscillator for clean sound; a squarewave oscillator for a nasty quality. Substitute any 4QM for 534: LM1496, OTA, etc.

A168 gives a generic means to adapt a pedal pot to voltage controlled effects.

Ring Modulators

Among guitar effects, *ring modulator* defines a four quadrant multiplier in which the axe feed meets a single frequency generated by an oscillator (Fig. A166). The output consists of sum and difference products whose dissonance resembles that of voice scramblers based on the same principle. Ring-modulator–like effects also arise when the axe feed beats against itself in a 4QM, but possess a more musical sound.

Ring modulation's potential seems as neglected as the effect is sparingly used. By making the oscillator a VCO, oscillator pitch becomes level-dependent or frequency-dependent; or varies rhythmically from a secondary oscillator, or manually through a pedal.
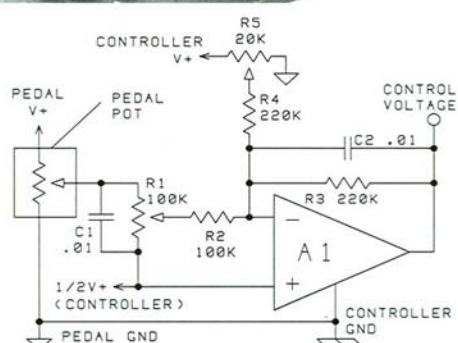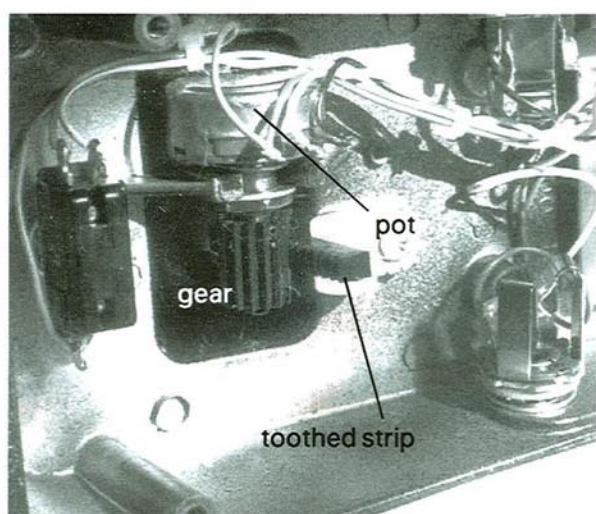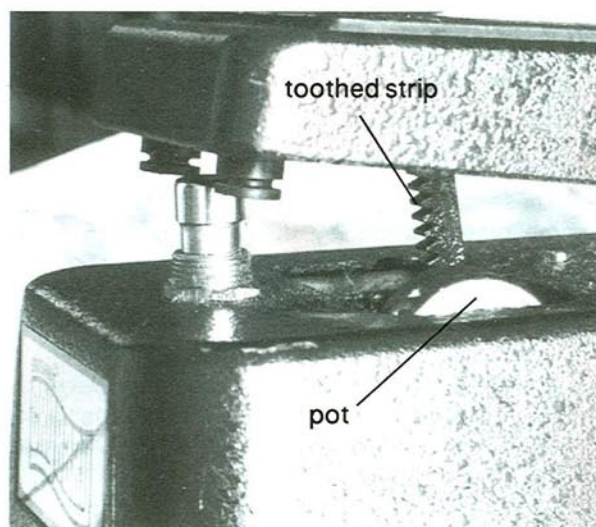

Fig. A168. Photos show mechanism found in many wah and volume pedals, a rotary pot w/gear attached to shaft, driven by toothed strip linked to treadle. Pedal is easily adapted to controlling external effects by desoldering the three wires, labeling them and insulating their ends, then wiring pot to battery as shown in schematic, a generic means to adapt pedal mechanism to drive voltage controlled effects. Output of pedal pot is a voltage that varies from less positive to more positive, because wiper does not traverse its full range. C1 & C2 help reduce scratchy response as pedal pot ages. R1 varies the percentage of pedal voltage into inverting amp A1; R4 determines static voltage at A1's output.
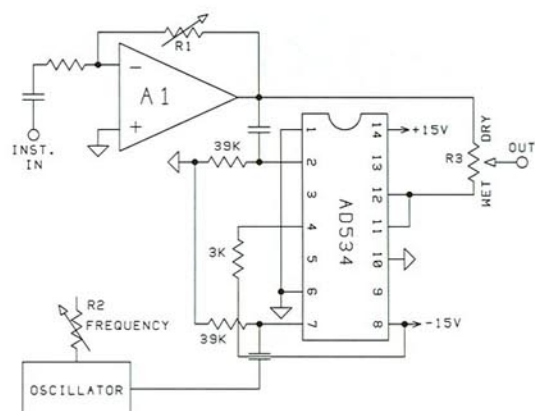
Cooking Up New Effects

The several stomp-box ingredients equip hungry builders to cook up effects that don't exist, or that can't be had in the form desired. The process follows a plan that quickly becomes intuitive:

1. Define the effect as specifically as possible.

2. Prepare a block diagram of the effect.

3. If the block diagram looks feasible, substitute specific circuits for blocks. This narrows many choices: single vs. dual supply, battery vs. AC power, discrete components vs. integrated circuits, and so forth.

4. Prepare a schematic, breadboard the effect, test the sound. This is the key step, for *sound defines success*. Cooking up effects is easy. Getting effects that sound good takes work.

5. Be aware that first attempts often miss the mark; reworking and fine tuning are ways of life. Return to Step 2 if the effect fails to dazzle.

6. Once you have a circuit that gives the desired sound, prepare a final schematic, lay out a printed circuit board, build the box.

Example 1

Let's design a compressor specifically for guitar. To simplify use without sacrificing performance, we'll incorporate auto-variable attack & decay; to suit stately tastes we'll add switchable soft-knee compression. A fixed-gain preamp means that the box needs only three knobs: threshold, ratio, and output level; plus a switch to choose the knee.

These specs suggest a block diagram which makes some other choices by default (Fig. A175). The SSM2120 defines a dual-supply system using op-amp ancillary circuits. Not only does this diagram look feasible, but ingredients from the Appendices let us plug in specific circuits. Since we're using the rectifier/log amp inside the 2120, threshold control is achieved with an external noninverting op amp. Auto-variable attack employs the diode-bypass method; auto-variable decay uses the falling-edge detector.

We could plug in any of a dozen blocks for soft-knee compression. To eliminate trimming, which requires access to an oscilloscope, we'll use a summing block and a diode. The only crucial feature is that the diode's forward drop match that of the prototype reasonably well, 1.82V in this case (the drop measured across the diode, not necessarily the drop specified in a parts catalog).

Making the box instrument-specific calls for limited level detector bandwidth. Listening tests defined a bandpass filter that rolls off at 12 dB/octave below 60 Hz and above 2500 Hz.

Because it's meant to live in a pedal and doesn't need tremendous headroom, we'll optimize the circuit for twin-9V power. This means choosing appropriate values for the 2120's bias resistors. This recipe resulted in Squeeze-O-Matic II.

Example 2

Confusion between tremolo and vibrato suggests an intriguing question: How do the two sound combined? Let's resolve the issue by building a box not just to pair the functions, but to run them synched or independently. We'll use 13600-based allpass filters, but these

Fig. A175. Block diagram of prospective guitar-specific compressor. Choice of SSM2120 defines several other options. From that point, design becomes a matter of chaining subcircuits, leading to a schematic detailed enough to allow sound tests.
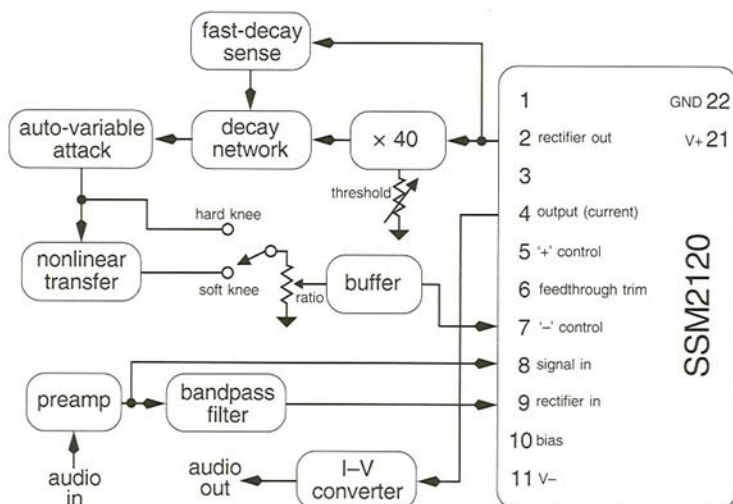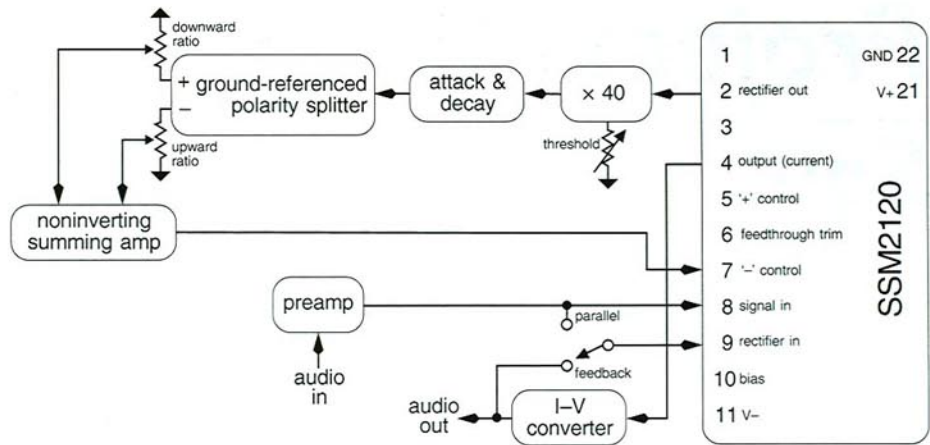
Fig. A176. Block diagram of prospective sustainer with separate upward and downward compression ratios. The only thing that distinguishes this device from an ordinary dual-mode compressor is the polarity splitter and dual ratio control. Because feedback sustain sounds distinct from parallel sustain, the box offers both modes.

saddle us with significant feedthrough. To vary from Phase-O-Matic's approach, we'll gate the effect, using essentially the same circuit as Tremolo-Matic III. The tremolo will use our standard no-feedthrough 570-based VCA. For variety we'll choose the ancient op-amp triangle oscillator rather than a sine oscillator. This recipe resulted in Tremolo-Matic V.

