TAPE NOISE LIMITER

Takes the hiss out of the quiet bits of your music, and does in a way which is simple yet effective, and is a replay only process so it will work on any tape!

DESPITE the small size, the performance obtainable from a cassette tape in a good recording deck is quite remarkable. In fact the latest top quality decks are so good that it is difficult to tell the difference between the recording and the original sound.

Unfortunately this is not true of the cheaper units — in which 'tape hiss' can be very prominent. Tape hiss is caused by random irregularities in a tape's surface coating. The effect is common to all tapes but some are marginally worse than others.

The annoying characteristic of tape hiss delayed the acceptance of cassette tape recorders in hi-fi systems for some years — until the advent of the Dolby system which was primarily developed as a cure for the phenomenon.

The Dolby system is often misunderstood — *it only works if the cassette tape itself has been recorded using the Dolby process* — and few commercially produced tapes are. Unless the tape cassette says specifically that it is Dolby processed then it's not! You can of course record your own tapes using Dolby if you own a Dolby machine.

Upper Limit

To overcome this limitation a number of cassette recorders are fitted with noise reduction circuitry which reduces the level of hiss on non-Dolby recordings. Most of these noise reducing circuits work by progressively reducing all high frequency signals when the output level falls below a preset minimum. Above that minimum level all sounds are allowed through because tape hiss cannot be heard once the sound level is substantially louder than the hiss. This effect is called 'acoustic masking'.

The circuit described in this project is a simple but very effective unit which may be used with any cassette recorder which is connected to a hi-fi system.

The unit should preferably be connected between the cassette recorder and the amplifier input — using short lengths of screened cable and suitable connecting plugs. If you really know what you're doing it may be actually built into the tape recorder or amplifier. Alternatively it may be connected between the pre-amplifier and power amplifier on those units which are so separated (note that many apparently integral

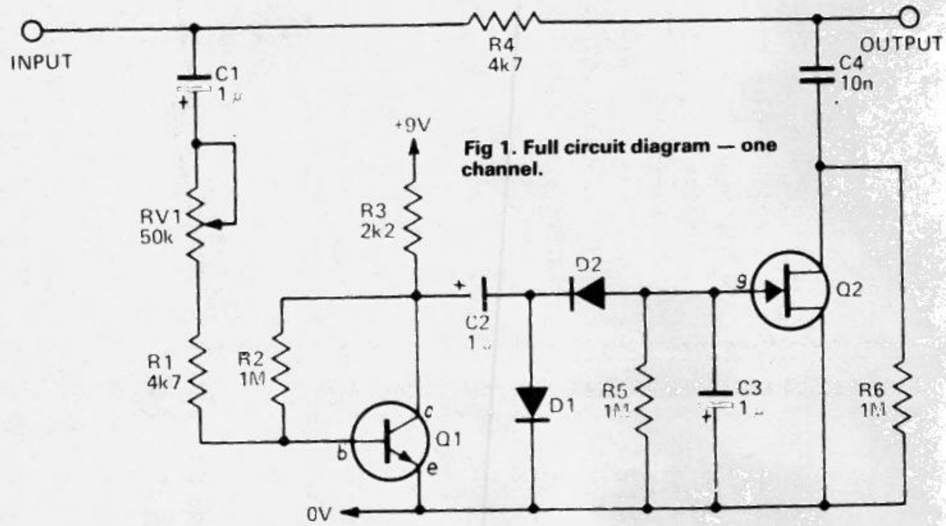


Fig 1. Full circuit diagram — one channel.

HOW IT WORKS

The circuit passes all frequencies (without attenuation) if the incoming signal is above a set minimum level. Signals below the preset minimum are progressively attenuated from 1 kHz upwards. The maximum attenuation of about 10 dB is applied at approx 10 kHz.

Resistor R4 and capacitor C4 form a filter in which Q2 is used as a variable resistor with the degree of resistance dependant on gate voltage. Thus, if the input voltage is at or near 0V then Q2 appears as a low resistance and C4 is in circuit. If on the other hand the input signal is higher than (say) four volts negative, Q2 has a very high resistance and C4 is effectively out of circuit.

The voltage applied to the gate of Q2 is that derived from Q1 — after rectification by D1 and D2. Transistor Q1 amplifies the input signal and with RV1 in minimum position, input signals above 10 mV or so will cause Q2 to be off.

Increasing RV1 raises the level below which high cut will occur. The change from full to zero cut occurs over a range of approx 5 dB input level change.

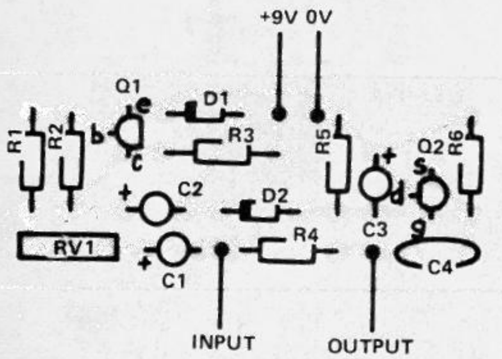
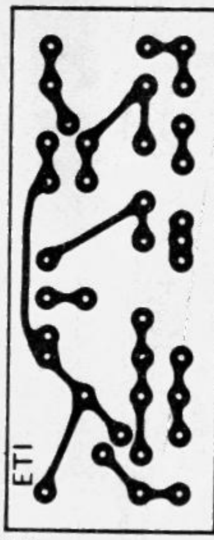


Fig 2. Above: Component overlay.



Foil pattern shown full size

PARTS LIST

RESISTORS 0.5 W 5%	
R1, 4	4k7
R2, 5, 6	1M
R3	2k2
POTENTIOMETER	
RV1	50 k trimpot
CAPACITORS	
C1-C3	1 uF 25 V
C4	10n polyester
TRANSISTORS	
Q1	BC548
Q2	2N5459
DIODE	
D1-D2	1N914
MISCELLANEOUS	
Nine volt battery and clip, PCB case	

amplifiers still have 'pre-amp out' and 'power-amp in' connectors on the rear panel. These connectors are normally bridged by 'U' shaped links — which should be removed to enable this unit to be plugged in).

Construction

As with most projects in this series you can use either Veroboard or the special printed circuit board shown here.

Take the ususal precautions about inserting components the right way round — taking particular care with the field effect transistor Q2. Note that the cathode lead of the diodes (shown as a horizontal bar on the circuit diagram) will be identified on the component by a black band or similar marking.

Unless the leads between this unit and the tape deck and amplifier are very short it is advisable to connect it via screened cable. Note that the 0V line shown on the circuit is also the 'earthy' side of the input/output connections.

To set up the unit simply choose a recording with a longish quiet passage and then adjust RV1 for the best compromise between tape hiss reduction and minimum loss of high frequency programme content.

ETI

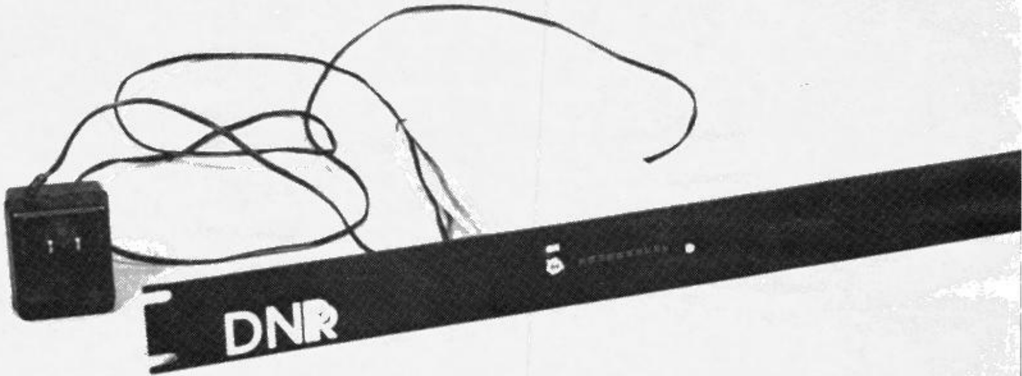
NOTE: If you listen only to hard rock — where there aren't any quiet passages — then this unit will be of little value to you. Its main effect is to reduce annoying tape hiss during otherwise quiet programme material.

Noise Reduction Unit

PROJECT

A tiny, inexpensive project from National Semiconductor that does an amazing job of reducing audio hiss without the need for encoding.

by Steve Rimmer



HUM IS a great thing in a sound system . . . it masks the hiss. After a while one inevitably fixes all the ground loops in one's stereo and gets the hiss really good and audible. Sadly, hiss is not as easily removed.

Hiss is caused by random energy floating around in circuitry. In vacuum tube equipment this is produced by electrons smashing into the metal anodes of the tubes. Semiconductors produce their own variety through the action of silicon junctions. In addition to this, there are special varieties of noise, such as tape hiss, which comes from the particle coating of the tape itself.

The elimination of noise has long been a really well sought after quest, because it is the "noise floor" that ultimately determines the quality of the sound that can be had from an audio system. Quiet can be awe-inspiring . . . there is no better buzz than to turn on your system and hear nothing until the first bars of music swell out of the ether . . . or beep out, if you listen to a lot of new wave.

There is also nothing quite so difficult to come by.

Boo . . . Hiss

The problem with killing noise is that it is part of the signal you'd like to keep around. It occupies part of the same frequency spectrum, which means that simple filters won't trap it. You can reduce the apparent hiss level in a signal by turning down the treble, but this also sends all the high frequency energy to Baffin island and makes your music sound fairly dismal and muddy. Besides, this is ultra-low tech.

High tech being a universally good wrench for fixing things that seem intractable, much thought has been given to the premise that this apparent paradox could somehow be overcome if only enough transistors were added to the circuitry. In fact, this idea is somewhat true . . . if you cheat a bit.

While noise is, by definition, energy right across the spectrum of interest, it happens that the noise which tends to plague sound systems is high frequency energy. That's why turning down the treble mortifies it. Furthermore, it will be noticed that high frequency hiss is only audible when there is no periodic high frequency energy to cover it . . . that is, you only hear it during the bass solos. High tech solutions are born of observations like this one.

To begin with, most of the circuitry involved in the accidental production of noise can be viewed as being signal processors of some sort. Tape recorders take sound and delay it. FM radio transmitters and receivers move it over distances. Spring lines add reverberation, and so on. Any processor can be viewed as a box with an input and an output.

The box itself produces an effect . . . and adds some noise.

The first effective noise reduction concepts involved double ended signal processing. This meant that one altered the signal before it went into the processor in some way so as to make it less susceptible to having noise attached to it, and then "de-altered" it once it came out to make it sound the way it was supposed to.

The most successful double ended processor is, of course, the mystical Dolby box. It uses a technique called compression-expansion on the high end of the audio spectrum . . . where the noise lives. It looks at the treble bandwidth and decides if there's any sound energy up there. If it turns out that things are pretty quiet at the moment, it expands the high end, that is, turns up the gain. This expanded signal gets fed to the signal processor . . . tape recorder, transmitter, or whatever . . . and, upon emerging, ready to be listened to, enters a box that reverses the process. On quiet passages it compresses the high end . . . turns down the gain . . . so as to effectively reduce the level of the noise.

The Dolby process, and other similar approaches, have several fundamental limitations. The most obvious one is that you have to have both ends of the thing happening for the system to do any good. If you have a noisy tape which was not recorded with the Dolby switched on, you have a permanently noisy tape.

Secondly, sound which is encoded under a double ended noise reduction process has to be played back through equipment capable of handling the other half of the process, or it sounds pretty weird in places.

Finally, all of the common noise reduction processes are designed around custom chips which can't be obtained by the average dude. They like circuitry which is excruciatingly critical . . . in short, you just can't build a Dolby system from scratch.

But wait, don't unplug your soldering iron yet!

Psychoacoustic and other Hitchcocks

As noted a few paragraphs ago, you only hear the noise when there is nothing happening in the high end of the spectrum to cover it up. This is the basis of the Dolby system. However, there is another way to use this effect, one of which, as we'll see, leads to a pretty decent single ended noise reduction system.

To grossly oversimplify things, one approach to reducing the high frequency noise in a sound system utilizes a graphic equalizer and a friend who has unspeakably fast reflexes. The guy with the reflexes sits beside the equalizer wearing headphones and, should he detect that there is nothing much happening in the high end, immediately hammers all the high end faders down. If something of a more treble disposition appears, he turns 'em back up.

You could, of course, produce some sort of circuit to do this if you didn't have a friend to hang around and mind your stereo every time you felt like checking out the Grateful Dead. In fact, such circuits do exist. They're called DNRs, or Dynamic Noise Reducers.

A DNR is a voltage controlled low pass filter and a spectral content detector. The detector watches the signal and generates a voltage which is proportional to the highest instantaneous frequency at which there is any meaningful level of periodic (sound) energy happening. It holds the filter open enough so that the highest sounds can pass through unattenuated, and the noise gets chopped off.

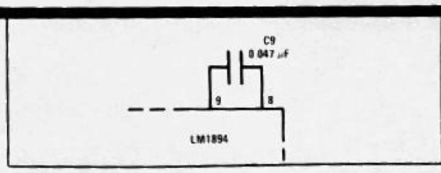


Fig. 2. If the multiplex filter is omitted, connect C9 across pins 8 and 9 as shown. Illustrations are courtesy of National Semiconductor.

The result of this is an application of a technique called psychoacoustic masking. In fact, it doesn't do anything for the sound from an electrical standpoint, but it does make things appear a lot more karmic to your ears. It stems from the observation that one sound can effectively "mask" another if the masking tone is sufficiently loud and removed spectrally from the tone to be masked. This also works with noise.

If you take a tone and add varying amounts of noise to it, it will happen that the tone will appear to vanish when the total energy of the noise exceeds that of the tone. It thus follows that if you increase the bandwidth of the noise that the overall amplitude can be smaller.

Of course, we're trying to mask noise with tones, rather than the other way around, but this is a trivial detail.

If you reduce the bandwidth of the noise that's masking the tone, with its amplitude remaining constant, then, you reduce the total noise energy, and, hence, its capacity for masking the tone. This can be interpreted as meaning that the tone becomes more audible as you reduce the bandwidth of the noise. This is effectively what the moving filter in the DNR box is for.

It is also useful to note that the ear cannot perceive distortion that lasts for less than a millisecond or so. Thus, the rapid swings of the filter one would expect to occur with complex music do gorch up the sound quite a lot . . . but your ear will never know it, because it all happens too fast.

Dynamic Tones

While DNR systems can be built from scratch, there is some fairly complex circuitry involved, and it is certainly not an easy thing to get one of them working properly unless you are really plagued with a lot of test equipment. Fortunately, it also isn't terribly necessary. Just recently, National Semiconductor turned up a specialized DNR chip, the LM1894. This little troll provides two channels of really fancy dynamic noise reduction with tight specifications, no weird parts and really simple external circuitry.

The circuit in figure 1 is a complete DNR stereo system built around a single LM1894. It produces 10 dB of effective noise reduction, although the apparent effect is rather greater. It runs on any supply voltage from 4.5 to 18 volts and has no critical adjustments. The single trimmer can be turned to its middle setting and left there for most applications. The circuit can be built on either Vero board or the PCB provided.

Figure 3 is an optional bandwidth meter. It uses a standard bargraph chip, the LM3915, and ten LEDs to show you what the instantaneous bandwidth of the filter is. It takes a bit of getting used to, because it does not show you the level of the signal.

In fact, there is no practical reason for having the bandwidth meter in the circuit at all. It can be left out entirely . . . unless you dig flashing displays.

The optional parts for the DNR circuit itself are only required if you intend to apply the box to an FM tuner. Many of these things don't filter out the 19 KHz FM pilot tone all that well and, while it would ordinarily never make it through your sound system, it will confuse the DNR box and cause it to keep its filter wide open all the time.

The DNR system can be built into any case you fancy . . . it's pretty tiny . . . or directly inside another piece of equipment. The prototype was done with a Hammond chassis box bolted to a 19 inch rack mount plate. This is a bit of a waste of space . . . you don't need anything like 19 inches to put the thing in . . . but it does keep the studio looking reasonably neat.

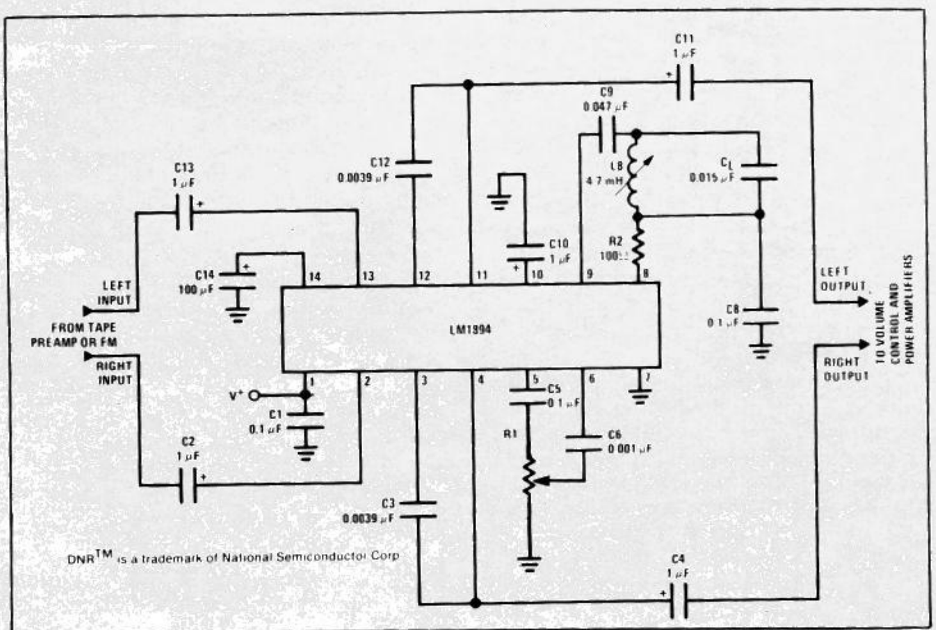


Fig. 1. The schematic diagram of the unit. L8, CL, C8, and R2 may be omitted if the FM multiplex filter is not required. See Fig. 2. ETI would like to thank Larry Clark of Canadian Micro Sales for his assistance in preparing this project.