Example 3

Passive in-axe mods are big at the moment. Let's get in on the action with something worthwhile, a mod not subtle; something to make the stock tone pot a quaint memory: a passive sweepable mid.

To vary frequency continuously using only caps and resistors, we have to resort to phase as the gain manipulation element. The technique is the one responsible for fixed passive notches in old Gibson amps. We'll put the network in series with the pickup; a dual pot replaces the stock tone pot. This recipe produced Axe-O-Matic II.
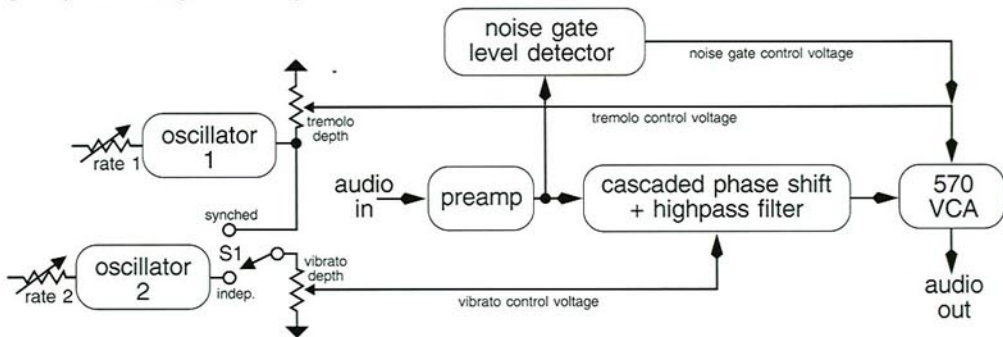
Example 4

The problem with some dual-mode sustainers is that the ratio control affects both modes. Some situations crave independently variable downward and upward sustain. By recognizing a sustainer as a dual-mode compressor, the only innovation required is a means to separate gain reduction from boost.

Using a 2120-type VCA, the '–' control port dictates unity gain when it sees ground potential; positive voltage reduces gain, negative voltage boosts gain. The box needs a circuit that seamlessly separates positive from negative with respect to a crossover at ground. We could use diodes, but their forward drop leaves a gap of ~0.6V on either side of ground; no big deal in many applications. Augmented diodes eliminate the forward drop. Once separated, each voltage is controlled by its own pot. The voltages recombine in a noninverting summing block (Fig. A176).

The remainder of the piece is completely stock but for a bit of auto-variable attack, to help prevent clipping caused by delayed attack in upward sustain. The box doesn't need manually variable attack & decay, so we'll put both functions in one stage; tune them to suit guitar or bass by choice of resistance during construction. The breadboard stage exposed a significant DC offset at the output when the VCA applied more than 20 dB of boost. In feedback mode this manifested as cycling. The offset was easily nulled using a trim network.

Because sustain sounds different with parallel and feedback control paths, we'll include both. Listening tests revealed grossly unnatural response in parallel mode that evened out by adding a series resistance to the parallel-mode control path. Sustain-O-Matic II resulted from this recipe.

Fig. A177. Block diagram of proposed tremolo/vibrato. Switchable oscillators let effects run in synch or at separate rates. Noise gate quells resting feedthrough of 13600-based phase shifters.

Project No. 35
Squeeze-O-Matic II

SOM2 invades the upscale world of auto-variable attack and decay; throws in soft-knee compression just for kicks. These features combine with a frequency-contoured level detector to make a smooth-sounding, easy-to-use compressor for guitar.

Circuit Description

Signal Path: Signal input couples through C4 to IC4-a, a noninverting preamp whose gain the builder sets during construction, nominally ~22. IC1-a output couples through R9-C9 to signal port of IC5, half an SSM2120 dynamic control chip. IC5 output (as a current) couples to current-to-voltage converter IC4-c; R10 varies the output level. Signal couples through R11-C12 to the output path. The signal path is noninverting.

Control Path: IC1-a output passes through bandpass filter IC4-b, thence through R8-C10 to IC5's rectifier input. Raw rectifier output feeds IC2-a, where it's boosted by a factor of 40. R22 applies a variable DC offset that sets the compression threshold. IC2-a output feeds decay network D7-C14-Q1-R24. When R24 sees ground, the charge in C4 takes ~3 seconds to decay from 5V to ground. However, if the program decays rapidly, a negative voltage is generated by a fast-decay sensing network made up of IC1 and associated components. This voltage causes Q1 to pass a constant current that neutralizes the C1's charge in as little as 50 milliseconds. The decay-sense network is a falling-edge detector whose function is detailed in an Appendix.

The decay network feeds an auto-variable attack network made up of D3-5, R30-R31, & C15. The basic attack time is about 20 milliseconds. If the program attacks fast enough and has sufficient amplitude, the voltage differential between D5's anode and cathode surpasses its forward drop, at which point the control voltage passes through R30, speeding attack by a factor of 10. If the program attacks extremely fast and has sufficient amplitude, the differential that develops across D4 reaches conduction, at which point attack becomes instantaneous. The attack network output is buffered by IC2-c, feeding D6, which blocks negative voltage. D6's cathode ties to one throw of S2, and to a nonlinear transfer block made up of IC3, IC4-d, and their associated components. Function of this block is detailed in an Appendix; Fig. 35–5 illustrates the transfer function. IC4-d's output ties to S2's second throw. S2's pole ties to pot R41, which varies the per-centage of control voltage into buffer IC2-d, thence through R16 to the VCA control port, IC5 pin 7.

When S2 selects hard knee, the control voltage

SQUEEZE-O-MATIC II PARTS LIST

Resistors
R1 470
R2, 21, 26, 27, 30, 34, 35, 36, 37 10K
R3, 38 150K
R4, 5 3.6K
R6, 7, 19, 20 39K
R8 7.5K
R9 36K
R10 20K audio-taper pot
R11 100
R12, 28, 31, 32 100K
R13, 15 200
R14 47
R16 8.2K
R17, 25, 29 1M
R18, 33 1K
R22 20K pot
R23 2.2K
R24 10M
R39 15K
R40 22K
R41 100K pot

Capacitors
C1 4.7μF aluminum electrolytic
C2 10pF
C3, 11, 16 100pF
C4 10μF nonpolar
C5, 6, 18 0.01μF
C7, 8, 15, 17 0.1μF
C9, 10, 12, 19, 20 10μF aluminum electrolytic
C13 0.0022μF
C14 0.0056μF

Semiconductors
D1, 4, 5, 9 red LED, forward drop 1.82V
D2, 3, 6, 7, 8 1N914
D10, 11 1N4001
IC1, 2, 3 TL064 quad low-power op amp
IC4 TL072 dual op amp
IC5 SSM2120 dual-channel dynamic controller
Q1 2N3904 NPN transistor

Miscellaneous
S1 DPDT switch
S2 SPDT switch
wire, circuit board, solder, ¼" jacks, knobs, etc.
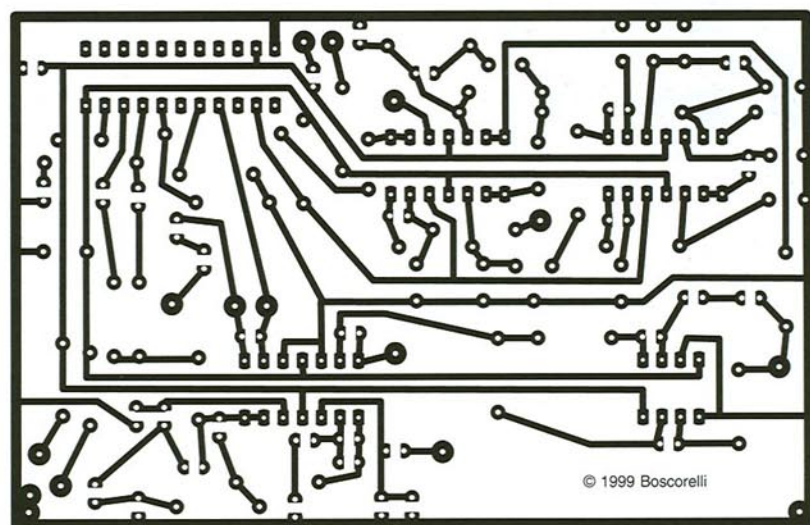
5" × 3.5" reference box

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Fig. 35–1. Squeeze-O-Matic II circuit board.

feeds straight off D6 to R41. When S2 selects soft knee, the control voltage passes through the nonlinear transfer block, then to R41.

Use

Pots & switches have these functions:

R10	output level
R22	threshold
R41	ratio
S1	compress/bypass
S2	knee select soft/hard

Initial settings: R10 fully CCW; R22, R41 fully CW, S2 hard. In this condition the box acts as a preamp. Connect unit to axe and amp, trim R10 for desired output level.

Lower the threshold until obvious compression is noted. Explore the sound of various threshold and ratio settings; try to 'fool' the compressor with fast attack, or a loud, fast chord followed immediately by a very soft passage. Switch to soft knee and again explore the ranges of threshold and ratio.

Using component values shown, threshold extends from ~10 mv$_{p-p}$ (−40 dBV) to >8V$_{p-p}$. Ratio measured in hard-knee mode extends from 1:1 with ratio control at minimum, to infinite with ratio control at 85%. Paradoxical compression results if the ratio is taken past 85%: gain falls as the signal intensifies. This feature was included to let the player emulate 'sag.' If the feature is not desired, change R16 to 12K.

Notes

SOM2 is optimized to run off a pair of 9V batteries. The

ratio control tolerates reasonable shifts in supply voltage without major shifts in threshold.

Auto-variable attack and decay are easy to accept once the player learns to trust them. Their convenience tends to spoil one for manual modes.

Differences Between the Squeeze-O-Matics

Both compressors use SSM2120 VCAs, and the chip's internal rectifier/log amp, but externalize the remainder of the level detector.

▶ SOM is a *spot compressor*: the level detector's bandwidth equals that of the signal path. It will compress any signal between 20 Hz and 20 KHz, and do so according to the control settings. Furthermore, attack, decay, threshold, and ratio exist under full manual control. Certain settings sound unmusical but prove handy for special effects. Versatility is further enhanced by the choice of feedback and parallel control paths.

▶ SOM2 is tuned to sound smooth and natural when compressing guitar feeds. The bandpass filter ahead of the level detector emphasizes the guitar's ~80–2500 Hz range, ignoring subsonic and ultrasonic energy. The auto-variable functions cloak compression while reducing the number of knobs. The control path is parallel, which listening tests showed met the objective of smooth sound.
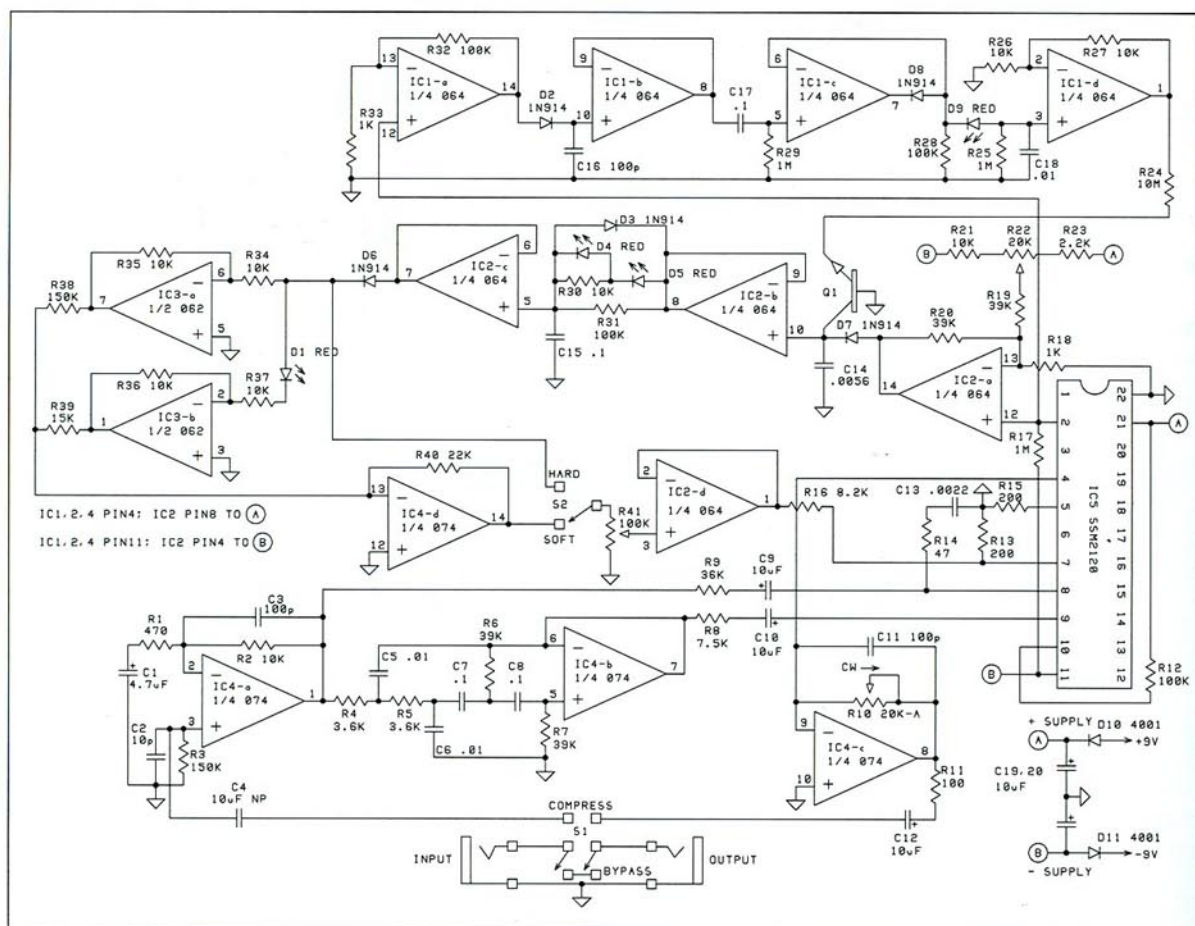
Fig. 35–2. Squeeze-O-Matic II schematic.
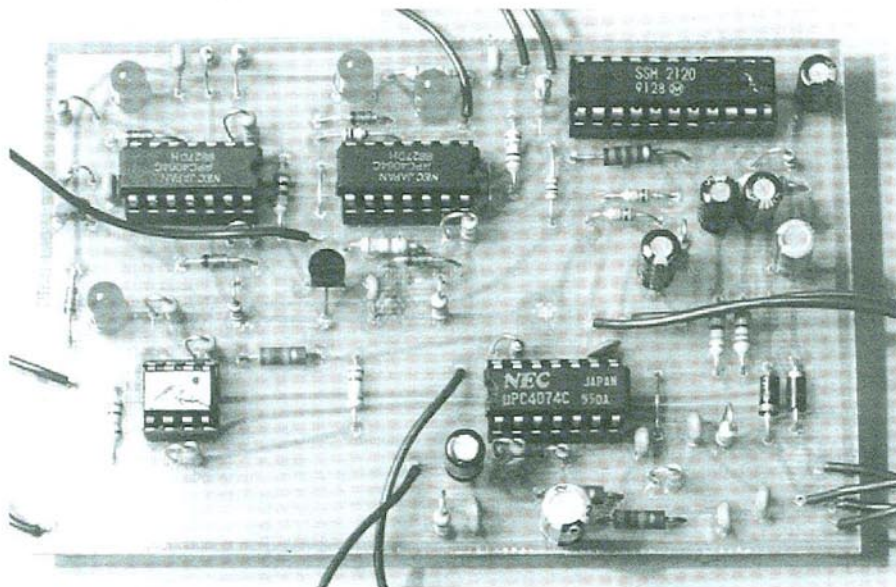
Fig. 35–3. Squeeze-O-Matic II prototype board.
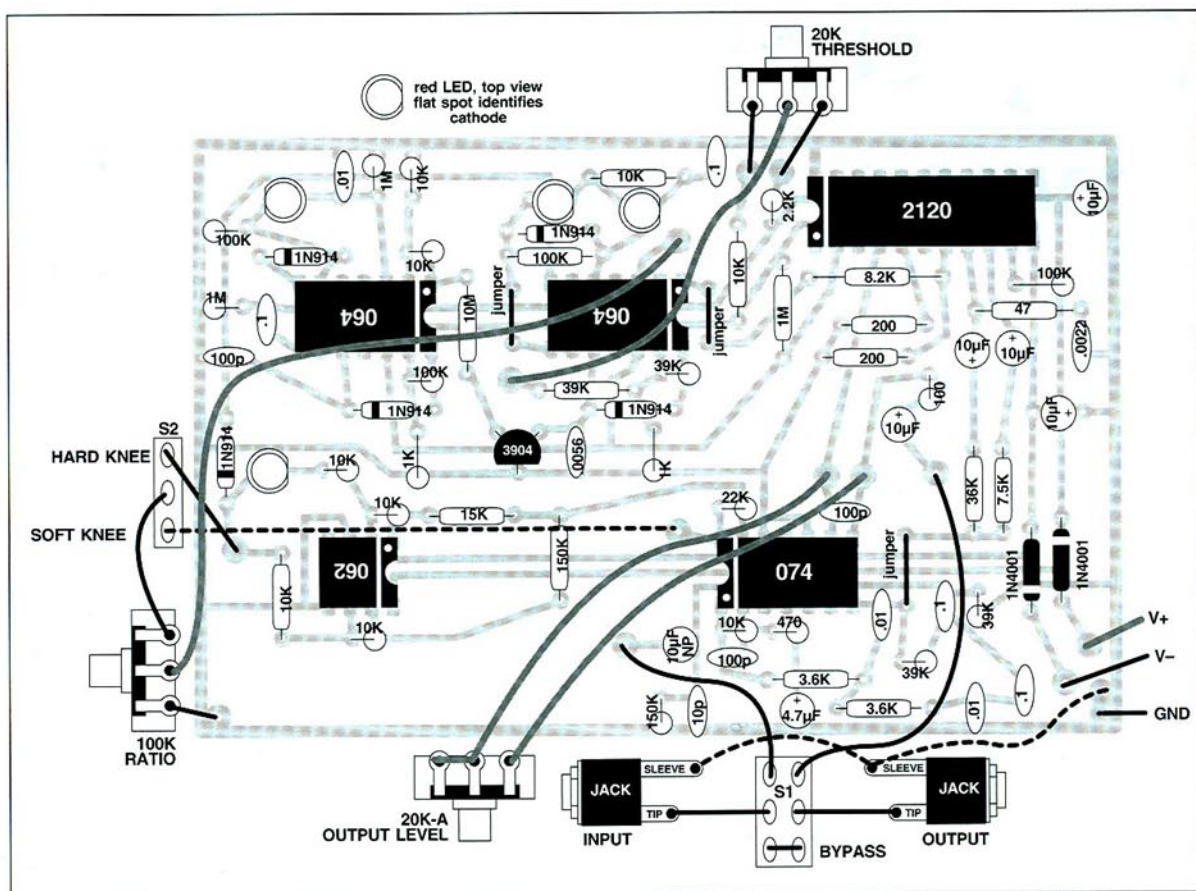
Fig. 35–4. Squeeze-O-Matic II layout & wiring diagram.
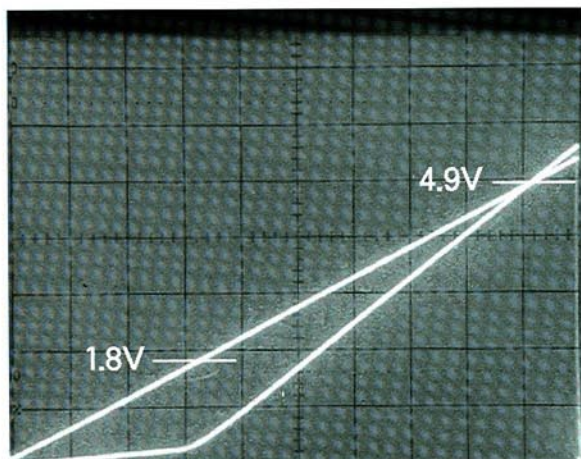
Fig. 35–5. I/O of SOM2's nonlinear transfer block. Below D1's 1.8V turn-on point, only ~15% of control voltage gets through (R40÷R38 = 22K÷150K = 0.15). Once D1 conducts, percentage of voltage increases progressively until, above ~4.9V, it exceeds the input voltage. Despite simplicity, this curve proved highly musical. Scale 1V, sweep 500μs.