Plugging In

The DNR box should be patched into your system before the tone and volume controls so that it gets a line level signal unaffected by treble setting. It is remarkably forgiving of volume settings, and

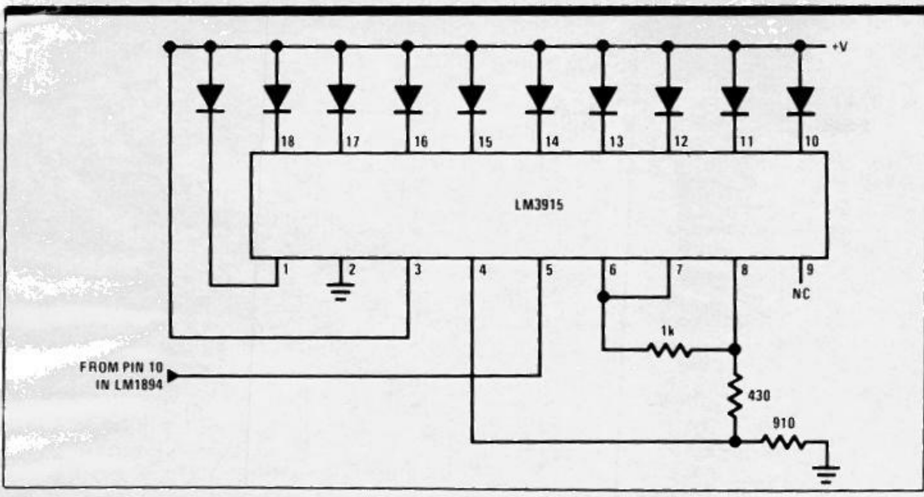


Fig. 3. The LED display circuit which displays the bandwidth of the unit. This display can be omitted.

most systems will take to it immediately. It will go a bit funny if you try hooking it in before your phono preamp or after the power amp, but most anywhere else should produce some sort of acceptable results.

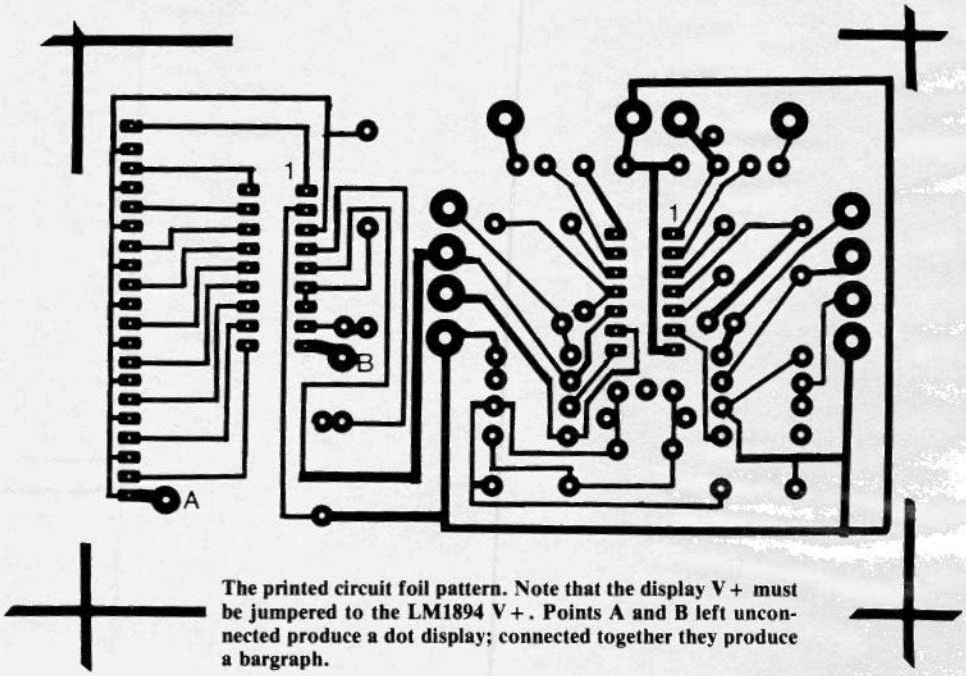
There are, obviously, no operating controls to fiddle with, with the exception of the power switch. The effectiveness of the DNR system is actually quite astounding . . . it doesn't pump, or do any of the unusual filter movement things one might expect. It also leaves the high end pretty well intact. However, it makes a great deal of difference in the perceivable noise.

Experiments using my 1946 Hallicrafters short wave radio and a really beat up Revox A77 tape recorder set to produce mega-tape echos (with well recycled noise bursts) have really given the DNR system a proper workout, and the sounds that emerged from it were a proper whiz . . . clean, sharp and without any unusual phasing noises.

There is only one really major caveat involved in using the DNR. It is tempting to try to cascade the two channels of the chip to obtain 20 dB of noise reduction. In fact, there is a National Semiconductor application note around that says you can do this. You can't. Well, okay, you can, but you'll be sorry. With both channels having a crack at the same sound source, the system does tend to pump, and the effect of the moving filter gets very, very noticeable.

The DNR box is a very simple and effective solution to a fairly complex problem. Despite its obviously short parts list, it functions very, very well. It's specifications are up to those of most home stereo equipment and semi-professional multitrack recording stuff.

Plus, the meter is neat to watch.



The printed circuit foil pattern. Note that the display V+ must be jumpered to the LM1894 V+. Points A and B left unconnected produce a dot display; connected together they produce a bargraph.

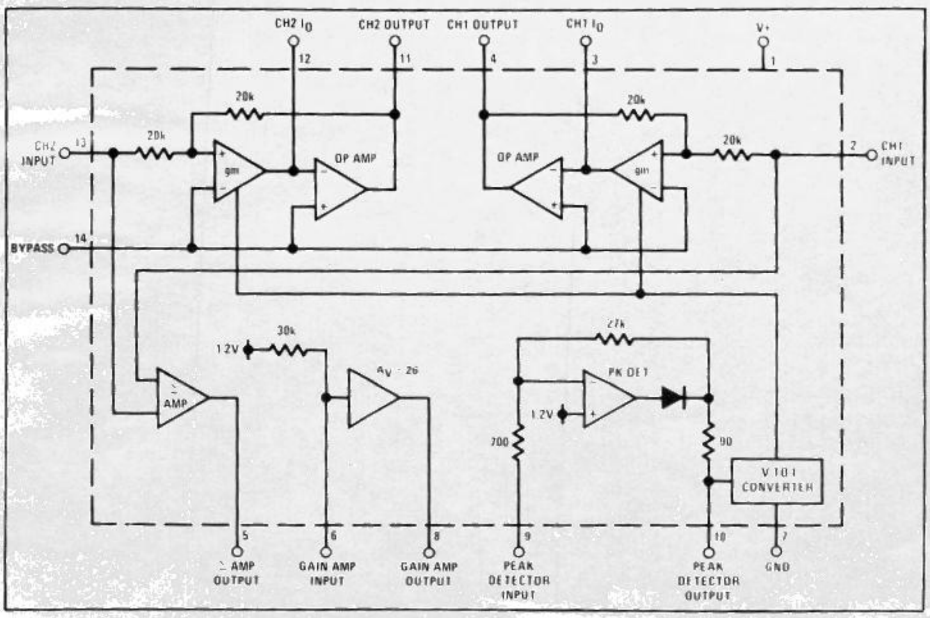


Fig. 4. The block diagram of the LM1894 noise reduction chip.

Powering Up

We used a 9 V DC plugpack from a portable radio, and no hum got through despite the output ripple. You can use one of these, from about 6 V to about 15 V, or you can tap off a DC feed from any preamp it gets installed in.

Calibration

This part is a cinch. As we mentioned, you can leave the trimmer set in its middle position and get good results. However, should you want optimum performance, National recommends playing tape noise (no signal) into the unit, and adjusting the trimmer until the first two LEDs come one. This sets the operation of the filter bandwidth threshold to suit the signal levels you'll be using.

Continued on page 22

TABLE 1

Consistent Units for Ohms's Law

V = I x R, where all units are taken from the same line of the table:

V	I	R
V (volts)	A (amperes)	Ohms
mV (millivolts)	mA (milliamps)	Ohms
V	mA	K Ohms

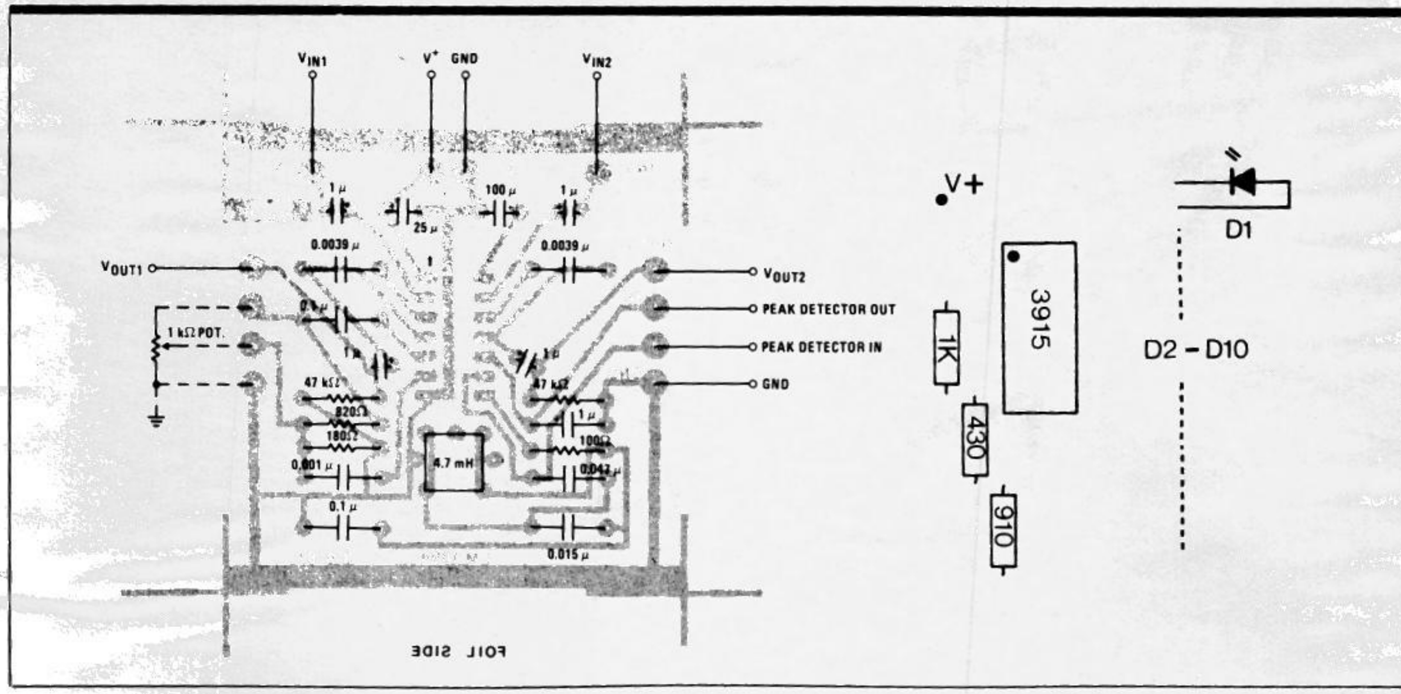
Gauge in mils	Diameter 1000 ft.	Ohms per 1000 ft.
24	20.20	25.67
25	27.90	32.37
26	15.94	40.81
27	14.20	51.47
28	12.64	64.90
29	11.26	81.83
30	10.03	103.20

REFERENCES

1. Increase Your Analog Meter's Range, by John Bailey. Popular Electronics, October 1982, page 72.

2. High Performance Op Amps, by Ray Marston. ETI, August 1982, page 50, figure 3.

Noise Reduction Unit



The component location for the noise reduction chip and display circuit.

PARTS LIST

Resistors (¼ W, 5%)

R1	1K ohm linear trim pot
R2	100 (see text)
R3	1k
R4	430
R5	910

Capacitors

C1	25 *u*, 16 V tantalum or electro.
C2,4,10, 11,13	1 *u*, 16 V tantalum or electro.
C3,12	0.0039 *u*
C5,8	0.1 *u*
C6	0.001 *u*
C7	not used
C9	0.047 *u*
C14	100 *u*, 16 V tan. or electro.
CL	0.015 *u* (see text)

Inductors

National Semiconductor designed the circuit for a coil which is unobtainable in Canada. We have substituted the Hammond 621 L. This is only necessary for the multiplex filter.

Semiconductors

LM1894 Dynamic Noise Reduction IC
LM3915 Level Indicator IC
D1-D10 Any size or colour LED

Miscellaneous

PCB, suitable case, power supply (see text), connectors if used.

HOW IT WORKS

There isn't a great deal to say that wasn't covered in the text; this very clever IC works it all out for you. The block diagram in Fig. 4 shows the interior of the chip. Both channels are summed, and the resultant signal converted to DC via various RC time constants chosen to minimize audible effects. This DC voltage then controls an electronic filter to open and close the bandwidth as necessary.

Input pins 2 and 13 like to see a signal level of about 300 mV for optimum performance. The maximum supply voltage is 18 V. Supply current will be about 17 mA for an 8 V supply. This does not include current drawn by the LED display, if used.

If the display is included, note that its power supply pin must be jumpered to the V+ of the DNR. A single-dot display results if points A and B (on the foil side of the PCB) are left unconnected; a bargraph displays results if they are linked together.

The multiplex filter may be omitted if the DNR is not used with FM broadcasts. C9 must still be included across pins 8 and 9.

BY COLLEEN McNEICE AND ROGER COTA

Build a DYNAMIC AUDIO NOISE FILTER

Cleans up radio, tape and record signals from any stereo system

THE "SILENCER" dynamic noise filter described here can eliminate tape hiss, record-surface noise, and atmospheric radio noise. Consequently, it is an ideal add-on device for stereo hi-fi systems. Moreover, it does not require encoding and decoding.

The device is essentially a voltage-controlled low-pass filter whose cutoff or break frequency is continually changing to accommodate program material and shut out any detracting noise. It only filters when noise and hiss are audible, when program material is at a low level or absent. The phenomenon of masking is utilized. That is, high-level signals mask noise that would be objectionable if program material level were low. When such masking occurs, the whole signal is passed. When there is no masking by program material, however, the filter extends the bandwidth only as far as required by the music. Beyond this, the high-frequency noise is attenuated. The frequency at which the filter begins rolling off to attenuate high-frequency noise is called the "break frequency."

About The Circuit. The silencer circuit constantly analyzes incoming signals for amplitude, frequency, and persistence. These factors determine the bandwidth at any instant, as well as how quickly the variable low-pass filter changes. Attack and release times vary with the music, thus eliminating a "pump and wheeze" effect of noise modulation.

The device has a continuously variable threshold control, with front-panel LEDs calibrated to indicate "Low," "Mid," and "High" break frequencies. The filter's break frequencies vary between 1.5 and 20 kHz with a roll-off slope of 9 'B/octave (maximum). The Silencer is a ngle-ended stereo device, making it ˀal for use with tapes, records, and ers for playback and record pur-es.

ˀe unit connects either in the auxili-

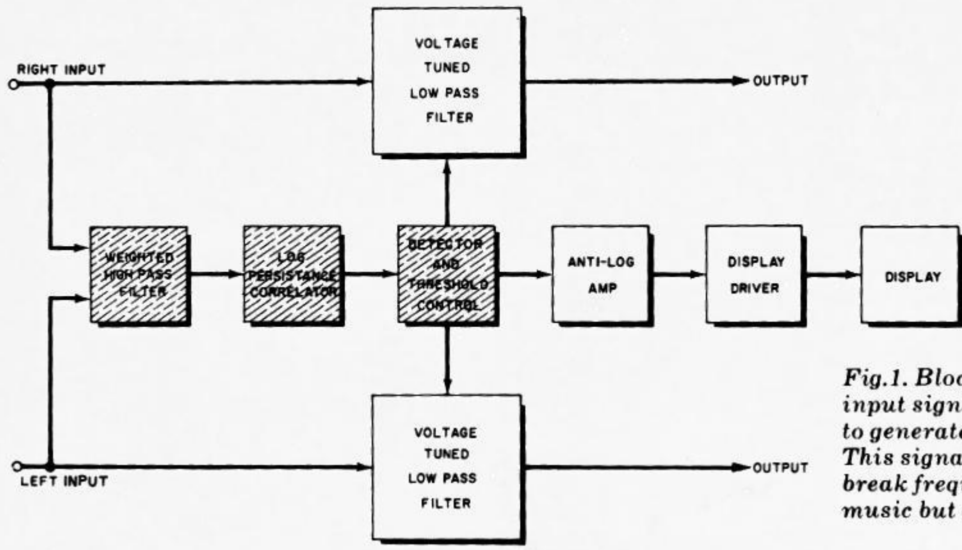


Fig. 1. Block diagram shows how input signals are processed to generate a control signal. This signal governs filter's break frequencies to pass music but attenuate noise.

ary mode or in the tape loop of your audio amplifier. On the back panel are IN and OUT jacks for the tape loop; the front panel also has a TAPE monitor button, and a system DEFEAT.

The block diagram of Fig. 1 shows the functions of the dynamic noise filter. The voltage-controlled low-pass filter is composed of *IC1A* and *IC1B*, as shown in the schematic of Fig. 2. (The components to the left of the dashed line make up one stereo channel; only one is shown in the schematic for clarity.) The gain of op amp *IC1A* is approximately *R3/R5*. At low frequencies, the capacitive reactance of capacitors *C4* and *C5* is very high, making the output of *IC1B* look like a low impedance source. The gain of *IC1A* is then:

$$A = R3/R5 = 10{,}000 \text{ ohms}/1000 \text{ ohms} = 10$$

At higher frequencies, however, the impedance of *C4* and *C5* decreases; *IC1B* generates an output and bootstraps *R5*. This bootstrapping effect causes *R5* to look larger. Therefore, gain A becomes smaller and the filter attenuates the high-frequency energy.

To vary the breakpoint of the filter, FET *Q1* has the ability to shunt the signal at the non-inverting input of *IC1B* to ground. Figure 3A shows the filter with the FET open and the high frequencies attenuated, while 3B illustrates the filter's action with the FET shorting the signal to ground. The control signal applied at the gate of the FET allows the bandwidth of the low-pass filter to be self-adjusting for any frequency. This allows high-frequency signals and subtle harmonics of fundamental bass frequencies to be passed, while unmasked noise is attenuated.

The circuits represented by the shaded blocks of Fig. 1 are the dynamic analytical controls. They automatically judge the program material, adjust the bandwidth to accommodate it, and change the attack and release times to maximize the masking effect and minimize noise modulation. The control signal is applied to the gate of *Q1*. It's determined by the (1) spectral content, (2) amplitude, and (3) persistence of the incoming signal.

The spectral content is sensed by the high pass weighting filter, a network made up by *R8, R29, R30, R31, C6, C17*, and *IC2A*. This network is driven by the output of *IC1B*, which actually determines the quiescent operating point of the low-pass filter. Amplitude is determined by threshold control, *R27*, a 100K-ohm front-panel potentiometer. This pot sets the voltage divider for the positive input to *IC2A*, and the dc level for *IC2A*'s output. The dc output level determines the quiescent operating point of the FET. The dynamic operation of the FET is adjusted by the ac control signal, allowing it to follow the program material. The ac component of *IC2A*'s output is determined by sensing the signal's amplitude on the output of *IC1B*.

The persistence log amp is formed by *R33, D2*, and *C20*. It checks the correlation coefficient of the signal, and adjusts the attack and release time of the low-pass filter to minimize any noise modulation problems. Variable attack and release times allow for the most effective masking of the noise.

The anti-log amplifier *IC2B* also senses the control voltage output of *IC2A*. This signal is then rectified and filtered by *D4* and *C21*, and is then used to drive threshold comparators *IC2C* and *IC2D*. These amps drive the logic network of *D5, D6*, and *D7*, which drives the display. The 10K-ohm trimpot, *R37*, is used to calibrate the LEDs. The red LED indicates a break frequency of 1.5 kHz, the yellow, a break frequency between 1.5 and 10 kHz, and the green that the filter is opening up above 10 kHz.

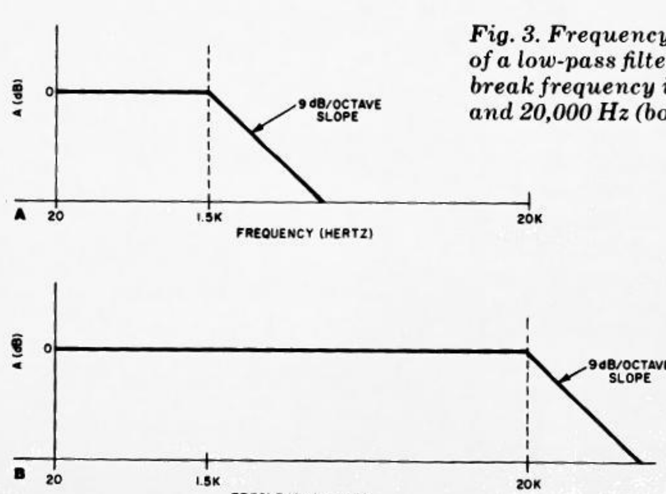


Fig. 3. Frequency response of a low-pass filter when its break frequency is 1500 Hz (top) and 20,000 Hz (bottom).

PARTS LIST

C1, C3, C9, C11, C21—1-µF 50-volt axial-lead electrolytic capacitors
C2, C10—680-pF disk ceramic capacitors
The following are 100-volt Mylar capacitors:
C4, C5, C12, C13—.022-µF
C6, C14, C17—.001-µF
C7, C15, C20—.1-µF
C8, C16—.01-µF
C18, C22, C23, C24—1000-µF 35-volt radial-lead electrolytic
D1—33-volt Zener diode
D2 through D7—IN914 signal diode
D8 through D11—IN4002 rectifier
F1—½-ampere fuse
IC1, IC2—µA4136 quad op amp (Fairchild)
J1-J8—RCA phono jacks
LED1—Red (Fairchild FLV 110 or equivalent)
LED2—Yellow (Fairchild FLV 410 or equivalent)
LED3—Green (Fairchild FLV 310 or equiv.)
Q1, Q2—Matched pair of 2N5458 JFETs.
The following are ¼-watt, 5% tolerance resistors:
R1, R13, R26—47,000 ohms
R2, R14, R35, R43, R44—4700 ohms
R3, R15, R33, R38—10,000 ohms
R4, R16—100 ohms
R5, R17—1000 ohms
R6, R18—39,000 ohms
R7, R19, R45, R46—2200 ohms
R8, R20, R29—15,000 ohms
R9, R21—11,000 ohms
R10, R22, R36, R39, R41—100,000 ohms
R11, R12, R23, R24, R34—1 megohm
R25—22,000 ohms
R28, R32—470,000 ohms
R30—680,000 ohms
R31—130,000 ohms
R40, R42—27,000 ohms
Other resistors and controls:
R27—100,000-ohm potentiometer with switch (CTS FR-GC-XM 450 or similar)
R37—10,000 ohm thumbwheel trimpot
R47, R48—10 ohms, ½ watt, 5% tolerance resistor
S1, S2—DPDT switches
S3—110-V, 2-A switch (part of R27)
T1—22-volt center-tapped, 50-mA transformer
Misc—Ac line cord, knob for threshold pot, buttons for switches, suitable enclosure, hardware, hookup wire, solder, etc.
Note—The following items are available from Logical Systems, 3314 "H" St., Vancouver, WA 98663 (Tel. 206-694-7905): Complete 318 Silencer kit, including 6063 extruded aluminum chassis and hand-finished black walnut end pieces, $129.00. Also available separately: Etched and drilled circuit board, $15.00; individually tested and matched 2N5458 FETs, $3.50. Washington state residents please add 5% sales tax.

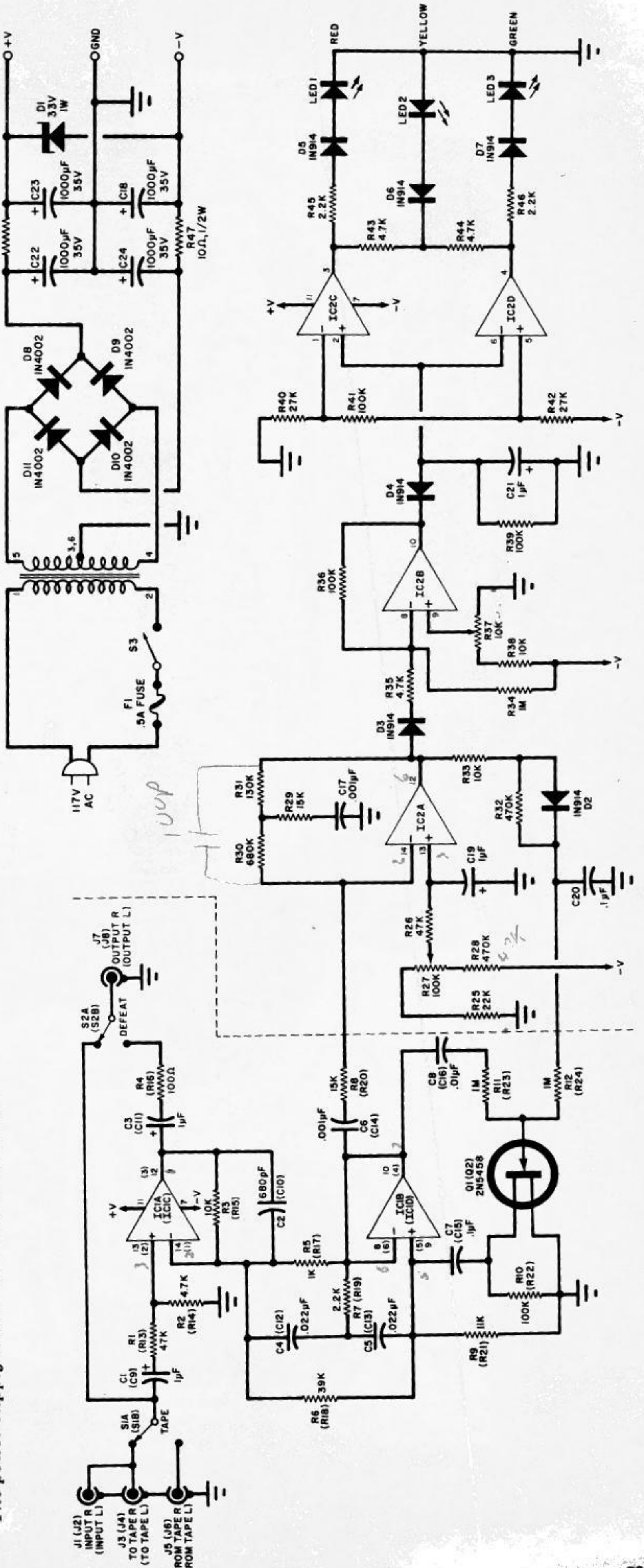
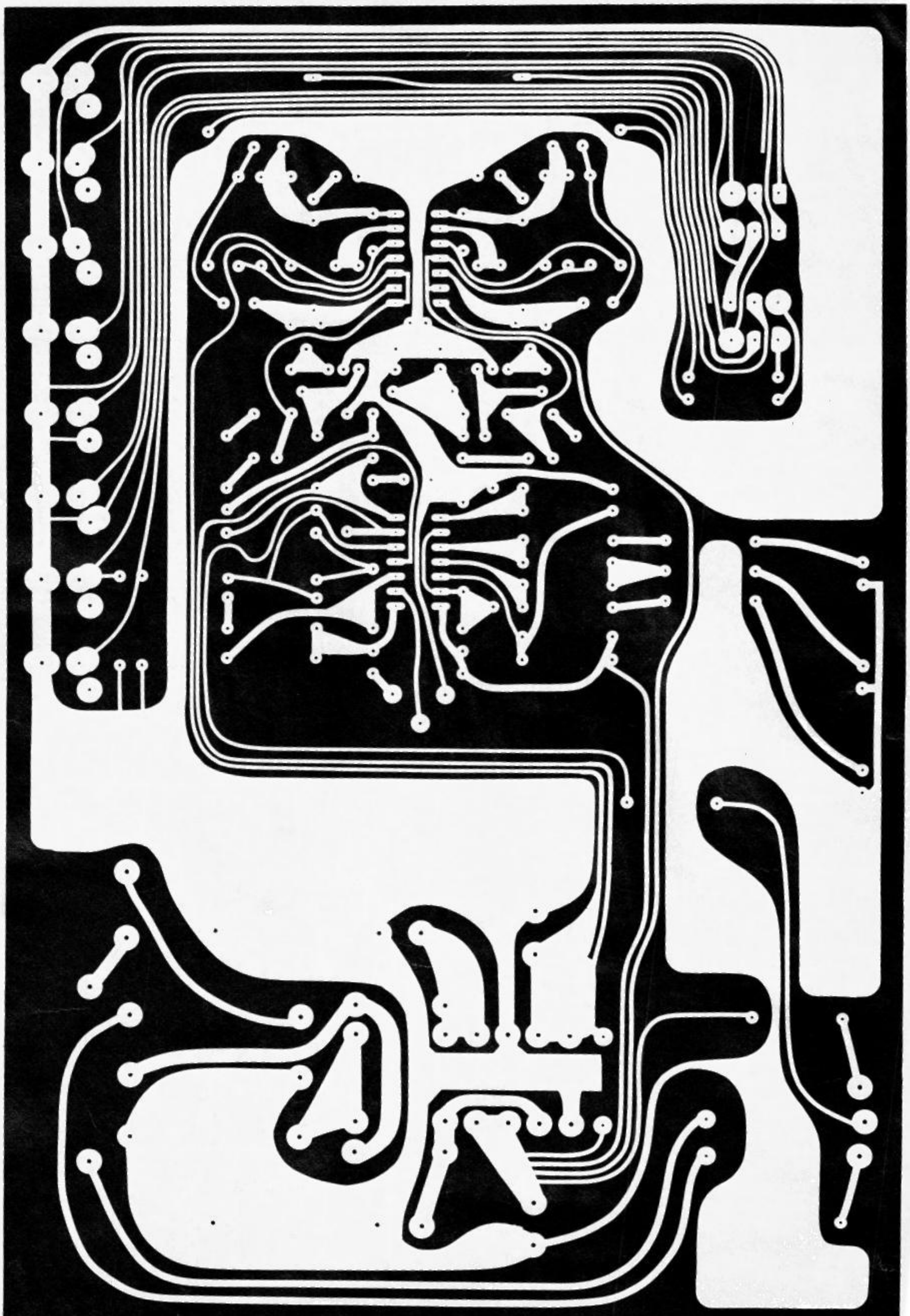
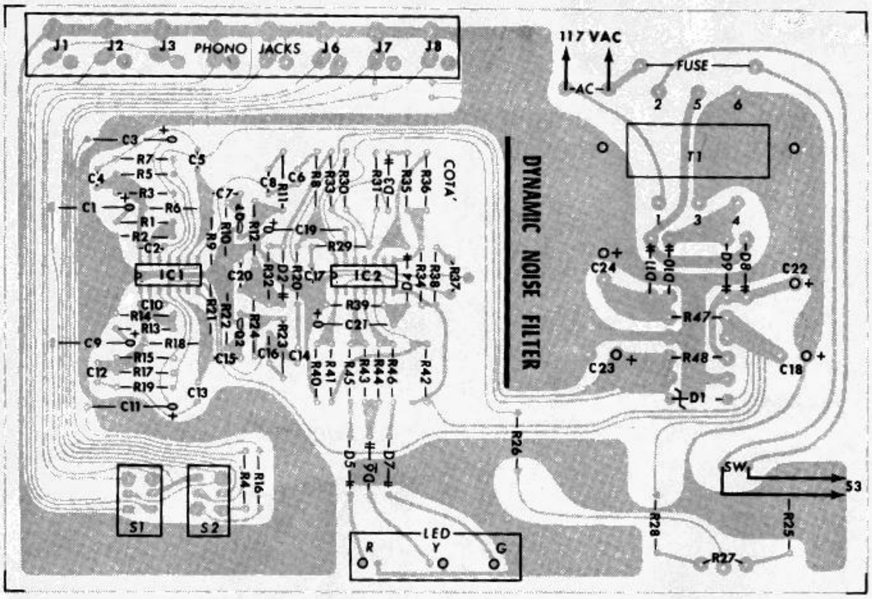


Fig. 2. Schematic diagram. Components of the left-channel filter are numbered in parentheses. The power supply is common to both channels.



A



B

Fig. 4. Actual-size etching and drilling guide for the "Silencer." Board is shown at (A); parts placement at (B).

OPERATING SPECIFICATIONS—"SILENCER"

Hiss Reduction:	15 dB at 10,000 Hz
Max. Filter Slope:	9 dB/octave
Frequency Response:	20 to 20,000 Hz ±0.5 dB
Minimum Bandwidth: (Filter Closed)	1500 Hz
Dynamic Range:	Output noise greater than 100 dB below max. output, 20 to 20,000 Hz
S/N Ratio:	Better than 85 dB below 2 V ac output 20 to 20,000 Hz
THD:	Less than 0.1%, at rated output, 20 to 20,000 Hz.
IM Distortion:	Less than 0.01% at rated output 60/7000 Hz mixed 4:1; typically less than 0.005%
Rated Output:	2 V ac into 10,000 ohms
Max. Output:	10 V ac into 10,000 ohms
Input Impedance:	47,000 ohms, single ended
Output Impedance:	100 ohms
Power requirements:	110/120 V, ac 50/60 Hz, 8 W

Note: All measurements made with filter bandwidth open maximum except where specified. (This is the worst-case condition.)

Construction. This unit is most easily constructed using a printed-circuit board. Complete etching and drilling guides are shown in Fig.4A, with the component guide shown in Fig. 4B. Proper orientation of parts is very important. Take careful note of how FETs *Q1* and *Q2* are mounted as well as op amps, diodes, and electrolytic capacitors. Also observe that the dynamic characteristics of the FETs must be matched. Moreover, when choosing op amps, it is important to make sure that the one chosen for the detection circuit, *IC2*, has an open-loop gain of at least 50 dB at 10 kHz. Op amps in the parts section were chosen for their excellent noise figures.

The unit is designed to fit into a custom aluminum extrusion, held by the eight screws in the wood ends. Any suitable enclosure will work, however. The circuit board itself measures 6" × 9". The RCA phone jacks, front-panel switches, and threshold pot are circuit-board mounted for ease of construction and minimum noise. LEDs may be circuit-board mounted or attached to your front panel and then wired. If you choose not to use the furnished printed-circuit guides, make sure that the power supply is as far away as possible from the rest of the circuit to eliminate stray hum.

Calibration. Calibration should be done before you fully enclose the unit. To calibrate, connect the noise filter into your amplifier's or receiver's auxiliary or tape input. Find a low-level noise source—an erased magnetic tape would be ideal. If you don't have tape facilities you may use the inside groove of an LP record. Increase the amplifier's gain so you can hear the noise very well.

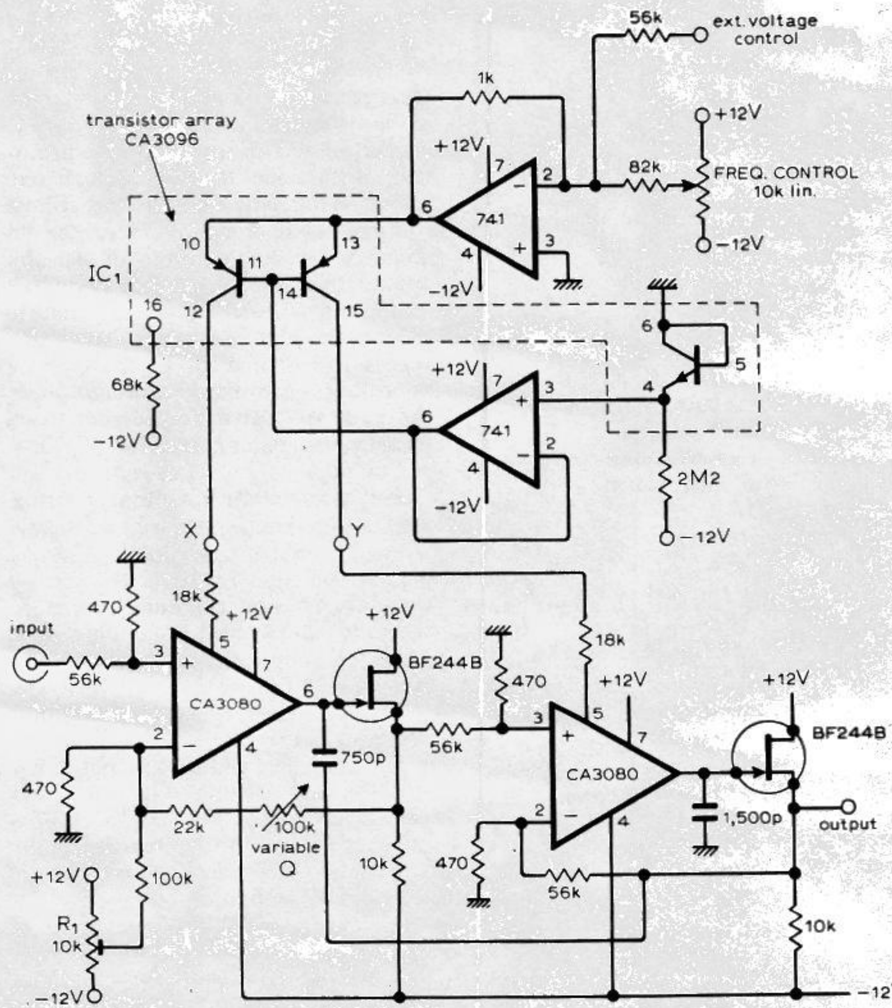
Start with the Silencer's threshold pot turned fully counter-clockwise and slowly turn the control knob clockwise. You will hear the noise change character and become more objectionable. Return to the position where the noise-content change just begins (listen several times so you will be able to identify this point). With the threshold knob in this position, adjust *R37*, the thumbwheel trimpot, so that the red LED lights. You should adjust the pot so that it is at the point where only a slight adjustment will cause the yellow LED to light.