The bandpass filter covers the guitar's fundamental tones and low harmonics. To retune the box for bass, double the values of the BPF caps.

The red LED used in the nonlinear transfer block is not critical, but should exhibit a forward drop reasonably close to 1.82V, measured as follows: one end of a 10K resistor ties to 10.0V, other end ties to the LED's anode, whose cathode ties to ground; voltage is measured at juncture of resistor and LED.

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Project No. 36
Tremolo-Matic V

TM5 resolves confusion between tremolo and vibrato by pairing the functions.

Circuit Function

Signal Path: Instrument feed couples through C1 to inverting preamp IC9-d whose output feeds four serial phase-shift networks similar to those used in Phase-O-Matic. Output of the final phase shift network feeds a quasi-18dB/octave highpass filter made up of IC9-c and associated components. The filter attenuates subsonic feedthrough from the phase network, and couples through C18-R47 to IC2, half an NE570 configured as a VCA whose gain varies 0–1, depending on control voltage feeding pin 16 through R48. IC2 output couples to pot R45, thence through R44-C17 to the output path. The net signal path is noninverting.

Tremolo Control Path: IC4 and associated components form a triangle generator whose frequency varies under control of R8. IC4 output couples to R37, which varies amplitude feeding summing amp IC3-d. R33 controls the DC offset present at IC3-d's output. Tremolo control voltage couples through R48 to pin 16 of IC2.

Vibrato Control Path: IC3-b & -c form a triangle oscillator identical to IC4. Output couples to one throw of S2, whose other throw ties to the output of the tremolo oscillator. S2's pole ties to pot R38 which varies the amplitude of the vibrato control voltage feeding summing amp IC3-a. Trimpot R41 controls the DC offset present at IC3-a's output. Vibrato control voltage couples through R54 to all four phase-shift ports wired in parallel. When S2 selects the tremolo control feed, vibrato and tremolo occur in sync; when S2 selects the second oscillator, the two functions observe independent rates.

Noise Gate Control Path: Preamp output couples through C3-R3 to a level detector essentially identical to that used in TM3, and whose function has been detailed elsewhere in the text. Output of the noise gate level detector couples through R20 to base of Q1.

TREMOLO-MATIC V PARTS LIST

Resistors
R1, 6, 9, 27 22K
R2, 43 220K
R3, 5, 11, 20, 29, 30, 31, 10K
R4, 10, 13, 18, 21, 22, 25 39K
R7, 8 100K reverse audio pot
R12 2.2K
R13, 15, 24, 32, 34, 35, 39, 40 100K
R14, 16 1M
R17, 23, 49, 54 4.7K
R19 10M
R26 1K
R28, 37, 38, 45 10K audio-taper pot
R33 20K pot
R36, 47, 55, 56 36K
R41 50K trimpot
R42 50K multiturn trimpot
R44, 53 100
R46 7.5K
R48 47K
R50, 51 330K
R52 33K
[resistors not individually identified on schematic: 16×10K, 4×36K, 8×1K]

Capacitors
C1, 17, 18, 20 10μF aluminum electrolytic
C2 22pF
C3, 8, 10, 11, 21, 0.1μF
C4, 5 2.2μF aluminum electrolytic
C6, 9, 13, 14, 24 100pF
C7, 12, 15 0.001μF
C16 10pF
C19, 21 220μF aluminum electrolytic
C22, 23 0.01μF
[caps not individually identified on schematic: 4×0.01μF]

Semiconductors
D1 1N4001
D2, 3, 6, 7, 8, 9, 10 1N914
D4, 5 red LED, forward drop ~1.8V
IC1 TL071 op amp
IC2 NE570 dynamic controller
IC3, 6, 7 TL064 quad op amp
IC4, 5 TL062 dual op amp
IC8 78L05 5V positive regulator
IC9-c/d ½TL074 quad op amp
Q1 2N3906 PNP transistor
Q2 2N3904 NPN transistor
[semiconductors not individually identified on schematic: 2×LM13600, 1×TL072, ½IC9-a/-b TL074]

Miscellaneous
S1 DPDT switch
S2 SPDT switch
wire, circuit board, solder, ¼" jacks, knobs, etc.

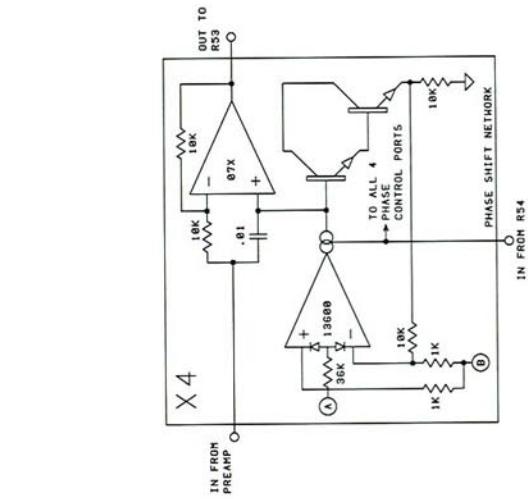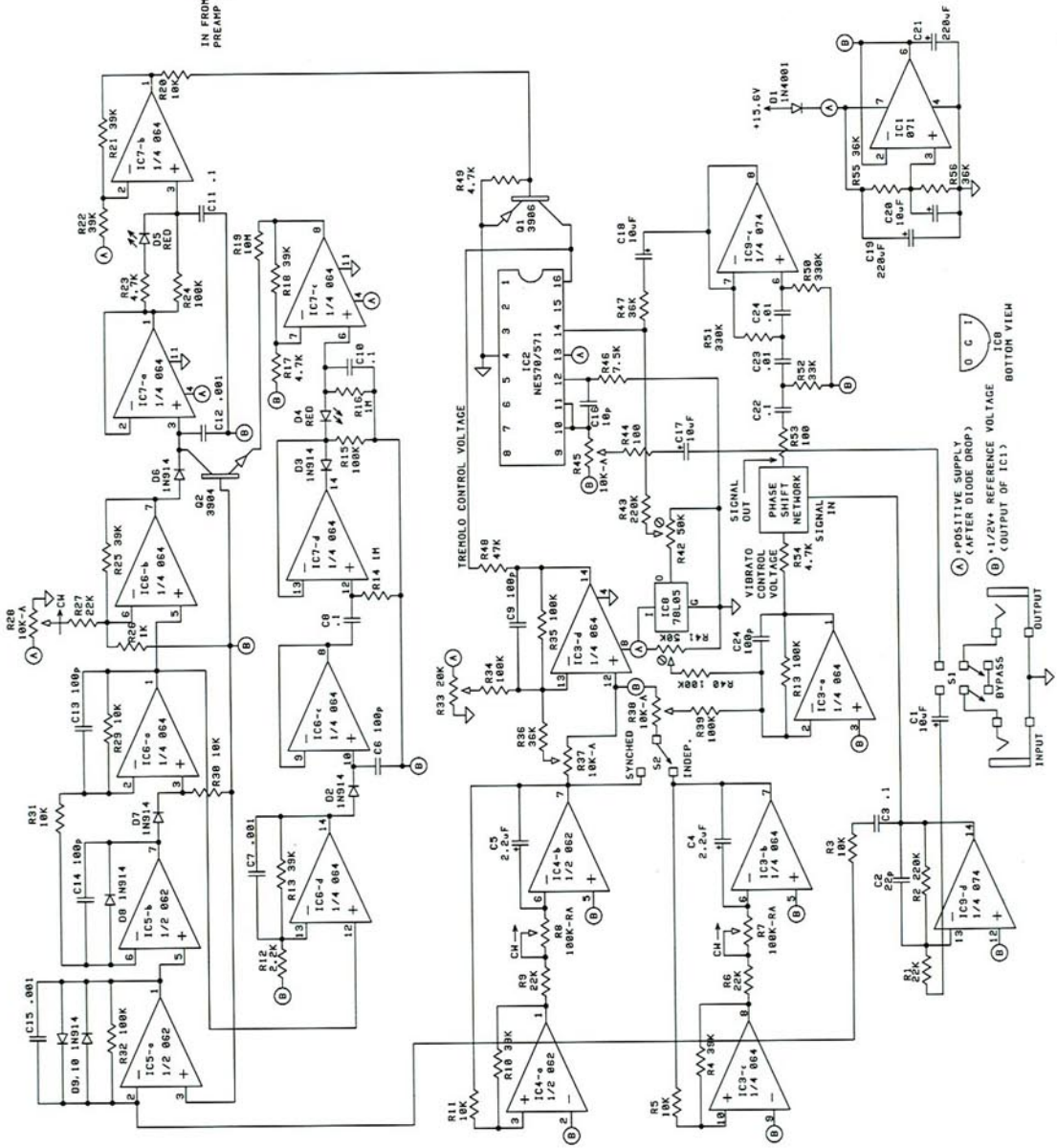Fig. 36–1. Tremolo-Matic V schematic.
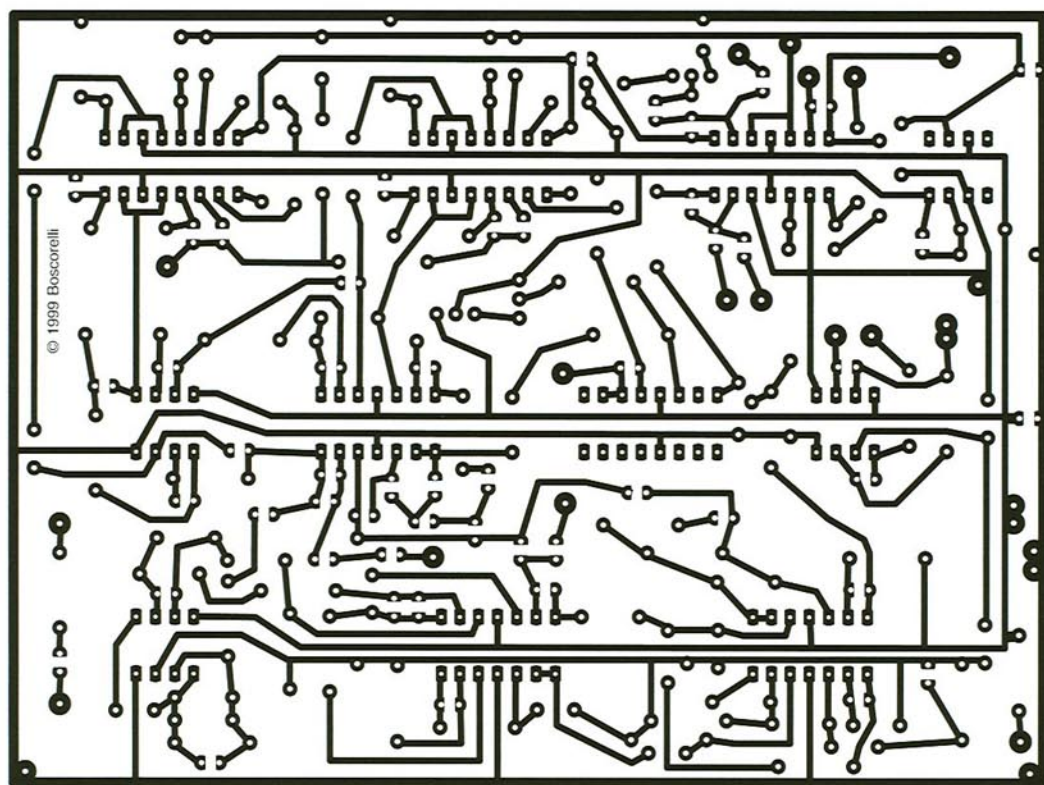
Fig. 36–2. Tremolo-Matic V circuit board.