In conclusion, this easy-to-build noise-reduction system will be a helpful and versatile addition to any stereo hi-fi system, cleaning up signals from any source. ◇

Voltage/current controlled filter

The circuit shows a controllable filter, having a −12dB/octave roll off. Frequency range is 15Hz to 15kHz, this frequency being controlled by either a voltage or current. Voltage to current conversion is achieved with a logarithmic characteristic (IC_1); thus the filter frequency moves in octaves/volt rather than in Hz/volt. The CA3080 operational transconductance amplifiers are used to produce variable resistors. If just manual control of the cut off frequency is required, only the bottom half of the circuit is used by shorting points X−Y and connecting them to the wiper of a 10kΩ log potentiometer between −12V and ground. If voltage control of the cut off frequency is required, the top half of the circuit is used. By using a transistor array, good matching and temperature stability is obtained. The separate transistor (pins 4, 5, 6) provides an offset bias voltage of the correct value and also a voltage to compensate for any temperature changes. The CA3080's may be selected for minimum d.c. offset change with respect to frequency control, or the offset may be nulled by R_1.

T. Orr,
Putney,
London SW15

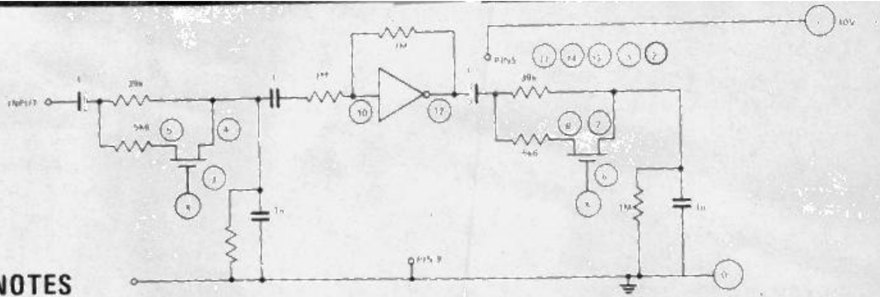


CHEAPO VCF!

Readers intending to build the dynamic noise limiter may be interested in the following circuit.

The circuit consists of two RC low pass filters connected by a unity gain buffer (inverting).

The n-channel MOSFETs are used as voltage controlled resistors to vary the cut-off frequency of the two filters which are controlled by a voltage entered at points X – The additional resistors limit the variation to limits of 5 and 50kHz.



NOTES

1. The control voltage should be positive-going, not negative-going as in the original.
2. Signal input should be resticted to 50mV, when distortion will be low.
3. The cut-off is less sharp than the DNF VCF so less trouble can be expected from changes in bandwidth, as such changes will be less obvious.
4. A high impedance buffer is required at the output.

development is now in excess of 9 million per year and rising. Urgent consideration is therefore being given to providing l.s.i. synthesizers for the equipments to reduce this phenomenal demand for individual channel crystals . . . If a parallel Citizens' Band development had been taking place in the UK there would now be probably some 500,000 vehicles equipped with CB radio, many more than currently fitted under the present conventional licensing basis. This US development is interesting from two points of view. First, it seems to argue that there is a very large pent up demand for mobile radio. Secondly, it indicates that a tremendous utilisation can be got out of 12 channels. All this on a.m. too! Should we, I wonder, introduce Citizens' Band radio in the UK? On the face of it I cannot see why not. I feel it might well be a very healthy development."

STEREO NOISE LIMITER IMPROVEMENT

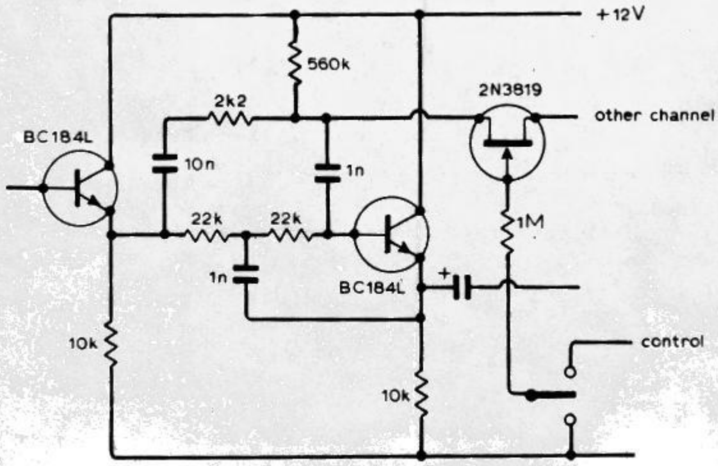
The two circuit ideas shown on p.474 of the October issue can be developed and combined in an interesting way. The dynamic noise limiter offered by Mr Richter is not really satisfactory as it stands since the switch-over from stereo to mono, even at low volume level, can be disconcerting to say the least.

However, the hiss which he is attempting to remove by this means is precisely an antiphase effect; thus a low pass filter, designed along similar lines to Mr Oldfield's stereo rumble filter, will remove it – with very little detriment to the overall signal. The f.e.t., driven by the amplifier output (or whatever – I prefer to drive it direct from the tuner), is now used to switch the filter into operation rather than to switch over to mono.

The component values shown in the circuit give very good results, and it was found that with a 2N3819 taken at random switch-over occurred abruptly at V_{gs} about –2V. Operating the f.e.t. from the positive line, as shown, facilitates switching the device in and out.

The bypass capacitor for non-filtering operation, 10nF, requires a resistor (2k2) in series with it to prevent excessive lowering of the input impedance at high frequencies. Otherwise the circuit is the low-pass corollary of Mr Oldfield's. I feel that the simple ingenuity of his circuit deserves considerable commendation.

Giles Hibbert,
Blackfriars,
Oxford.



Mr Hibbert's improved noise limiter circuit

PHASE AUDIBILITY: RATE OF CHANGE

I have in my laboratory a Fourier synthesiser, consisting of a fundamental oscillator at a frequency of 256Hz and harmonic oscillators up to the twelfth, all phase-locked to the fundamental. It is possible to alter independently the amplitude and phase (through a full 360°) of each of the harmonic oscillators. Thus it is possible to synthesise a wide variety of waveforms*, for listening tests.

The consensus of opinion among my colleagues is that the tonal quality of a sound depends solely on the amplitudes of the harmonic components and not on the phases. The phase of any harmonic may be altered individually by any amount without, apparently, altering the tonal quality. This applies when the phase-control knobs are stationary; while any one of them is being moved, i.e. while there is a rate of change of phase, the ear readily detects that "something is happening" (it is difficult to describe the effect in words). But since rate of change of phase is synonymous with frequency it is arguable that the detectability of the effect is due to the fact that a harmonic component becomes slightly inharmonic while the phase knob is being rotated.

The effect is very similar to that caused by any movement in the laboratory where the tests were carried out. This is acoustically quite lively, so that there must have been a marked standing-wave pattern, which would alter as a result of movement by anybody in the room. What seems to me surprising is (a) the sensitivity of the ear to movement, and (b) the fact that when movement ceases the sound appears to revert to its original tone quality, in spite of the fact that in this case one would expect the amplitudes as well as the phases of the harmonic components to have altered at the position of the ear.

These were all rather rough experiments, and I hope in time to do something more precise. But the fact is that we know so little about how the ear and brain perceive sounds that we do not even know what are the crucial experiments we should perform.

P. L. Taylor,
University of Salford.

* See Mr Taylor's article "Frequency modulation illustrated" in this issue. – Ed.

TV SOUND: BOOSTING WEAK SIGNALS

I have noted with interest the recent articles on television tuner design (Oct. 1975—Jan. 1976 issues), especially with reference to television sound reception. In this area of Ireland the cross-channel u.h.f. TV stations provide alternative programmes – if they can be received. U.h.f. signals in the 600MHz region from a 100kW e.r.p. transmitter 120 miles away are usually very weak and suffer from severe tropospheric fading, even at elevated sites. From experience I have found that only at 1000ft a.s.l. are signals acceptable.

Those of us at lower altitudes receiving u.h.f. signals have to cope with signals on television receivers that are loaded with noise to say the least. I have been experimenting for some time in order to get less noisy reception and offer the following comments.

At extreme distances fading occurs on signals at different rates at different frequencies and this includes u.h.f. television signals. Even when receiving steady but noisy video signals the audio signals are usually quite noisy also, not due to deficiencies in the f.m. system but to the fact that they are attenuated in the inter-carrier sound detection process. Having a few various u.h.f. tuners, I tried feeding them directly into the input of a sensitive f.m. portable (Tandberg TP41). The reception of weak signals using this method was much superior to that of the normal television receiver, especially when the outboard tuner was re-aligned to the v.h.f.-f.m. band frequencies acting as i.f. and detection stages. Mechanical and varicap tuners gave similar results. In fact when tested on a signal generator, signals of 3 to 5 microvolts of sound carrier in Band 4 gave good acceptable signals, and I am sure these figures could be bettered.

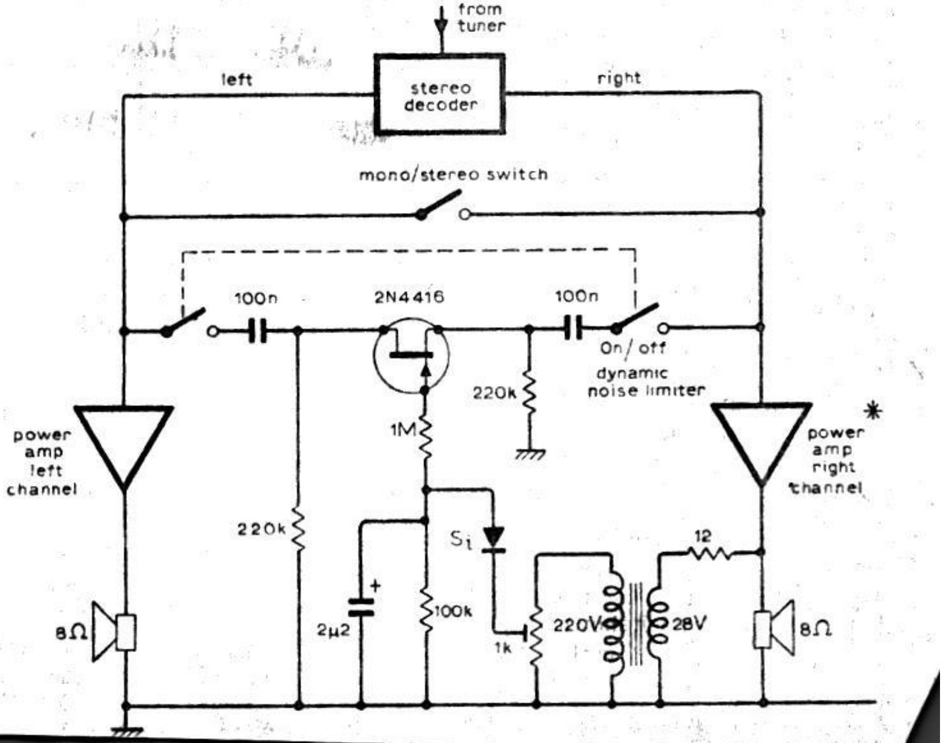
Desmond J. Walsh,
Carrick-on-Suir,
Co. Tipperary, Republic of Ireland.

Stereo dynamic noise limiter

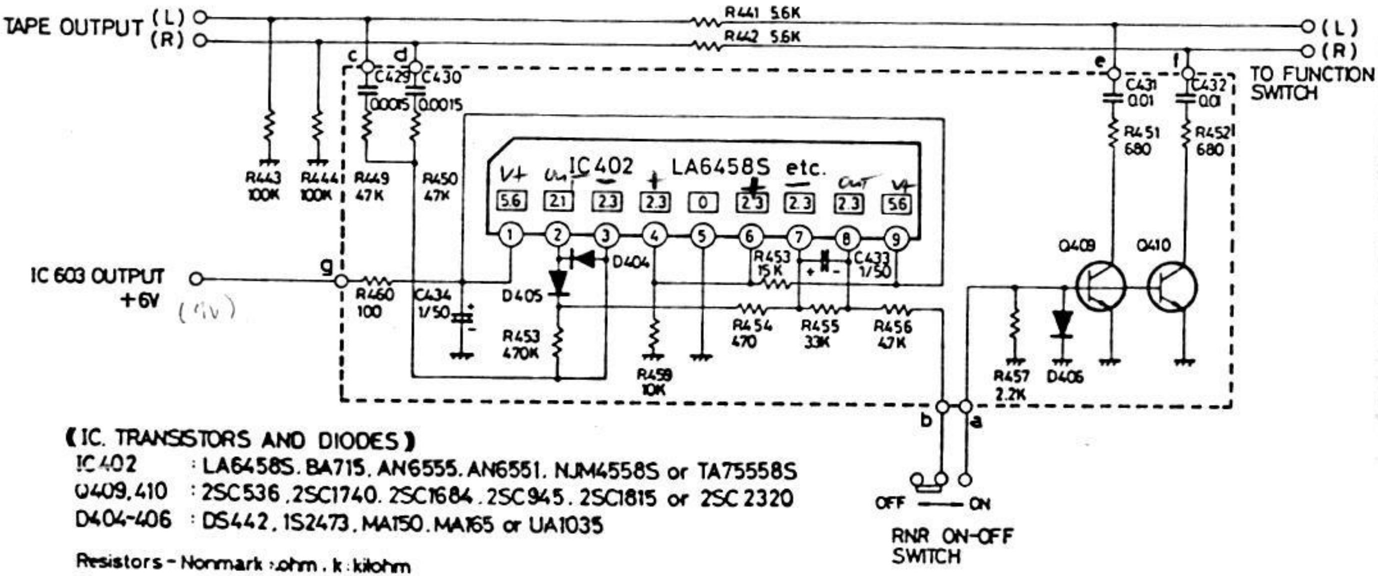
Stereo reception of f.m. stations is often accompanied by a slight noise signal. This signal is only heard at weak passages; in this case a noise limiter will help to produce a pseudo-stereo sound, which has reduced noise.

In practice both audio channels are short-circuited, depending on the audio signal strength. The short circuit is realized by a field effect transistor 2N4417, whose gate is controlled by the output voltage. If this voltage is not sufficient to drive the f.e.t, an amplifier or transformer must be used.

J. W. Richter,
Eindhoven.



SCHEMATIC DIAGRAM (RNR SECTION)



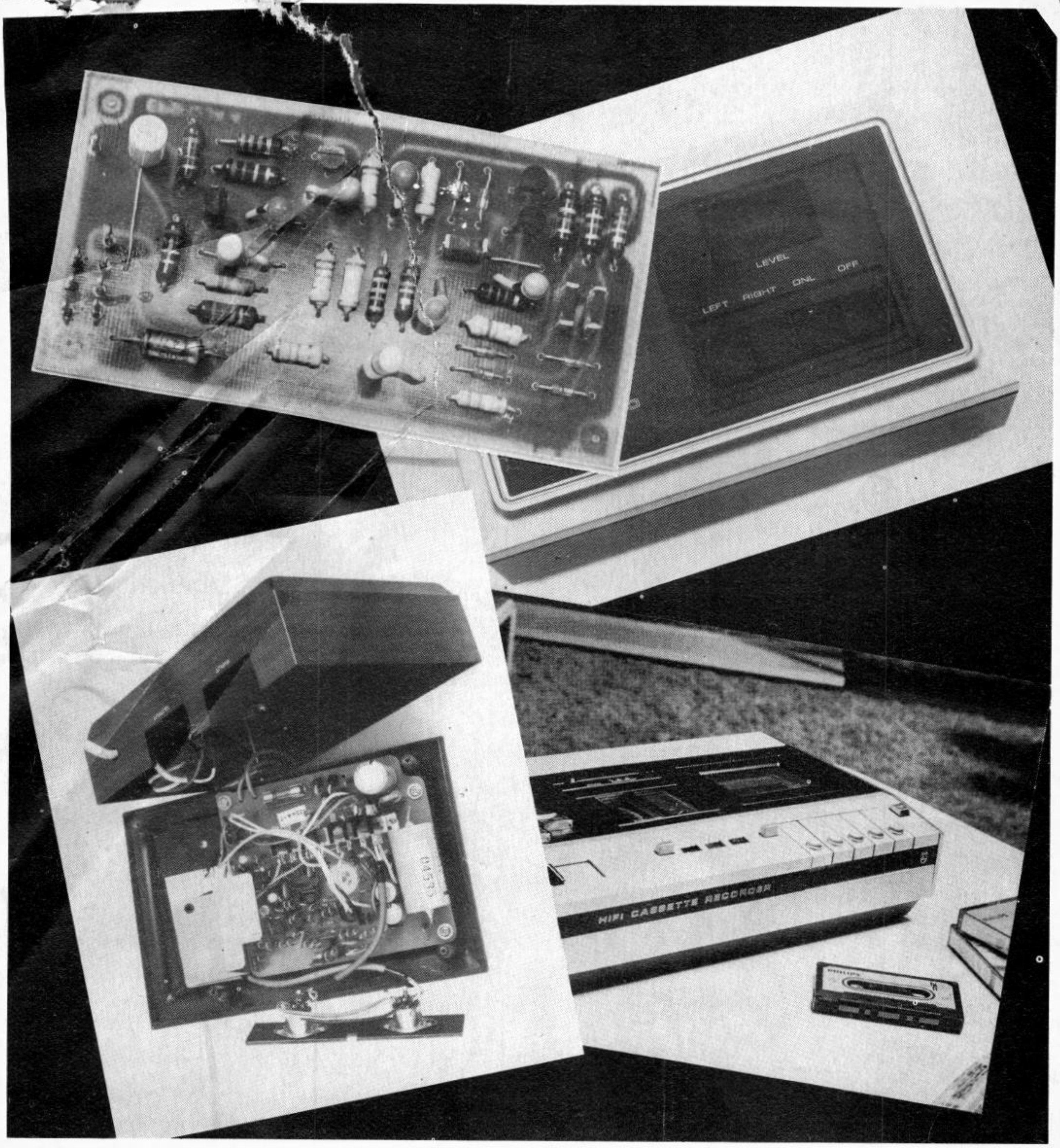
(IC. TRANSISTORS AND DIODES)

IC402	: LA6458S. BA715. AN6555. AN6551. NJM4558S or TA75558S
Q409, 410	: 2SC536, 2SC1740. 2SC1684. 2SC945. 2SC1815 or 2SC 2320
D404-406	: DS442, 1S2473. MA150. MA165 or UA1035

Resistors - Nonmark : ohm . k : kilohm

Capacitors-Nonmark: μF. P : μμF

Voltage specified are positive with respect to chassis.



dynamic noise limiter

One of the major problems with tape recorders in general, and cassette recorders in particular, is tape noise. For this reason, noise reduction systems are in demand — and several systems have been launched in past years.
The Philips DNL system described here has several advantages: it is relatively cheap, it can be added to existing equipment without difficulty and it only affects reproduction so that it can be used on existing (conventional) recordings.

The DNL system reduces or eliminates high frequencies (noise) during the quieter passages and pauses of a recording. During loud passages (near maximum modulation) the system is not operative – tape noise is masked by the audio signal in this case. The noise reduction during quiet passages results in an apparent increase in dynamic range.

The block diagram of the circuit is shown in figure 1. The input amplifier stage (A) is used for impedance matching to the tape recorder. From here the signal is fed into two parallel channels. The upper channel consists of a high-pass filter (B), amplifier (D) and variable and fixed attenuators (E) and (G). The lower channel consists of a phase-shifting network (all-pass filter, C) and a preset (or fixed) attenuator (F). The final output is the sum of the outputs of both channels.

The operating principle can be described briefly as follows. The output V_1 of the all-pass network (C) is equal to the original signal, except for having an additional phase-shift. There will be no audible difference between this signal and the original. The output V_2 of the high-pass filter contains the high-frequency portion of the original signal. For all frequencies the signals V_1 and V_2 are in anti-phase so that, if these two signals are summed, the high-frequency portion of the original signal is cancelled out. The net result is a low-pass filter.

For large input signal levels the variable attenuator (E) becomes operative, thereby reducing the contribution of V_2 to the output signal. The high frequency content of V_1 is no longer cancelled out: it is passed without attenuation to the output.

For readers who would like to see a more mathematically exact description (and we suggest that readers who don't like mathematics skip this paragraph): the transfer functions of the high-pass

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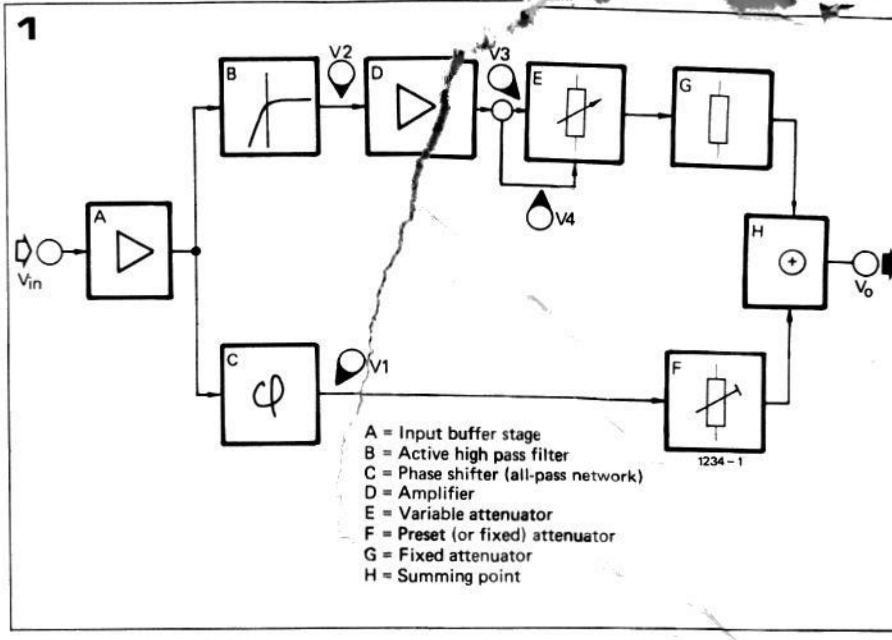

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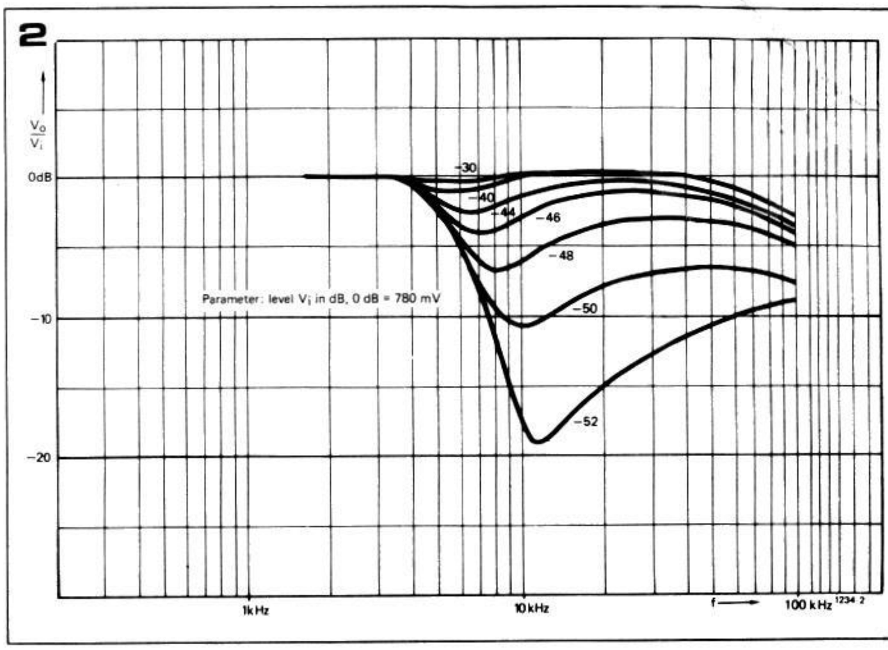

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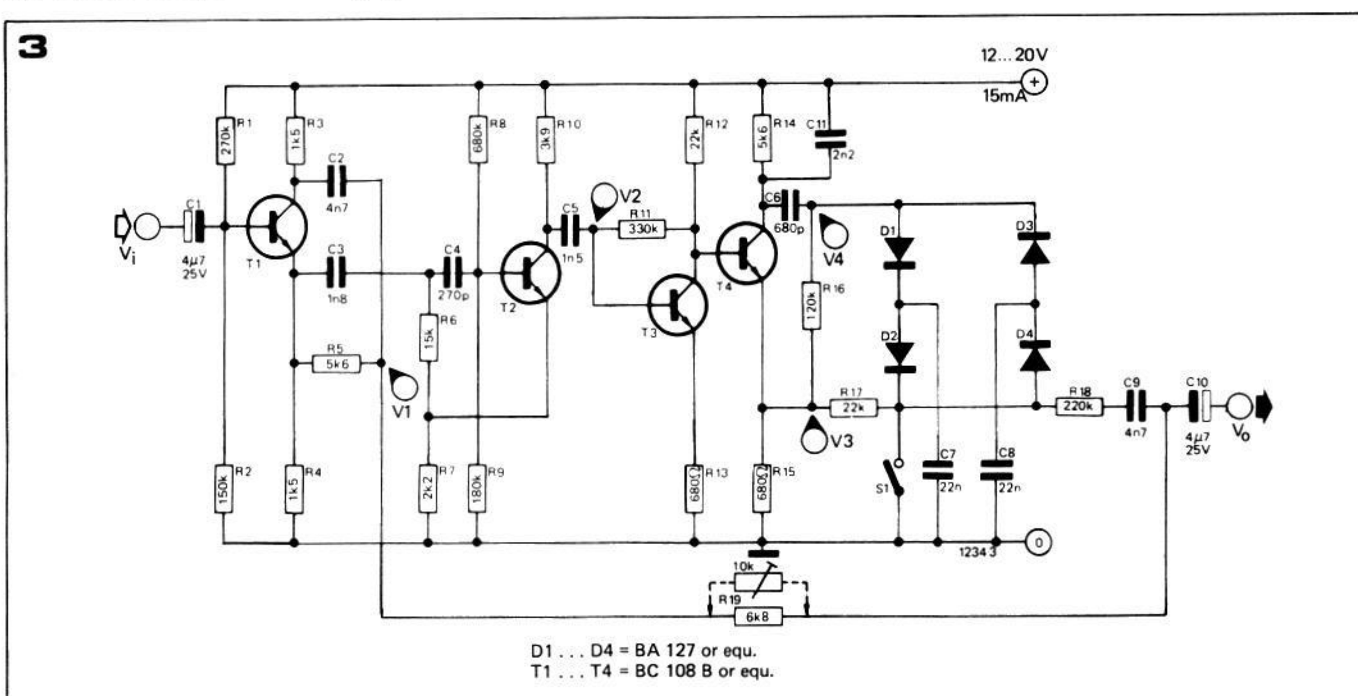

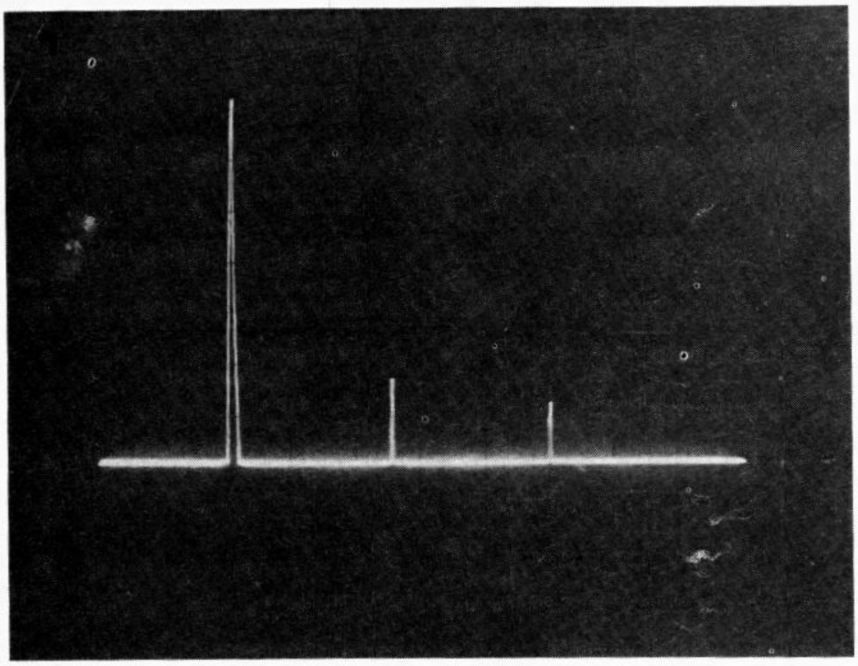

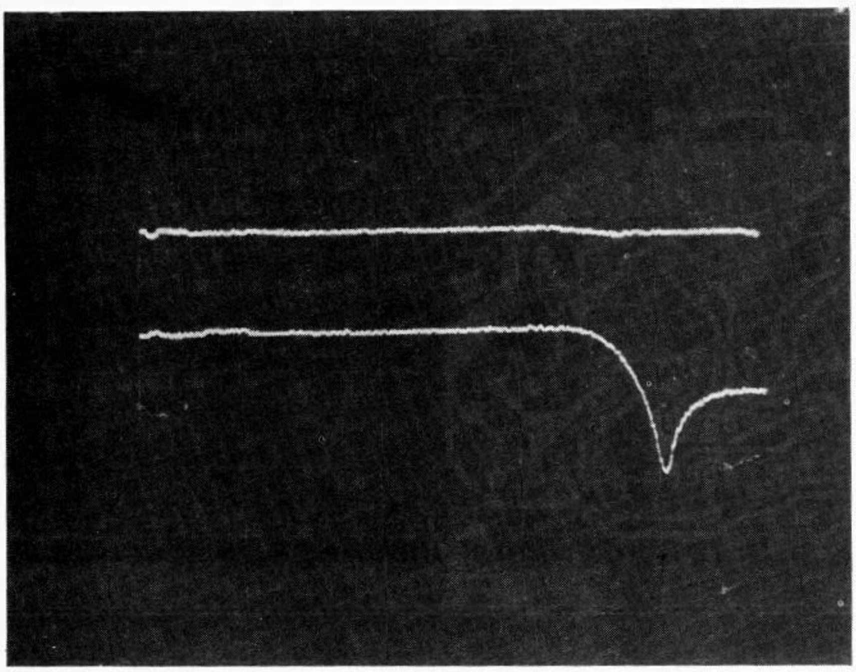

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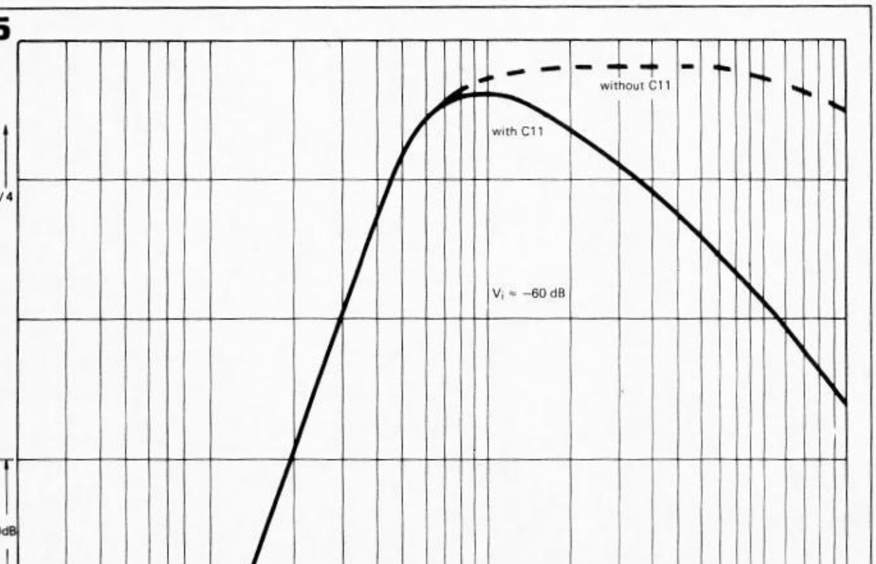


low that, even after amplification in T1 it cannot produce a sufficiently large control voltage to forward-bias D1 or D3. It will be obvious that, in this case, the input signal (V_3) will certainly be insufficient to forward-bias D2 or D4. V_3 is approximately one-eigth of V_4, and V_4 is itself insufficient to forward-bias the diodes! The cancellation signal is thus passed without any further attenuation: the DNL is at its most active. Next consider the situation where the input signal is at a much higher level – say 500 mV. The peak value of the control voltage will be some 5.8 V in this case, so that C7 and C8 would be charged to more than 5 V above or below their original potential – if that were possible. However, D2 and D4 are connected in series between the two capacitors, so that the voltage difference between them can never be larger than about 1.4 V, i.e. two diode-drops. D2 and D4 are now forward-biassed, so that they form an effective short-circuit for audio signals. In this situation, R17 and C7/C8 constitute a low-pass filter with a corner frequency of approximately 300 Hz – giving an attenuation of 20 or more above 5.5 kHz. The DNL as a whole is inoperative: there is practically no cancellation of high frequencies.

It is possible to estimate the effect of the attenuator between the two extreme cases described above by referring to figure 2. The results of some brief calculations, based on these characteristics and the estimated gain or attenuation of the various stages, are summarised in Table 1.

Reading first from left to right: the input voltage to the DNL is given in dB, 0 dB being defined as 780 mV. The input level in mV can thus be calculated, and from this the input voltage to the attenuator (V_3) can be estimated. This, in turn, gives the peak value of the control voltage (V_4).

Now, reading from right to left: V_o/V_{in} is given in dB for the various levels of V_{in}; V_o can be calculated from this. The difference between V_o and V_{in} is equal to the high frequency cancellation component coming from the variable attenuator; an estimate of the fixed attenuation due to R18 gives the signal level at the D2-D4 junction: αV_3. Since V_3 is known, the attenuation factor α can be calculated.

It will now be obvious that the first extreme case outlined above – very low signal level – corresponds to the situation for signal levels of 2 mV or less. The control voltage is 460 mV or less, so that the diodes are blocked and the attenuator is almost inoperative ($0.9 \leqslant \alpha < 1$). The second extreme case – high signal level – corresponds to signal levels of 25 mV (or more): the diodes are almost in saturation, and the attenuation factor has practically reached the theoretical limit of 0.05 determined by the low-pass network R17, C7/C8. Finally, to pick one intermediate value: at an input level of 3.1 mV the peak value of the control voltage will be approximately 720 mV.

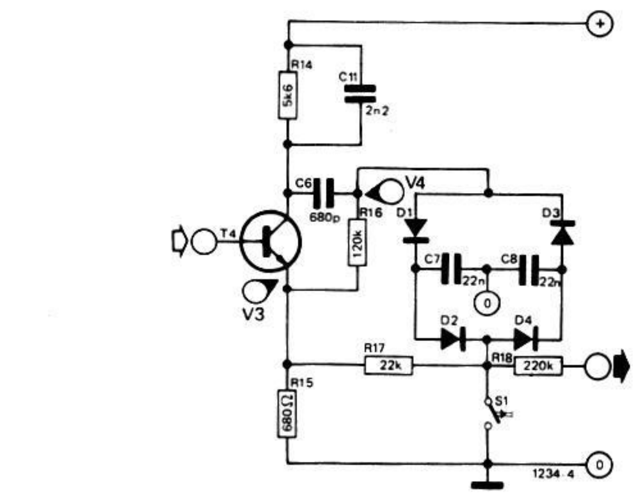
Figure 1. Block diagram of the DNL unit.