Use

Switches & pots have these functions:

S1	effect/bypass
S2	vibrato control select synched/independent
R7	vibrato rate (independent mode)
R8	tremolo rate
R28	noise gate threshold
R33	VCA static gain
R37	tremolo depth
R38	vibrato depth
R41	vibrato static point trim
R42	VCA feedthrough trim
R45	output level

First, trim the vibrato static point. Turn R38 fully CCW, attach voltmeter or scope probe to pin 1 of IC3. Trim R41 to give 2 volts.

Next, trim VCA control feedthrough. Center R33, turn R37 fully CW. In this state the tremolo control voltage clips at both extremes. Connect scope probe to IC2 output pin 10; trim R42 for minimum feedthrough. If no scope is available, trim feedthrough by ear, as detailed in connection with prior Tremolo-Matics.

Initial settings: S1 effect, S2 synched; R7, R8, R28 fully CW; R33, R45 straight up; R37, R38 fully CCW. In this state the box acts as a preamp with gain of ~2. Connect unit to axe and amp, establish desired listening level.

Verify the tremolo functions of rate, depth, and static point as detailed for prior Tremolo-Matics. Set tremolo depth to minimum, explore the vibrato depth & rate functions. With independent tremolo and vibrato established, test their combined sound, running synched and independently. Finally, turn R28 CCW until the desired gating action occurs.

Notes

TM5 is optimized for a 15V supply. Significant departure from this voltage is not recommended. Despite extensive use of low-power op amps, the prototype drew ~40 ma, making a 1W DC–DC converter an attractive power option (e.g., Digi-Key p/n 179-1072 or Mouser p/n 580-NME0515S). Use of non-low–power op amps will require a 2W converter.

Fig. 36–3. Tremolo-Matic V layout & wiring diagram.

Fig. 36–4. Tremolo-Matic V prototype board.

Project No. 37
Axe-O-Matic II

Some of the best-sounding mods prove to be the simplest. Case in point: Axe-O-Matic II.

Circuit Function

AOM2 consists of a phase-based tunable passive notch of the type whose function was discussed in the Appendix on Tone Control. The axe retains its stock volume pot; the notch frequency control replaces the existing tone pot.

Notes

No circuit board is required; solder the parts point to point. Be sure to ground the pot's case.

The pot tunes a notch ~20 dB deep over the approximate range 80–15,000 Hz, offering more tonal versa-

tility than one might anticipate from the nature of the function. If limited range of the stock tone control has you frustrated, definitely give AOM2 a listen.

For bass guitar, change C1 to $0.1\mu F$, change C2 to $0.0047\mu F$, and increase R1 & R2 to 10K.

AXE-O-MATIC II PARTS LIST
Resistors
R1, 2 1K
R3 250K dual audio-taper pot
Capacitors
C1 $0.047\mu F$ polypropylene or polystyrene
C2 $0.0022\mu F$ polypropylene or polystyrene
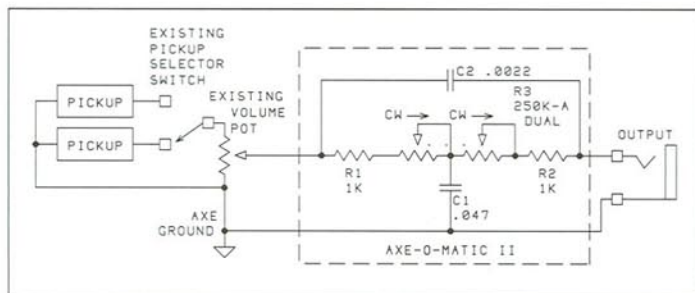**Miscellaneous**
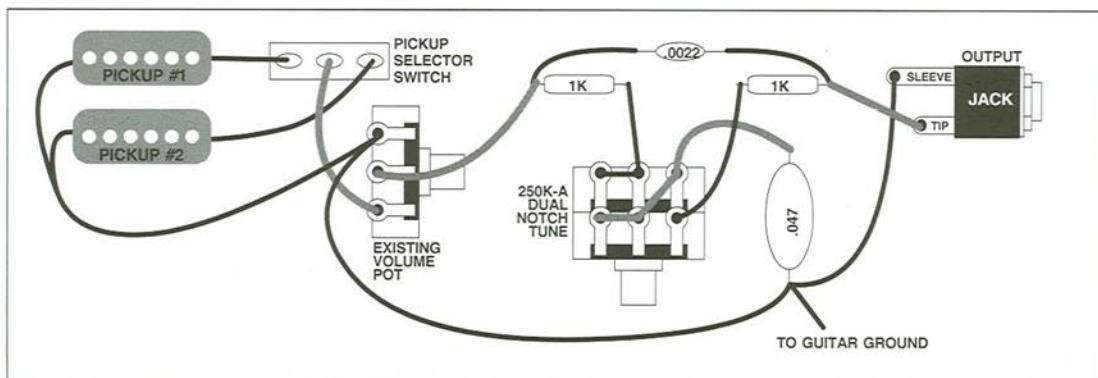wire, solder, etc.

Fig. 37–1. Axe-O-Matic II schematic.

Fig. 37–2. Axe-O-Matic II wiring diagram.

Project No. 38
Sustain-O-Matic II

Sustain-O-Matic II separates the Siamese twins of downward and upward sustain, giving the player independent control of each.

Circuit Function

Signal path: Signal feed couples through C10 to noninverting preamp IC4-b, whose gain the builder determines during construction according to the value of R26. Preamp output couples through C4-R19 to signal input of IC1, half an SSM2120 dynamic control chip. IC1 output, as a current, is taken off pin 4 and feeds current-to-voltage converter IC5-a, whose output ties to output-level pot R24, feeding voltage follower IC5-b. Signal couples through R23-C7 to the output path. The net signal path is noninverting.

Control path: One of S2's poles couples through C3-R18 to rectifier input of IC1. One throw of S2 ties to the preamp output, the other to the output of IC5-a. These define parallel and feedback control paths, respectively. (S2's other pole and one throw tie to each end of R29, such that, when the feedback control path is selected, R29 is shorted; when the parallel control path is selected, the control voltage passes through R29 in series with R30. This mitigates the exaggerated response that otherwise accompanies parallel control.) IC1's raw rectifier output is taken off pin 2, feeding IC2-a, where it's boosted by a factor of 40, and where the variable bias fed through R9-R11 creates a DC offset at IC2-a's output, which varies the sustain threshold. IC2-a's output feeds a fixed decay network made up of D3-R5-C1; and a program-dependent attack network comprised of D4-6, R6-7, and C1. The resultant control voltage is buffered by IC3-a, whose output swings negative in the absence of a positive voltage from the level detector, due to the bias applied through R4. IC3-a output couples to a variable voltage divider consisting of IC2-b, IC3-b/c/d, and associated components. This network seamlessly separates positive from negative control voltages; R1 varies the fraction of positive voltage, which controls downward sustain; R2 varies the fraction of negative voltage, which governs upward sustain. These separate voltages recombine in noninverting summing amp IC4-a, whose output ties through R29-R30 to IC1's '−' control port, pin 7.