Figure 2. Performance of a basic DNL unit. As can be seen from this graph, it works as a low-pass filter as long as the high-frequency component of the signal is at a low level. As soon as the high-frequency component reaches a significant level the DNL becomes inoperative (flat frequency response).

Figure 3. Complete circuit of the DNL (one channel; for stereo two of these units are required).

Figure 4. This part of the circuit is the variable attenuator.

4



and all-pass networks are as follows:

high-pass:

$$H_h(p) = \frac{(pT)^3}{(1 + pT)(p^2T^2 + pT + 1)},$$

i.e. a third order Butterworth response;

$$\text{all-pass: } H_a(p) = \frac{1 - pT}{1 + pT}.$$

When its input signal level is high, the variable attenuator will suppress the output from the high-pass filter; the output of the DNL is equal to the output of the all-pass filter in this case: a flat amplitude response. For low input levels, however, the variable attenuator adjusts to a setting where the total attenuation in the amplifier (D) and attenuators (E and G) is equal to the fixed attenuation in F. In this case the total transfer function of the DNL becomes the sum of the two transfer functions:

$$H_t(p) = H_h(p) + H_a(p) = \frac{1}{(1 + pT)(p^2T^2 + pT + 1)},$$

i.e. a third order Butterworth low-pass filter.

Summing it all up briefly: in the absence of high frequency signals of any importance, the DNL circuit will operate as a sharp (18 dB/octave) low-pass filter and thus reduce tape noise; however, if the original signal has a significant high-frequency content the filter action will become progressively less, until at a certain level it will be totally inoperative. The graphs shown in figure 2 illustrate this.

For the actual design, values must be chosen for three independent variables:

- the *corner frequency* of the filter: if this is too high there will be little or no audible noise reduction; if it is too low, noise modulation effects will be audible on program material with mainly low-frequency content (e.g. pianosolo). The value chosen in this design is 5.5 kHz.
- the *critical signal level* at which the system starts to become inoperative. The choice depends on the nominal signal level at the input to the DNL and on the S/N ratio of the program source. The value chosen is approximately 2 mV, corresponding to −52 dB with respect to a 780 mV nominal level.
- the *attack time constant* of the variable attenuator. A slow attack will lead to some deterioration in transient response, but a fast attack can give rise to distortion at high frequencies – particularly in the critical region where the system is beginning to become inoperative. The value chosen is approximately 0.1 ms.

The circuit

The complete circuit is shown in figure 3. Referring to the block diagram (figure 1), the circuit can be analysed as follows:

T1 is the input stage (A in figure 1). The input impedance is approximately 75 k. Simultaneously, T1 with the network C2/R5 works as all-pass filter (C in figure 1). The time constant is approximately C2 · R5 ≈ 27 µs.

R19 is the preset (or fixed) attenuator (F in figure 1); a 6k8 fixed resistor can be used for most applications. If a preset potentiometer (10 k) is used instead, this can be adjusted by playing an unmodulated tape and setting R19 for minimum hiss.

The active high-pass filter (B in figure 1) consists of T2, C3, R6, C4, $R8 /\!/ R9 /\!/ R_{in,T2}$ and the feedback loop over R7. This is a second order filter; the third element of the total filter is C5 and $R10 + R_{in,T3}$. The time constants are chosen to obtain the required corner frequency (5.5 kHz). The actual component values used differ slightly from the theoretical values, to compensate for a certain amount of mutual loading of the circuits.

Part of the signal amplification (D in figure 1) is already achieved in the active high-pass filter (T2); T3 is the second half of the amplifier.

The components from T4 up to R18 are the variable attenuator (E in figure 1). This will be discussed in greater detail further on.

The fixed attenuator (G in figure 1) is simply R17 + R18.

The summing point (H in figure 1) is the junction C9 – R19 – C10.

The variable attenuator

This part of the circuit is shown separately in figure 4.

The input signal is amplified by T4 and passed through a further high-pass filter (C6, R16) to obtain the desired control voltage (V_4) for the variable attenuator. C11, in conjunction with R14, gives a high frequency roll-off at a slightly higher corner frequency. The result of this is that the attenuator becomes progressively less effective at higher input frequencies as the control voltage decreases (figure 5). Conversely, if the attenuator is less effective, the DNL as a whole becomes more effective towards higher frequencies: a larger high-frequency component will be subtracted from the main signal at the summing point. This effect can be seen by comparison of the characteristics shown in figure 6 with those shown in figure 2: the latter characteristics are those of the circuit without C11, whereas figure 6 shows the performance of the total circuit, including C11.

The attenuator itself consists of D1 . . . D4, R17, C7 and C8. The high-frequency input signal (V_3) is fed to R17, and an amplified and filtered version of this signal is fed (in anti-phase) to the junction D1 – D3: this is the control voltage (V_4).

In the absence of any input signal, the tank capacitors C7 and C8 will be charged through R16-D1 and R17-D4 respectively to approximately the emitter potential of T4. C6 will also charge until the junction C6-R16 reaches this potential.

Let us now first consider the situation where the signal level at the input is so

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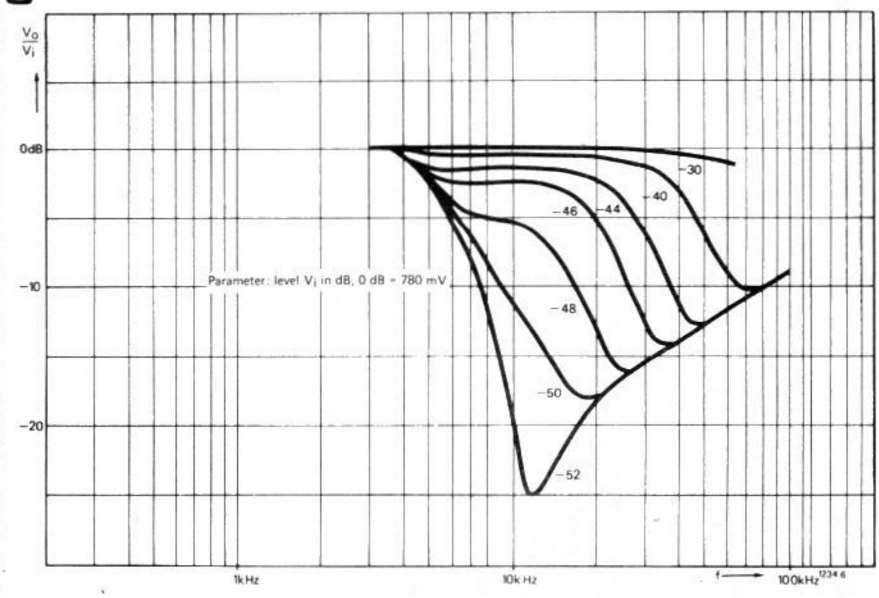

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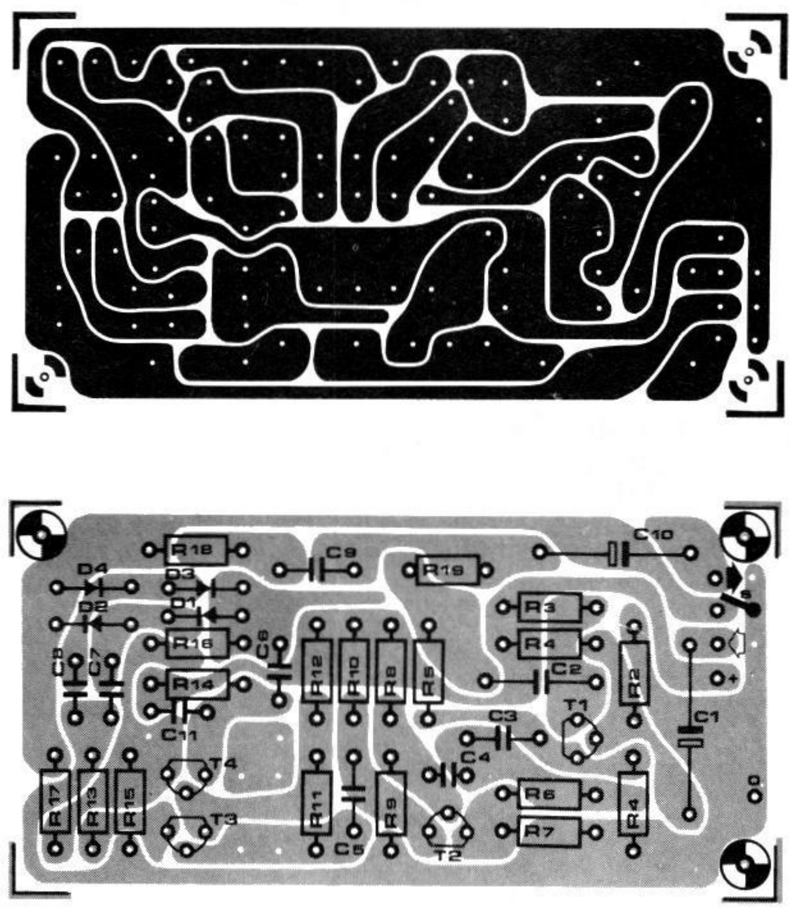

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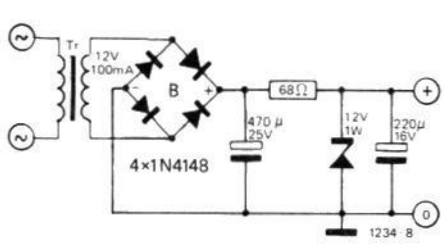

Figure 5. Control voltage (V_4) as a function of frequency, at a fixed input level, showing the influence of the capacitor C11.

Figure 6. Performance of the DNL unit described here, i.e. with capacitor C11 included.

Figure 7. The p.c. board and component layout (EPS 1234).

Figure 8. A simple power supply, which can be used if the supply cannot be derived from the main equipment.

Photo 1. Harmonic distortion of the DNL. Test signal 5 kHz, 500 mV; harmonics less than −54 dB (0,2%).

Photo 2. Frequency response for an input level of 25 mV (upper trace) and for 2.5 mV (lower trace). The dip is at approximately 11 kHz, and 25 dB down.

Parts list for figures 3 and 7

Resistors:
R1 = 270 k
R2 = 150 k
R3,R4 = 1k5
R5,R14 = 5k6
R6 = 15 k
R7 = 2k2
R8 = 680 k
R9 = 180 k
R10 = 3k9
R11 = 330 k
R12,R17 = 22 k
R13,R15 = 680 Ω
R16 = 120 k
R18 = 220 k
R19 = 6k8 fixed or 10 k preset (see text)
P1 = 100 k or 220 k preset (see text)

Capacitors:
C1,C10 = 4μ7/25 V
C2,C9 = 4n7
C3 = 1n8
C4 = 270 p
C5 = 1n5
C6 = 680 p
C7,C8 = 22 n
C11 = 2n2

Semiconductors:
T1 . . . T4 = BC108B or equ.
D1 . . . D4 = BA127 or equ.

Miscellaneous:
S1 = SPST (mono version), or
DPST (stereo version).

Table 1. An estimate of the variable attenuator action, as derived from the DNL performance shown in figure 2.

Figure 9. Two possible arrangements for the input attenuator. The arrangement of figure 9a reduces the input impedance, whereas the other arrangement increases it.

Figure 10. The DNL can be connected between the cassette recorder and a main amplifier (10a) or it can be included in the cassette recorder itself (10b).

Table 1

V_{in} (dB)	V_{in} (mV)	V_3 (mV)	$V_{4, peak}$ (mV)	α	$\alpha \cdot V_3$ (mV)	$V_{in} - V_o$ (mV)	V_o (mV)	V_o/V_{in} (dB)
−30	25	500	5800	0.06	28	1.4	24	− 0.5
−40	7.8	156	1800	0.11	17	0.8	7.0	− 1.0
−48	3.1	62	720	0.53	33	1.6	1.5	− 6.5
−50	2.5	50	580	0.72	36	1.8	0.7	−11
−52	2.0	40	460	0.90	36	1.8	0.2	−19

The diodes are then just on the verge of conduction, and the attenuation factor proves to be approximately 0.53.

One final point is perhaps worthy of note. The fact that the control voltage V_4 can be 5.8 V or more does not mean that the voltage across D1 or D3 can reach this value. Nor does it mean that the voltage on C7 or C8 will swing by this much. C7 and C8 are more than 30 times greater than C6, so the effect of a large swing in control voltage will be minor 'charge-pumping' from C8 to C7, followed by the charge flowing back from C7 to C8 through D2 and D4. It is this flow which keeps D2 and D4 in conduction.

The DNL can be switched out of operation entirely by closing S1.

Construction and alignment

The p.c. board and component layout are shown in figure 7. The power supply can be relatively simple. The required 12 . . . 20 V at 15 mA can usually be derived from the main equipment. If this is not possible for some reason, a simple supply as shown in figure 8 will suffice.

For the unit to operate properly, it is essential that the load impedance at the output of the DNL is greater than 20 kΩ. This should not create any real problems with modern equipment.

It is even more essential, however, that the input signal to the DNL is at the correct level. As Table 1 shows, the noise level at the input should be 2 . . . 3 mV. If the noise is at a lower level the DNL will still work, but it will reduce the low-level, high-frequency portion of the audio signal more than is necessary. On the other hand, if the noise is at too high a level, the DNL will not operate at all!

The S/N ratio of the average cassette recorder will be 46 . . . 48 dB. The requisite noise level (2 . . . 3 mV) thus corresponds to a nominal signal level of approximately 500 mV; if the recorder output is higher than this, an attenuator will have to be added at the DNL input. A simple solution is to use a 100 k preset potentiometer, as shown in figure 9a; alternatively, a 220 k preset pot can be included in series with the input (figure 9b).

If there is any doubt as to the correct setting of the input level, a simple alignment procedure is as follows:

– open switch S1.
– connect a dc voltmeter (10 . . . 50 kΩ/Volt), set at its highest sensitivity, between the D1-D2-C7 junction and the D3-D4-C8 junction; + to C7, – to C8. The meter should read 0 V.
– play an erased tape. The tape noise may cause the meter to show a slightly higher reading.
– set the input level to the DNL, by adjusting P1, until the pointer just starts to move up from 0 V.

The input level is now correctly set, so the meter can be removed. The DNL unit should now operate satisfactorily.

If a preset potentiometer is used instead of R19, this may also need adjustment:

– with S1 open, play an erased tape.
– set R19 for minimum noise at the output.

From the above it will also be obvious that the input signal to the DNL unit must be at a fixed level. This means that it must come straight from the playback preamplifier, before the tone and volume controls. Most cassette recorders have such an output; if this is not the case, the signal can usually be taken from the top of the volume control potentiometer. This is illustrated in figure 10.

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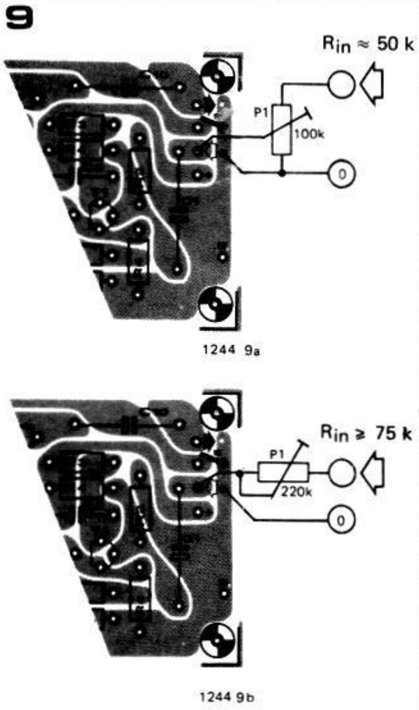

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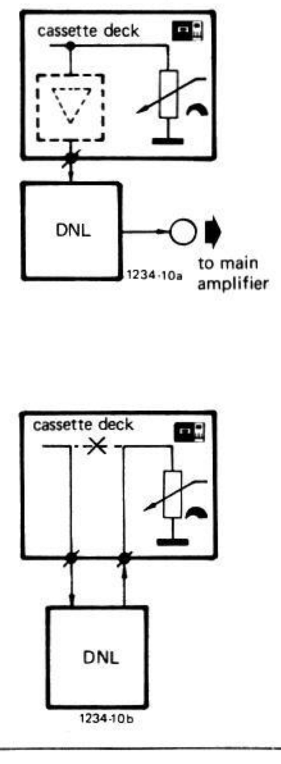
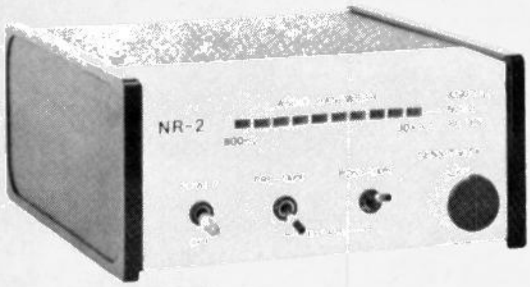


Adaptive Noise Filter



Simple to construct and operate, this dynamic variable-cutoff low-pass filter removes the snap crackle, and pop from your favorite records and tapes.

TIM SKORMOND AND GENE GARRISON

MANY POPULAR NOISE-REDUCTION SYSTEMS are complementary in nature; i.e., the signal is processed by some compression technique prior to being recorded and then decoded by the equivalent expansion technique during playback. This leaves a large quantity of older recorded material, both disc and tape, unable to benefit from the noise-reduction circuits built into many hi-fi systems. Since many of us have treasured recordings that fall into this category, the need exists for a single-ended, add-on Adaptive Noise Filter that will permit us to enjoy these recordings at the same quality level we expect from encoded material.

A great deal of the noise in tape and FM sources is located in the frequency band from about 1 kHz to 7 kHz. This frequency band is also the area in which the human ears are most sensitive. Fortunately, when musical signals are also present in this same frequency band, at a few dB higher level, they effectively mask the noise. This makes it possible to use a variable-frequency low-pass filter in the audio channel to reduce the noise without audibly affecting the dynamics of the desired signal. If no signal is present, between musical passages for instance, a low-pass filter with a cutoff frequency of 800 Hz will substantially reduce the level of the audible noise. When music of sufficient amplitude is present at frequencies above 800 Hz, the filter opens to pass the music, and since the noise and music occupy the same spectral area, the noise is masked and the music faithfully reproduced.

About the circuit

A block diagram of a suitable circuit is shown in Fig. 1. The circuit consists of three major sections: a signal path, a control path and an LED bar-graph display.

The signal path is straightforward, consisting of two (one per channel) current-controlled low-pass filters with 6-dB-per-octave slopes. The two filters are designed to roll off at a corner frequency of 800 Hz with no input and at a corner frequency of 30 kHz when fully driven by the common control voltage.

Figure 2 gives a detailed look at the actual current-controlled filter circuits. Most existing variable filter designs use one or more field-effect transistors as voltage-controlled resistors in an active or passive filter configuration. This can lead to two troublesome problems. One is the need for some means of offsetting the difference in pinch-off voltages between individual FET's. A trimmer from each gate to a fixed bias voltage must be adjusted to calibrate each device. The second problem is the modulation of the drain-to-source resistance with moderately large input signals. This modulation quite often results in excessive distortion at the filter output and, in some cases, will require another trimmer from each gate to minimize the effect.

Both of these pitfalls are successfully avoided in this circuit by using a new operational transconductance amplifier (OTA) as the controlled device. National Semiconductor has recently introduced the LM13600, a dual OTA with Darlington buffers and input linearizing diodes in a single 16-pin DIP package. The linearizing diodes compensate for the logarithmic characteristics of the input stage of the OTA, enabling it to pass relatively large signals with low distortion. (OTA circuits in the past have required very small input signals to obtain respectable distortion levels.)

The transconductance of the OTA's is set by the amplifier bias current (supplied to pins 1 and 16 in the case of the LM-13600). The signal path of Fig. 2 uses this variable transconductance (conductance being the inverse of resistance) to implement a current-controlled filter (CCF).

The filter cutoff frequency is initially set at 800 Hz by the current through

TABLE 1—SPECIFICATIONS (typical)

Attack time (to within 10% of final value):	1 ms
Decay time (to within 10%)	50 ms
Minimum bandwidth:	800 Hz
Maximum bandwidth:	30 kHz
THD (1 kHz, 1 V_{RMS}, max. sensitivity):	0.11%
S/N ratio re: 1 V_{RMS}:	
800 Hz, BW, 20 Hz to 20 kHz:	−88 dBV
CCIR/ARM weighted:	−98 dbV
30 kHz, BW, 20 Hz to 20 kHz:	−85 dBV
CCIR/ARM weighted:	−87 dBV
Effective noise reduction (CCIR/ARM weighted cassette tape noise):	14 dB

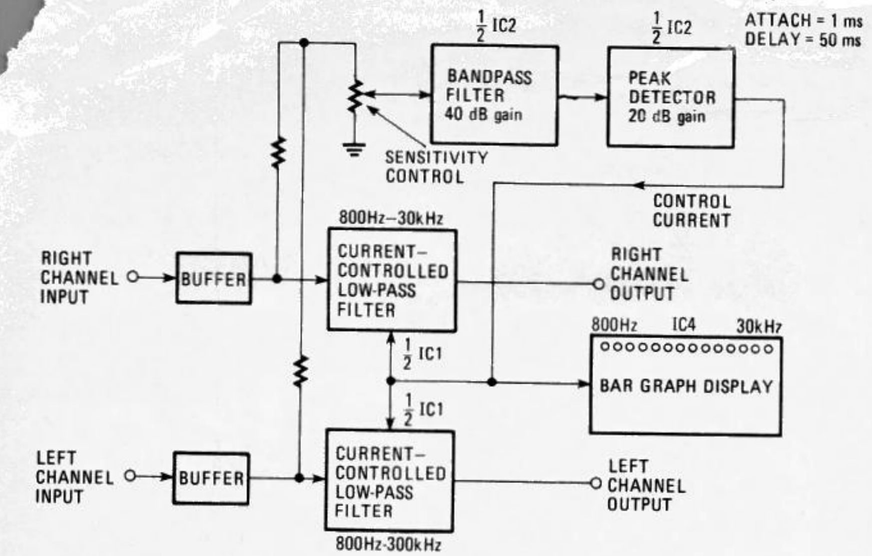


FIG. 1—BLOCK DIAGRAM of the adaptive noise filter. Audio signals are summed into the bandpass filter and then rectified to develop a control current that varies with noise in the signals.

R15. Additional current supplied through R16 by the control path circuitry increases this −3-dB frequency to as high as 30 kHz, this bandwidth being directly proportional to the applied current. Figure 3 shows the filter bandwidth vs. the amplitude of an 8-kHz input signal.

The control path consists of two basic stages: a bandpass filter with 40-dB gain and a specially configured peak detector with 20-dB gain. The LM387 operational amplifier was chosen because of its high-gain capability at 20 kHz and its high slew rate (required at the output of the peak detector).

The left and right input signals are summed together by resistors R17 and R18. Capacitor C7 provides a rolloff above 16 kHz, while a rolloff below 1.6 kHz is provided by C8. This low-level input signal is then amplified by the first half of IC2, whose gain is attentuated below 4.8 kHz by C10 with R20. R22, C11, C12, L1 and C13 couple the signal into the peak detector and provide additional frequency shaping, L1 being tuned to provide a filter notch at 19 kHz for use with FM stereo sources. Figure 4 shows the response of the control path.

R-E TESTS IT

LEN FELDMAN

THE NR-2 NOISE FILTER WAS EVALUATED IN our laboratory, both for its measured characteristics and its performance as a single-ended noise reduction filter. Its principle of operation depends upon signals passing through variable-bandpass filters whose cutoff frequencies are made to vary between 800 Hz and 30 kHz, depending upon a control current derived from the original audio signal. While this is certainly not an original concept, the execution of the design is excellent. Attack time (confirmed as approximately 1 millisecond) and decay times (around 50 milliseconds) are ideally chosen so that, in use, the action of the filter is almost inaudible—except of course for the great audible improvement in noise. The NR-2 can be used to reduce noise in program sources such as records, weak FM signals and, most significantly, playback of cassette tapes that have not had the benefit of a two-sided noise reduction system such as Dolby or dbx encoding and decoding.

Figure 1 illustrates how different *levels* of swept signals (from 20 Hz to 20 kHz) affect overall response of the device. We see that when the device "senses" low-level signals

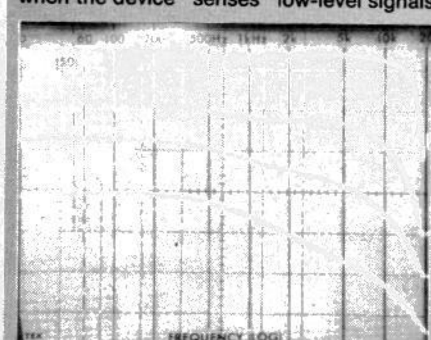


FIG. 1

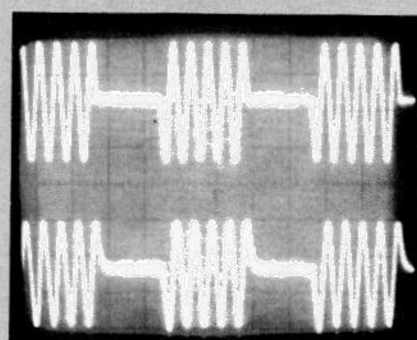


FIG. 2

(equivalent to *noise*) bandwidth is restricted. On the other hand, when high-amplitude signals, corresponding to musical signal content, are fed through the device, bandwidth is fully restored so that overall frequency response is virtually flat to 20 kHz.

Figure 2 illustrates the attack and decay times of the device. A tone burst, consisting of several cycles at 1 kHz, was fed through the filter. Upper trace is the input signal, while lower trace is the output signal. Note that the first cycle of each burst, whether negative going or positive going, is accurately reproduced at the output, while the *last* cycle in each burst is somewhat retarded in the output trace, indicating the slower but desired decay time of the device.

Table 1 summarizes our measured findings as compared with the author's claimed specifications.

TABLE I

SPECIFICATIONS	DESIGNERS' CLAIM	MEASUREMENTS
Attack Time (ms)	1.0	0.8
Decay Time (ms)	50.0	40.0
Minimum Bandwidth (Hz)	800	800
Maximum Bandwidth (kHz)	30	30
THD (1V, Max. Sensitivity) (%)		
1 kHz	0.11	0.12
20 Hz	N/A	0.3
20 kHz	N/A	0.11
S/N (re: 1V)		
800 Hz B.W.	88 dB (98 Wt'd)*	90 dB (95 W'td)
30 kHz B.W.	85 9B (87 W'td)*	85 dB (90 W'td)
Effective Noise Reduction (tested with cassette tape)	14 dB (Wt'd)*	12 dB (Wt'd)*

* Small differences in weighted results arise from the fact that the author used CCIR/ARM weighting whereas we used IHF "A" weighting, but, as can be seen, unweighted results are as good or better than claimed.

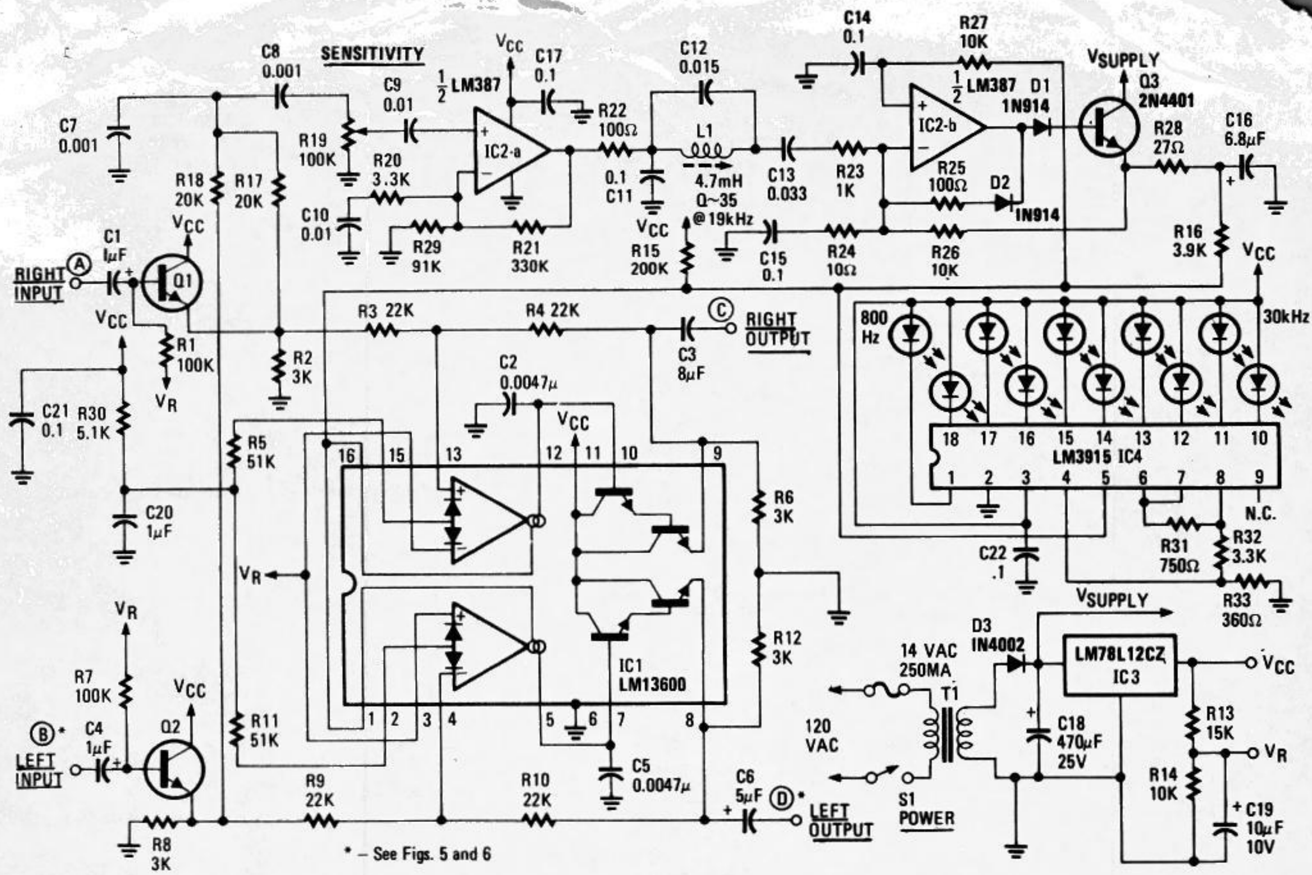


FIG. 2—SCHEMATIC DIAGRAM of the current-controlled variable-bandwidth low-pass filter circuits. Bandwidth is controlled by a current developed from the input signal.

The peak detector amplifies the AC output from the frequency-selective control amplifier and converts it into a DC voltage at the emitter of Q3. Resistor R28 sets the system attack time or charging rate into the peak-detector storage capacitor, C16. Capacitor C16 is discharged at a much slower rate (decay time) by the current drains of the two CCF's through resistor R16. Resistor R27 sets the initial no-signal DC level at the peak-detector output.

The attack and decay times and response characteristics were chosen with great care as they determine the subjective effects of the filter. An attack time of 1 ms was selected because it is fast enough to accommodate the response time of the human ear and yet slow enough to be relatively insensitive to "clicks" and "pops" on the record surface. The decay time was set at 50 ms so the filter could close quickly after the passing of high-frequency musical information and thus pass a minimum of noise. A shorter decay time would cut into the natural reverberation time of some program sources, making them sound "sterile" or flat.

The third section of the Adaptive Noise Filter is the bandwidth bar-graph display. A bar graph was used instead of a meter because of the millisecond response time of the control signal. The National LM3915 bar-graph display driver was chosen as it requires only a few external parts and contains all the necessary circuitry for a 10-point logarithmic bar-graph display. The common control voltage at the top of R16 has upper and lower limits of 9.3 VDC and 1.1 VDC, respectively. Therefore, the upper and lower limits of the LM3915 are set accordingly at pins 4 and 6 (internal logarithmic resistor string between these two pins sets the DC levels at which each of the internal comparators drives its associated LED). The left-hand LED corresponds to an 800-Hz bandwidth and the right-hand LED corresponds to a 30-kHz cutoff. The LED's between these two extremes represent steps of approximately 1.5 times the frequency display by the preceding step.

A logarithmic display was selected because it was most indicative of the audible

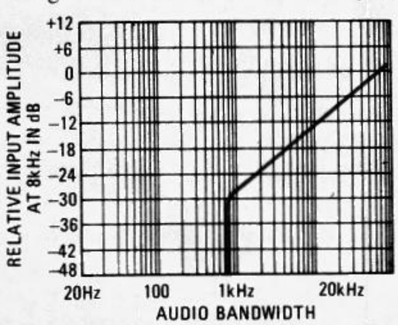


FIG. 3—FILTER BANDWIDTH and how it varies with the input of an 8-kHz signal.

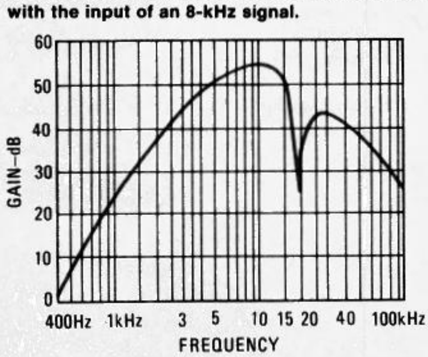


FIG. 4—RESPONSE of the final control path sensitivity vs. input frequency.

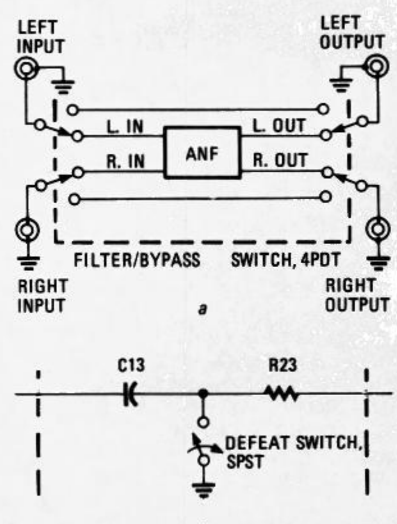


FIG. 5—CONTROL SWITCHING illustrating how the adaptive noise filter can be switched in and out of play-only circuit.

PARTS LIST

All resistors 5%, ¼ watt unless otherwise noted

R1, R7—100,000 ohms
R2, R6, R8, R12—3000 ohms
R3, R4, R9, R10—22,000 ohms
R5, R11—51,000 ohms
R13—15,000 ohms
R14, R26, R27—10,000 ohms
R15—200,000 ohms
R16—3900 ohms
R17, R18—20,000 ohms
R19—100,000 ohms, miniature pot, audio taper (Clarostat 389 N 100K-Z)
R20, R32—3300 ohms
R21—330,000 ohms
R22, R25—100 ohms
R23—1000 ohms
R24—10 ohms
R28—27 ohms
R29—91,000 ohms
R30—5100 ohms
R31—750 ohms
R32—360 ohms

Capacitors

C1, C4, C20—1 µF, 16 volts electrolytic, radial leads
C2, C5—.0047 µF, 50 volts, Mylar, 10%
C3, C6—5 µF, 10 volts, electrolytic, radial leads
C7, C8—.001 µF, 50 volts, ceramic disc
C9, C10—.01 µF, 50 volts, ceramic disc
C11, C14, C15, C17, C21, 022—0.1 µF, 50 volts, Mylar
C12—.015 µF, 50 volts, Mylar, 5%
C13—.033 µF, 50 volts, Mylar
C16—6.8 µF, 25 volts, tantalum, 10%, radial leads
C18—470 µF, 25 volts, electrolytic, radial leads
C19—10 µF, 10 volts, electrolytic, radial leads

Semiconductors

D1, D2—1N914
D3—1N4002
Q1, Q2, A3—2N4401
IC1—LM13600 dual operational transconductance amplifier (National)
IC2—LM387N dual low-noise preamplifier (National)
IC3—LM78L12CZ (National)
IC4—LM3915N logarithmic bar-graph display driver (National)
LED1-LED10—NSL57124 rectangular LED for bar graph (National)

Miscellaneous

S1—miniature SPST toggle switch
S2, S3—miniature 4PDT toggle switch
T1—power transformer, 14 VAC, 250 mA (Triad F-112X)
L1—adjustable inductor, 4.7 mH, Q = 35 at 19 kHz (TOKO CLN20 740 HM)
J1-J8—panel-mount RCA-type phono
F1—¼-amp slow-blow fuse
Fuse holder, line cord, PC boards, control knobs, hardware

The following parts are available from Advanced Audio Systems, PO Box 24, Los Altos, CA 94022:

DX-244 (NR-2) complete kit including case—$69.95

DX-245 (NR-2) main and display PC boards—$19.95

DX-247 (NR-2) component kit; includes D1, D2, D3, IC1, IC2, IC3, IC4, Q1, Q2, Q3 and L1—$27.50

California residents add state and local taxes, as applicable.