The result of this control path is for bipolar control voltages to be generated in response to input dynamics; and for both polarities to reach the VCA control port. The polarity splitter allows independent control of upward and downward compression ratios. S2 allows use of a feedback or a feed-forward control path, and automatically varies the attenuation applied to the control voltages, to maintain comparable tracking between modes.

Trim network R21-20 exists to null the DC offset that develops at IC5-a's output when the VCA applies more than 20 dB of gain. If not nulled, this offset dis-

Fig. 37–1. Sustain-O-Matic II schematic.

Fig. 37–2. Sustain-O-Matic II prototype board.
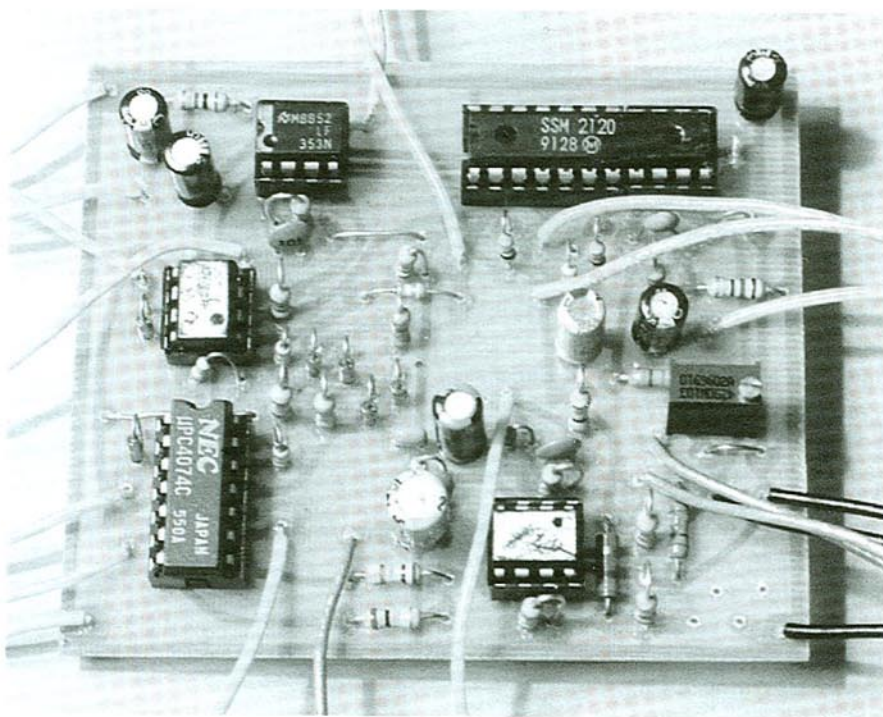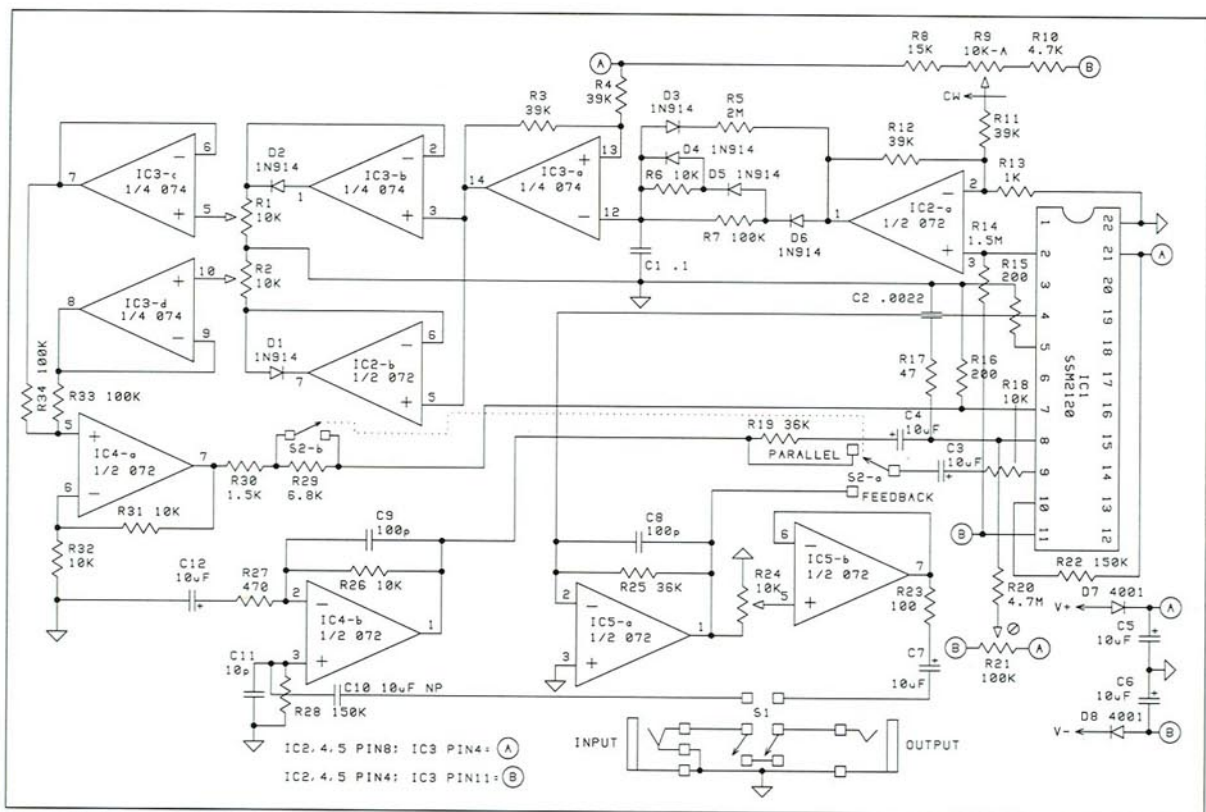
Fig. 37–3. Sustain-O-Matic II circuit board

3.75" × 3.25" reference box

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Fig. 37–4. Sustain-O-Matic II layout & wiring diagram.

ALL vertically mounted 1N914 diodes use this orientation, EXCEPT the one noted by the arrow.

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Use

Switches & pots have these functions:

R1	downward compression ratio
R2	upward compression ratio
R9	sustain threshold
R21	output DC offset trim
R24	output level
S1	sustain/bypass
S2	control path select feedback/parallel

First, trim the DC offset. Short the input, set upward sustain ratio and threshold at 1 o'clock, downward ratio and output level at minimum, control path to feedback. Set scope sweep to 20 ms/div., scale to 1V/div., DC-coupling. Connect scope probe to pin 1 of IC5, power up the circuit; trim R21 to give ground at pin 1 of IC5-a. This gives best suppression of cycling, but does not completely quell a minor downward pulse that repeats at 0.5–1 Hz. This trim is sensitive to the supply voltage.

Power down, remove scope probe, set R1, R2, & R24 fully CCW; R9 fully CW; S2 to feedback. Connect unit to signal source and target output device; slowly advance R24 to set desired listening level. In this state the box acts as a preamp, with no effect on dynamics.

Turn R1 fully CW to give maximum downward compression ratio; slowly turn R9 CCW until obvious downward compression is noted.

Now turn R1 to fully CCW, which lowers downward compression ratio to 1:1. Using material suitable to detect upward compression, slowly advance R2 until the effect is noted.

Next, explore the effect of both modes operating simultaneously.

Return controls to initial settings, switch S2 to 'par-allel,' and repeat the checkout sequence.

Notes

SM2 is optimized for a regulated ±15 supply. The prototype ran off a discrete version of the Append-A-Board Power supply, and for that reason does not include rectifier diodes. If you elect to run the box off batteries, install the polarity protection diodes shown on the wiring diagram.

Appropriate preamp gain depends on whether the box treats instrument-level or line-level feeds. The stock 26 dB gain suits instrument-level inputs. Taking advantage of this box's large headroom calls for an average preamp output of at least $5V_{p-p}$. This leaves room for peaks, while prolonging downward sustain.

SM2 produces very natural-sounding sustain, particularly in feedback control mode, when the upward and downward compression are applied with restraint. For natural sound, keep R1 at or below 50%; R2 at or below 60%. Higher settings produce longer sustain, at the cost of increasingly audible artifacts. That option was retained for players who prefer a conspicuous sound.

The 2M value of R5 sets decay at a rate that produces low distortion down to the guitar's lower limit. Increase R5 to 4.7M for bass; reduce R5 to 1M for extra-fast onset of upward sustain.

High preamp gain coupled with more than 20 dB of boost in the VCA in parallel mode predisposes to instability. When upward compression beyond 50% is selected, and when not playing, either turn the axe's volume all the way down, or unplug it from the box. Intermediate settings of axe volume leave a high-impedance node tied to the input through a long path, which may result in instability.

Troubleshooting

Careful construction technique is something you could apply to a Heathkit®. Building projects from scratch demands anticipatory troubleshooting, an active search for errors at each phase of construction. Those who built one of the now-defunct Heathkits probably recall how much fun the process gave them. Building a stomp box can do the same if the builder treats it as recreation and sets a leisurely pace.

The first chance to err comes in transferring the circuit pattern to copper-clad board; error modes depend on the method. Photomethods tend to suffer mechanical errors: toner fails to stick to the board; the photoresist missed a swatch at the edge; and so forth. These dictate manual repairs at which photomethod-users have to be adept. Patterns transferred as crepe tape and rub-on pads present many chances for application error, minimized by careful technique. Take your time, divide the work into blocks, do not work to the point of fatigue. Carefully compare traces against the pattern before etching.

Etching errors occur when wax-based transfers lift off, especially in hot etchant. Crepe tape can lift at the end and reattach to an adjacent pad. The resultant unsuspected connection may be fiendishly difficult to spot. If a floating end is seen early in the etch, cease etching and rinse the board. Press the loose end back on the correct pad, reinforce the connection with rub-on etch resist, then continue the etch.

Manually transferred patterns benefit from a post-etch comparison against the circuit board pattern. Errors that escaped pre-etch scrutiny often stand out once the board is etched.

The importance of inspecting the etched board under magnification cannot be overstressed. Look for undercutting, in which traces have been notched or severed by etchant; and incomplete etching, in which copper bridges remain between pads or traces. If traces run at all close, as they do in many projects in this book, run a continuity test on adjacent traces. Bridge undercut traces with wire and solder; break bridges between traces.

Soldering presents countless opportunities for error, reduced by a methodical approach. First, install parts in like groups. Start with IC sockets and flat-mounted resistors. Follow with vertically mounted resistors, then vertically mounted caps, then miscellaneous parts, such as transistors, regulators, and trimpots. At the start of each step, verify that the right parts sit in the right holes, and that the parts are correctly oriented. After soldering each group and clipping their leads, inspect the board for solder bridges.

Most bridges are obvious, but some lurk in the flux between closely spaced pads. If in doubt, test continuity to verify that the pads do not connect. Before soldering wires to the fully stuffed board, make a last visual comparison against the layout diagram. Then solder wires to board, pots, and jacks.

Double-check the polarity of power diodes and bypass caps. Tantalum bypass caps pose a particular hazard in that reversed wiring usually causes them to short when powered up. Also, verify that ICs' power pins match the correct power bus. Incorrect socketing is a common mistake.

Once all pots are connected and chips socketed, *run an ohmic check on the power leads before powering up the board*. Connect a DMM set to a 20K or 100K scale across V+ and GND, and V− and GND in dual-supply circuits. Typical readings fall in the range 2000–15,000 ohms, depending on the nature of the circuit and on which meter lead connects to which power lead. The reading first fluctuates, then stabilizes as bypass capacitors charge. If no reading is obtained with either polarity, short the polarity protection diode(s). The author did not observe a stable reading below 1800 ohms on any of the projects in this book. Do not power up the board if you get a reading below 500 ohms. A potentially serious error requires correction. This test catches some disastrous errors, such as a shorted power bus. It will not detect incorrectly mounted parts unless they result in a low ohmic reading.

Perform the initial power-up prepared to cut power instantly should you note hissing, smoking, or hot parts or wires. These indicate a major error that prior steps were meant to catch.

If the board passes all tests but fails to work—relax. You're going to find out why. Frustration breeds haste that practically guarantees error while troubleshooting. A major screw-up, including one that dictates starting anew, isn't the end of the world. Experienced builders gained much of their experience through error. Take a break if frustration intrudes. Return to the problem with a clean perspective.

First, check battery voltage. Depending on the circuit's current drain, batteries that read good can fade under load. Measure the voltage under load.

Next, check voltage at all circuit points having a known, constant voltage, such as chips' power supply pins. If you find that a certain pin is not getting voltage, trace the voltage at all accessible circuit points between the battery terminal and that pin. If a circuit point reads the wrong voltage, confirm that the supply

R21 2M-RA
CW →
R20 150K
D2,3 1N914
R19 47K
R14 10K (A)
R22 2.2K
R24 2.2K
R17 2.2K
R18 200
R15 36K
R12 100K
R13 100K
R11 47K
IC2-a 1/4 064
2
1
6
IC2-b 1/4 064
7
5
C1 1uF
C2 1uF
R16 10K
IC2-c 1/4 064
9
8
10
(B)
8 7 6 5 4 3 2 1
IC4 NE570/571
9 10 11 12 13 14 15 16
D1 4001
+9V ← (A)
R1 36K
C8,9 10uF
13 4
IC2-d 1/4 064 14 (B)
R2 36K
12 11
R9 10K-A
C3 10p (A)
R10 100
R7 4.7K (B)
R8 36K
R6 220K (A)
R3 22K
C6 47p
R4 250K
C5 1uF
IC1 MC33171
2
7 (A)
6
3
4
C7 10uF
BYPASS
S1
C4 10uF
R5 50K
IC3 78L05
INPUT
OUTPUT

(A) = V+ AFTER DIODE DROP
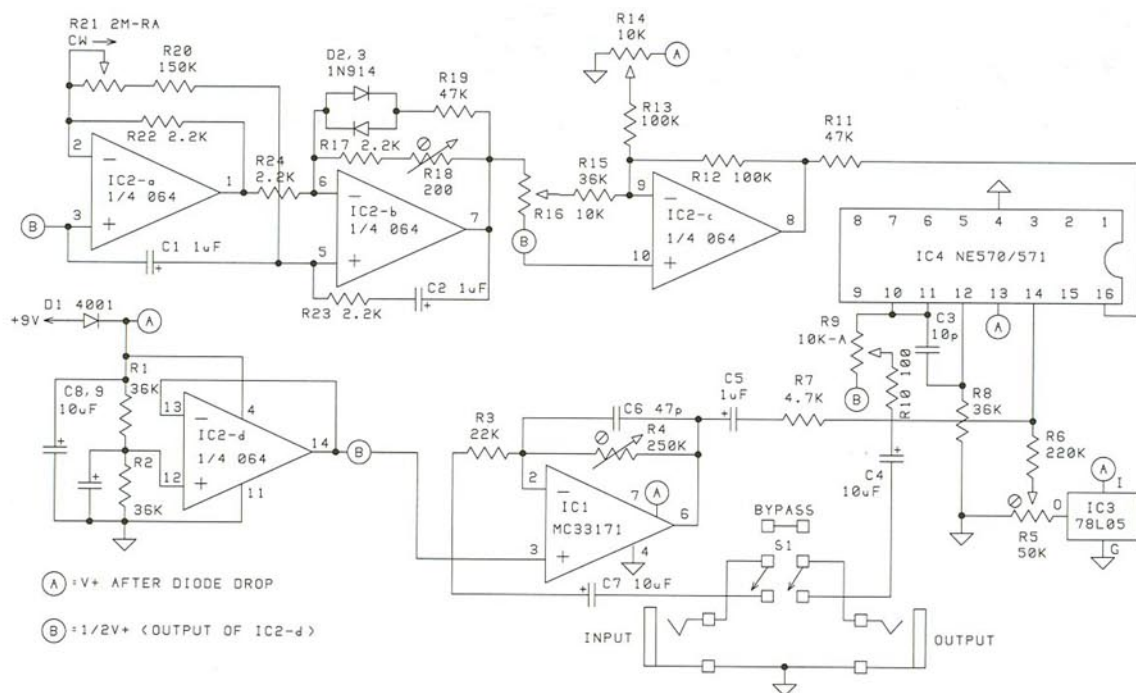
(B) = 1/2V+ (OUTPUT OF IC2-d)

Fig. A178. As an example of signal tracing, take Tremolo-Matic. Problem: no output. Procedure: assuming voltages checked good, apply a 1V 1 KHz sinewave to the input. Set oscilloscope for DC coupling, to display DC offsets as well as AC signals.

First, place scope probe on output pin of IC1. Signal should be inverted; amplitude depends on setting of R4. A DC offset equal to $\frac{1}{2}$V+ should be present at IC1's output. Assuming IC1 is good, trim R4 for unity gain.

Next, check to see that the signal is getting into the VCA, IC4. Touch probe to IC4 pin 14. If no signal registers, the problem lies in the C5-R7 path. The amplitude at pin 14 may be reduced due to divider action between R7 and the resistor inside the chip at pin 14; the DC offset at this pin is ~1.8V, which comes from the 570's internal voltage reference, and explains the orientation of C5, since IC1 pin 6 exists well above 1.8V.

If no signal is present at IC4 pin 10 (DC offset approximately $\frac{1}{2}$V+ in a 9V system), it's logical to assume that no control voltage is getting into the VCA control port, pin 16. Place scope probe on IC2 pin 8. Turn R16 fully CCW to eliminate the sine component. The static DC voltage should respond as R14 is taken through its range. If it does not, check voltage at the wiper and the ends of R14, and check the integrity of R13's connections at both ends. If the correct voltage is entering R11 but there is no VCA output, either the voltage isn't getting through R11, or IC4 is bad (or IC4 pin 4 isn't making contact with system ground, a problem that voltage tests don't detect).

If the signal is present at IC4 pin 10 but not at the output jack, you've narrowed the problem to the path that includes R9, R10, C4, & S1. Place scope probe on all junctures between pin 10 and the output jack to locate the break.

If the problem happened to be a good static signal but no tremolo, you'd suspect that the sine oscillator wasn't working, or that the sinewave was being interrupted somewhere between IC2 pin 7 and IC4 pin 16. If the sinewave is present at IC2 pin 7, stepwise tracing with the scope will reveal the interruption point. If the sine generator isn't working, check the trim state of R18, and verify that the R21-R20 path is not an open circuit.

These steps diagnose the problem in 99% of cases, though sometimes they turn up errors that should have been caught earlier, such as omitted circuit traces or shorted pads, or an improperly socketed chip.

is not at fault, then trace the voltage from the supply to the point that reads wrong. Location of the departure should reveal the problem.

Check to see that variable voltages vary with adjustment of the appropriate pot. Verify that DC offsets in the signal path fall within spec. If seemingly bizarre DC offsets are found in the signal path, and if the signal path uses polarized coupling caps, suspect reversed orientation of the coupling caps.

A meter takes this process just so far. Serious troubleshooting demands an oscilloscope. Assuming the supply voltages checked good, trace the signal path. Feed a 1 KHz sinewave to the input, trace it through the signal path. Tracing means viewing the signal in sequence, at every accessible circuit point from input to output (some prefer to trace from output to input). Note whether the signal possesses proper shape, amplitude, and DC offset.

The schematic tells what you should see at each point. Fig. A178 uses Tremolo-Matic as an example.

Faulty parts are rare enough not to justify a working assumption. You can swap out socketed chips, but desoldering parts without evidence that they're bad should be a last resort.

Troubleshooting is simple, logical, stepwise; but the better alternative to is to take enough care building the box that it works straight away.

Stomp Box Glossary

~ approximately

aliasing generation of artifacts by a digital or quasi-digital system attempting to process a signal whose frequency exceeds half the clock frequency

bleeder resistor a high-value resistor tied between the floating end of a coupling capacitor and ground, whose purpose is to prevent accumulation of a charge in the cap

BPF bandpass filter; sometimes shortened to *BP*

C short for capacitor

CCW counterclockwise

class A amplifier bias class in which the stage conducts 100% of the input cycle

class A/B amplifier bias class in which the stage conducts >50% but <100% of the input cycle

class B amplifier bias class in which the stage conducts 50% of the input cycle

class C amplifier bias class in which the stage conducts <50% of the input cycle

CMRR common mode rejection ratio

corner short for corner frequency; when speaking of a resistor in series with a capacitor, or in parallel with a capacitor, the corner is the frequency at which the cap's reactance, X_C, equals the resistance; corresponds to the point at which response of a lowpass or highpass network is 3 dB down

CW clockwise

cycling a concomitant of dynamic systems that use a feedback control loop; in limiters, fast attack and decay cause gain reduction to cycle on and off at rates in the KHz; in upward sustainers, cycling results when VCA gain rises to the point that system noise or some minor transient generates a control voltage that flips the VCA into unity or low-gain mode, often accompanied by an audible pop

de-esser sibilance reducer

DIP dual in-line package, the name of standard-size integrated circuit packages

direct box a circuit that converts an axe's high-impedance unbalanced signal to a low-impedance balanced one, to enable the signal to feed line-level gear directly without suffering loading losses; might or might not include gain; can be built from active circuitry or from a transformer

DPDT switch nomenclature: double pole double throw

dry referring to a musical instrument effect, the sound of the instrument before alteration by the effect; see *wet*

duty cycle referring generally to squarewave trains, the percentage of time spent high relative to the time spent low

dynamic tracking in loose usage, the extent to which the dynamics of an effect's output follow those of the input

envelope short for music envelope, the low-frequency DC that results from fullwave rectification of audio, followed by integration; the quantity obtained gives an index of music dynamics

feedthrough the fraction of control voltage applied to a voltage controlled block that appears at the output along with the signal

FET field-effect transistor; short for JFET (junction FET)

f_{-3} corner frequency

GND ground

h_{FE} loosely, the current gain of a transistor, specified in data sheets and in some parts catalogs; can be measured by many digital multimeters; used in the calculation of I_C that gives minimum total noise of a preamp mated to a given source impedance

Hz Hertz; cycles per second

IHF Institute of High Fidelity

IC integrated circuit

I_C collector current; the current flowing through the collector resistor of a bipolar transistor

I_D drain current; the current flowing through the drain resistor of a field-effect transistor

instrument-level referring to audio, the raw output of an electric guitar or bass; between microphone-level and line-level; usually more than 50 mv_{p-p} but less than $1V_{p-p}$

I/O input/output

ips inches per second (tape speed)

I–V current-to-voltage

K kilohm; 1000 ohms

KHz kilohertz; 1000 cycles per second

knee the point and manner in which a compressor shifts from linear transfer to compression-law transfer

LC inductor-capacitor

LDR light-dependent resistor, a/k/a cadmium sulfide or cadmium selenide photocell

LED light-emitting diode

line-level referring to an audio signal, at least 200 mv_{p-p} but often $>1V_{p-p}$

LP long-playing vinyl phonograph record

M megohm; 1,000,000 ohms

ma milliamps; 10^{-3} amps; a common unit of current

MHz megahertz; 1,000,000 Hertz

microphone-level referring to an audio signal, the output of an unpowered microphone, typically no more than 10 mv_{p-p} and requiring >30 dB of preamp gain

MOSFET metallic oxide surface field effect transistor

mv millivolts

NAB National Association of Broadcasters

nm nanometer, one billionth of a meter; unit of length, commonly used to designate wavelengths of light

OTA operational transconductance amplifier

optocoupler an electrical component in which a light source, usually an LED, is potted with a phototransistor or a photoresistor

period $1 \div \text{frequency}$

pF picofarad; 10^{-12} farad; a common unit of capacitance

p–p peak-to-peak, as in volts measured peak-to-peak (V_{p-p})

R resistor

rail splitting the practice of converting a single power supply voltage into a dual supply, by creation of a reference point halfway between V+ and ground; the reference becomes the new ground, the existing ground becomes V–

rail to rail extending from the positive supply to the negative supply

RC resistor-capacitor

RIAA Record Industry Association of America

RMS (or rms) root mean square, as in volts RMS (V_{rms})

sag a level-dependent fall in volume caused by a drop in tube bias voltage; occurs in some tube amps when driven hard

scope oscilloscope

S/N signal-to-noise (ratio)

snubber a resistor-capacitor network that suppresses ultrasonic oscillation of a circuit

SVF state variable filter; an active filter built from op amps, resistors, and capacitors, which directly provides highpass, lowpass, and bandpass outputs, and from which a band-reject (notch) can also be derived; frequency and bandwidth are readily tuned; commonly used to realize parametric EQ

SVR sliding or stepwise variable ratio, the formal definition of *soft-knee compression*

tremolo an effect produced by rhythmic amplitude modulation, usually at a rate between 1 and 5 Hz

μa microamp; 10^{-6} amp; a common unit of current

μF microfarad; 10^{-6} farad; a common unit of capacitance

UL Underwriters Laboratories

varistor an electronic component that exhibits very high resistance in the presence of low voltage, very low resistance above a threshold which ranges from 4V to several hundred volts, depending on type; varistors are usually used to protect electronic gear from spikes in the AC line voltage, but were adapted to phase-shift circuits in Magnatone amplifiers

VCA voltage controlled amplifier

VCF voltage controlled filter

vibrato in a stomp box or amp, the sound of rhythmically shifted phase, which manifests as shifting pitch only while phase is changing; also called *pitch vibrato*

V_{p-p} volts peak-to-peak

V+ the positive power supply

V– the negative power supply

V–I voltage-to-current

wet referring to a musical instrument effect, the sound of the instrument after alteration by the effect; see *dry*

X_L inductive reactance

X_C capacitive reactance

Z impedance

Symbol Legend

	SCHEMATIC	LAYOUT		SCHEMATIC	LAYOUT

FIXED RESISTOR

horizontal   vertical

SPST SWITCH

VARIABLE RESISTORS (PANEL POTS)

CW →

SPDT SWITCH

DPDT SWITCH

VARIABLE RESISTORS (TRIMPOTS)

single-turn

multiturn

2-COND. JACK

TIP

SLEEVE

JACK

SHORTING JACK

TIP

SHRT

SLEEVE

TIP

SHORT

SLEEVE

JACK

CAPACITOR (NONPOLAR, CERAMIC, ETC.)

CAPACITOR (TANTALUM, ALUMINUM)

vertical

3-COND. JACK

TIP

RING

SLEEVE

TIP

RING

SLEEVE

JACK

SMALL SIG. DIODE

anode   cathode

NPN TRANSISTOR

C

B

E

C

B   top view

E

RECTIFIER DIODE

ANODE   CATHODE

PNP TRANSISTOR

C

B

E

C

B   top view

E

LIGHT-EMITTING DIODE

top view   bottom view

flat=cathode

N-CHANNEL FIELD EFFECT TRANSISTOR

D

G

S

G

S   top view

D

TRANSFORMER

Parts Sources

DC Electronics, Box 3203, Scottsdale, AZ, 85271. Parts include: NE570, XR2206, MC3317X op amps, FET-input op amps, Mouser pots & transformers. Free catalog.

Digi-Key, Box 677, Thief River Falls, MN, 56701. Carries Panasonic BBD chips & databook, OP-27, 1W DC–DC converters; most of the ICs made by Maxim, Linear Technology, and National Semiconductor; pots, jacks, switches, and many other parts. Free catalog.

Groove Tubes Electronics, 12866 Foothill Blvd., Sylmar, CA, 91342. Sells tube amps and tube mics; plus a large selection of screened and graded triodes & pentodes; also carries *The Tube Amp Book*. Free catalog.

Hosfelt Electronics, 2700 Sunset Blvd, Steubenville, OH, 43952. Carries 2N2646, Vactec VTL2C2, cheaply priced CdS photocells for building custom optocouplers; XLR hardware; large selection of wall warts. Free catalog.

Jameco Electronics, 1355 Shoreway Road, Belmont, CA, 94002. Carries 40-KHz ultrasonic transducers, NE570, AD536A, OP-27, AD633 analog multiplier, large semiconductor selection. Free catalog.

Mouser Electronics, 2401 Hwy 287 N., Mansfield, TX, 76063. Inexpensive audio transformers, 1W & 2W DC–DC converters, XLR hardware, pots, semiconductors; Clairex CdS and CdSe photocells. Free catalog.

PAiA Electronics, 3200 Teakwood Lane, Edmond, OK, 73013. Carries a stomp-box case w/six punched holes for pots & jacks and a larger hole for the stomp switch; SSM2120; kits to build stomp-box effects & rackmount processors; books. Free catalog.

Parts Express, 725 Pleasant Valley Dr., Springboro, OH, 45066. Sells Accutronics reverb tanks, semiconductors, XLR jacks, studio cables, Celestion guitar speakers, 2N2646, triodes, pentodes, tube sockets, tube-amp transformers, old tube-amp manuals, *The Tube Amp Book*; rich complement of guitar-amp-related parts. Free catalog.

Stewart-MacDonald's Guitar Shop Supply, Box 900, Athens, OH, 45701. Carries a heavy-duty DPDT stomp-box–style switch; also pull-switch-equipped pots; single & dual 9V battery compartments designed to mount in the body of a guitar but adaptable to stomp boxes; large selection of books, pickups, guitar construction & repair supplies. Free catalog.

Torres Engineering, 1630 Palm Ave, San Mateo, CA, 94402. Vast selection of tube-amp parts and modification kits; short-spring and long-spring reverb pans; *Inside Tube Amps*. Free catalog.

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