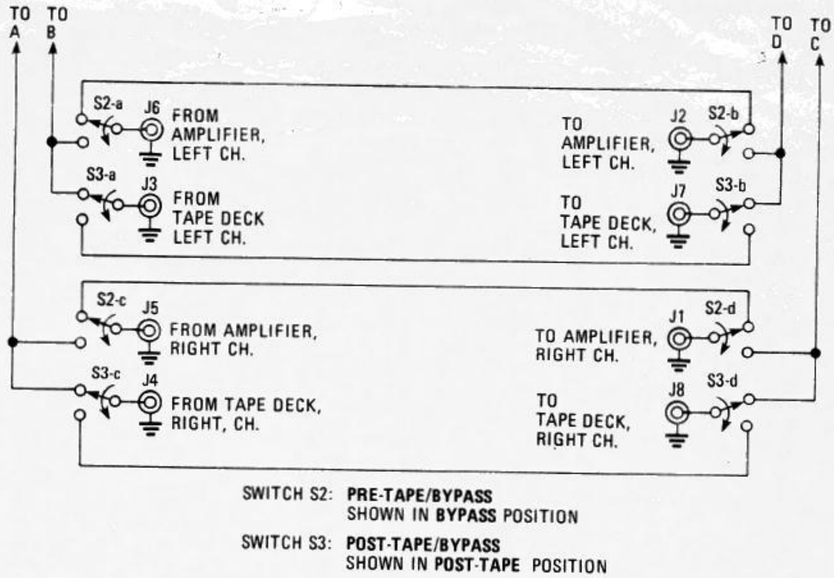
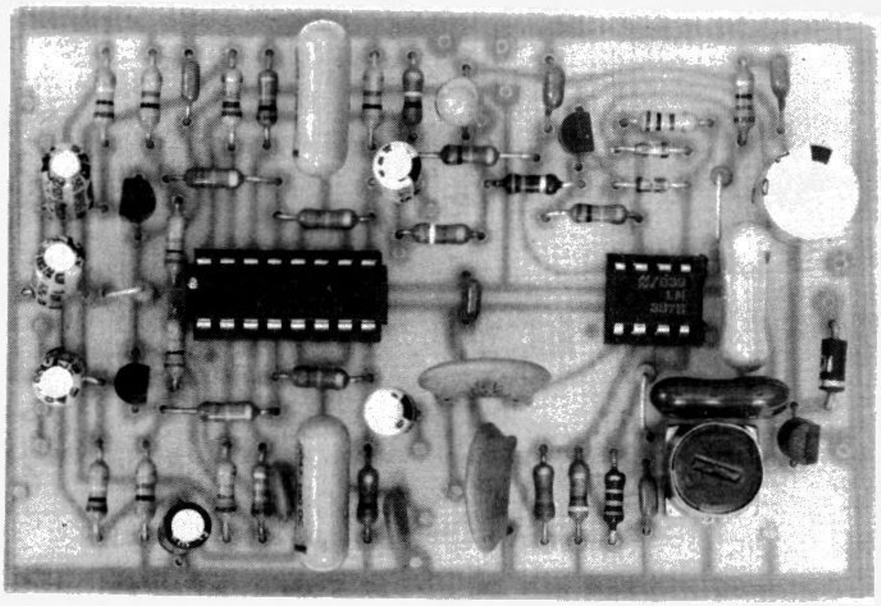


FIG. 6—IDEAL SWITCHING ARRANGEMENT is simple and flexible. User must be careful not to place both switches in non-bypass positions simultaneously.



TOP VIEW OF THE PC BOARD shows the locations of most of the components. Parts not shown include the bar-graph display and power supply.

action of the filter. It should be noted that the LED bar graph does not indicate signal level, but rather the instantaneous bandwidth of the two filters and, as such, should not be used as a signal-level indicator.

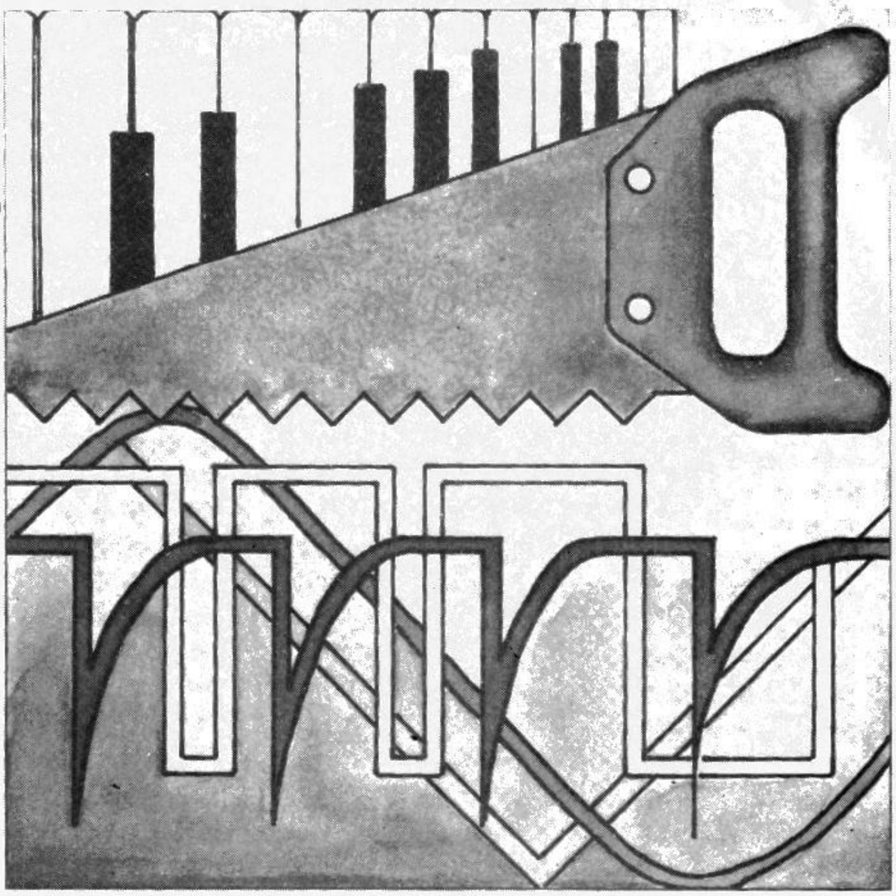
If the Adaptive Noise Filter is used for only one source, such as for tape playback, then a BYPASS switch will be the only switching needed in addition to the power switch. Figure 5-a shows a suitable arrangement. Alternatively, the filter may be defeated by switching the junction of C13 and R23 (Fig. 2) to ground, as shown in Fig. 5-b, thereby forcing the filter to a constant 30-kHz bandwidth. This approach has the advantage of requiring only a single-pole switch as opposed to the four-pole switch and associated wiring required for the straight-wire bypassing.

A more flexible system, as used in the prototype model, can be obtained by providing for insertion of the filter either before or after the tape deck (record or playback). The rather unusual switching scheme shown in Fig. 6 was selected for its relative simplicity. A setup having a PRE TAPE/POST TAPE and a FILTER/BYPASS switch would require one four-pole and one eight-pole switch. The scheme of Fig. 6 only requires two four-pole switches, but the user should be careful not to place both switches in the non-bypass positions simultaneously.

Printed-circuit construction techniques are used in the assembly of the prototype adaptive noise filter. Foil patterns for the printed-circuit boards used for the filter and associated display board will be published next month along with complete construction details and calibration and operating instructions.

continued next month

24 dB-VCF



In response to requests from readers who have built the Formant synthesiser the following article presents a design for a voltage-controlled filter whose slope is considerably steeper than that of the original VCF, in fact 24 dB/octave as opposed to 12 dB/octave. The filter offers a choice of highpass or lowpass modes and slopes of 6, 12, 18 or 24 dB/octave.

C. Chapman

VCFs with an extremely steep slope seem to have a particular appeal for most synthesiser enthusiasts because of the greater range of tonal possibilities that they offer. Formant users are evidently no exception to this rule judging by the number of requests for a 24 dB/octave VCF. Of course, the filter described here is not restricted to use with the Formant synthesiser, but may also be used with other synthesiser designs.

New possibilities

It should be stated at the outset that the 24 dB VCF does not render the existing 12 dB design obsolete. On the contrary, the two filters are complementary to one another and can be used in combination to provide greatly increased possibilities for tailoring the harmonic structure of the sounds produced by Formant.

For example, the 12 dB VCF can be used in the bandpass mode together with the steep filtering of the 24 dB VCF to produce selective tone coloration. The two filters can be controlled by the same envelope shaper or by different envelope shapers, and may be connected in cascade or in parallel. The latter arrangement offers several interesting possibilities. For example, hard, metallic sounds can be produced by applying a short, steep envelope voltage to the 12 dB VCF and a longer, shallower contour to the 24 dB VCF.

If the filter inputs are connected in parallel then interesting effects may be obtained by connecting one VCF output to one input of a stereo amplifier and the other VCF output to the other input. This gives rise to a very distinctive dynamic amplitude characteristic and stereo imaging, particularly if the two VCFs are controlled by different envelope shapers.

The audible differences between the 12 dB VCF and the 24 dB VCF are quite prominent. The 12 dB VCF produces sounds that are distinctly 'electronic', which can have a slightly fatiguing effect on the listener during extended playing sessions. The sounds produced by the 24 dB VCF, on the other hand, are much more 'natural', and can be listened to for extended periods without fatigue. This effect is probably due to the more severe filtering of higher harmonics which the 24 dB VCF provides when used in the lowpass mode, since these harmonics tend to make the sound of the 12 dB VCF much more shrill than that of the 24 dB VCF.

The effect of the steeper filter slope of the 24 dB VCF is illustrated in figure 1, which shows the different outputs from the 12 dB VCF (dotted line) and 24 dB VCF (continuous line) when fed with a sawtooth waveform. It is apparent that, due to the almost complete removal of the harmonics of the sawtooth, the output of the 24 dB VCF is practically a sinewave, whereas the original waveform is still apparent at the output of the 12 dB VCF since the harmonics are only partially removed.

It is clear from the foregoing that a 24 dB VCF greatly extends the musical possibilities of a synthesiser and is virtually a must for the serious user.

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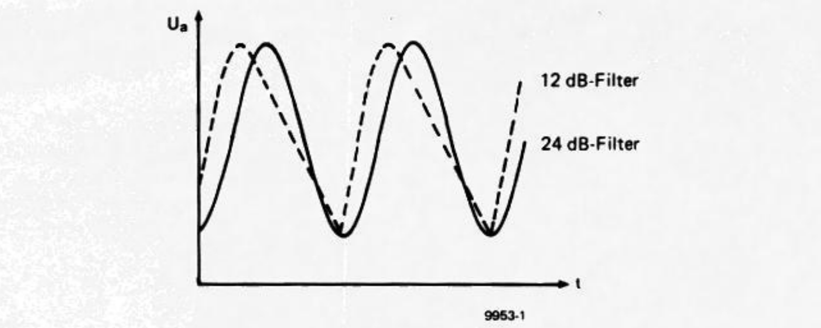

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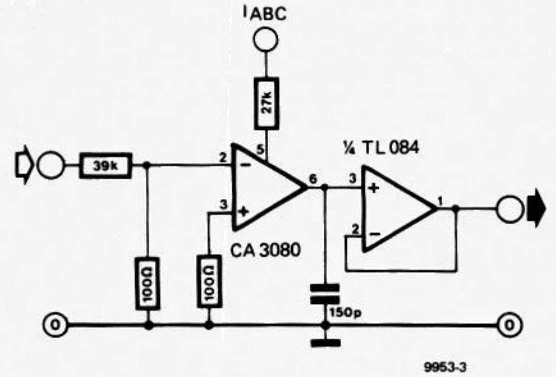

Figure 1. This illustrates the difference between the outputs of a 12 dB/octave VCF and a 24 dB/octave VCF having the same turnover frequency, when fed with a sawtooth input. The 24 dB VCF removes practically all the harmonics giving a sinewave output, whereas the original waveshape is still distinguishable at the output of the 12 dB VCF.

Figure 2. The basic filter section of the 24 dB VCF is the same as that of the 12 dB VCF, i.e. an OTA integrator followed by a FET op-amp buffer.

Figure 3. The highpass function is obtained by connecting the 6 dB lowpass section in the feedback loop of an operational amplifier.

Figure 4. To obtain a 24 dB/octave filter, four 6 dB/octave sections are cascaded.

Design of the 24 dB VCF

Most 24 dB VCFs are variations on the heavily-patented design by R.A. Moog, which has been around for a number of years. However, thanks to the advent of inexpensive IC OTAs (Operational Transconductance Amplifiers) a more versatile design than Moog's is now possible, which can be operated in highpass or lowpass modes with slopes of 6, 12, 18 or 24 dB/octave. Even greater slopes than 24 dB would be possible, but experiments have shown that a greater slope does not result in a corresponding increase in tonal quality.

The design of the basic filter section shown in figure 2 is very similar to that of the 12 dB VCF, which was described in detail on page 12-29 of Elektor 31, December 1977. However, advantage has been taken of recent developments in FET op-amp technology to simplify the design. As described in that article the basic filter section is an integrator or 6 dB/octave lowpass section consisting of an OTA driving a capacitor. The voltage/current transconductance (g_m) of the OTA can be varied by an external control current and hence, via an exponential voltage/current converter, from an external control voltage. This control current alters the time constant of the integrator and hence the turnover frequency of the filter section.

The output current of the OTA must all flow into the capacitor, otherwise the integrator characteristic will be less than ideal. This means that the output of the OTA must be buffered by an amplifier with a high input impedance. In the

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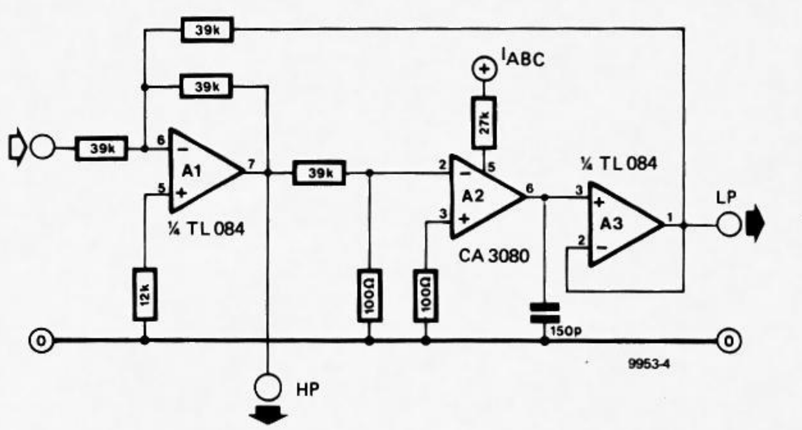

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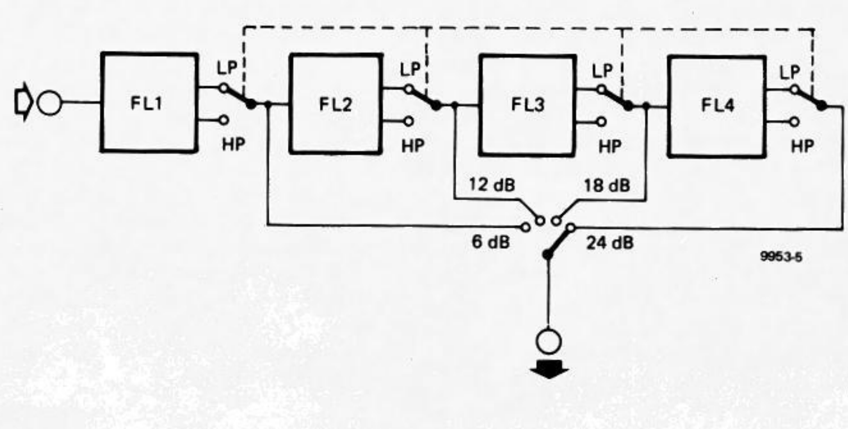

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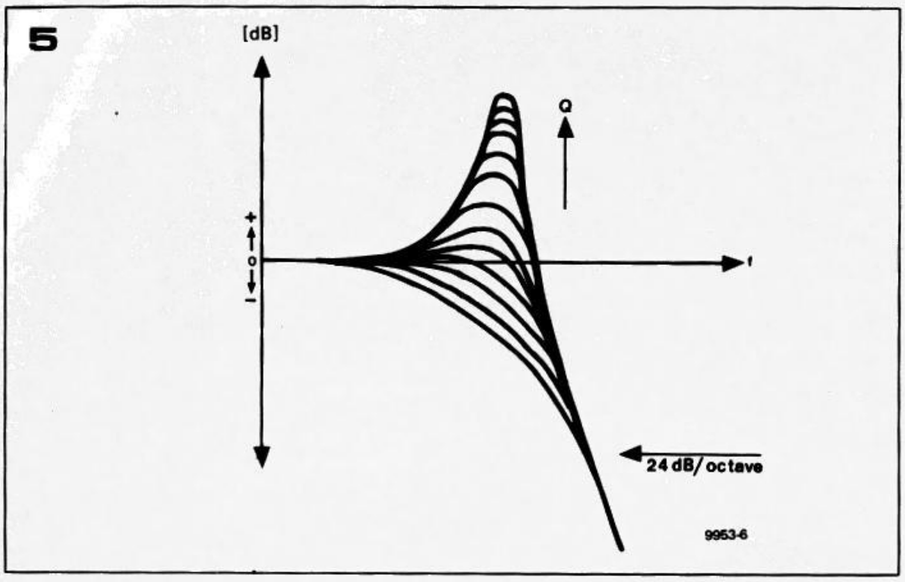

Figure 5. Positive feedback around the entire filter allows the response to be boosted about the turnover frequency. The degree of boost can be varied by a 'Q' control.

Figure 6. Block diagram of the 24 dB/octave filter, showing how the Q control is incorporated.

Figure 7. Complete circuit of the 24 dB VCF. The exponential voltage/current converter is identical to that used in the 12 dB VCF.

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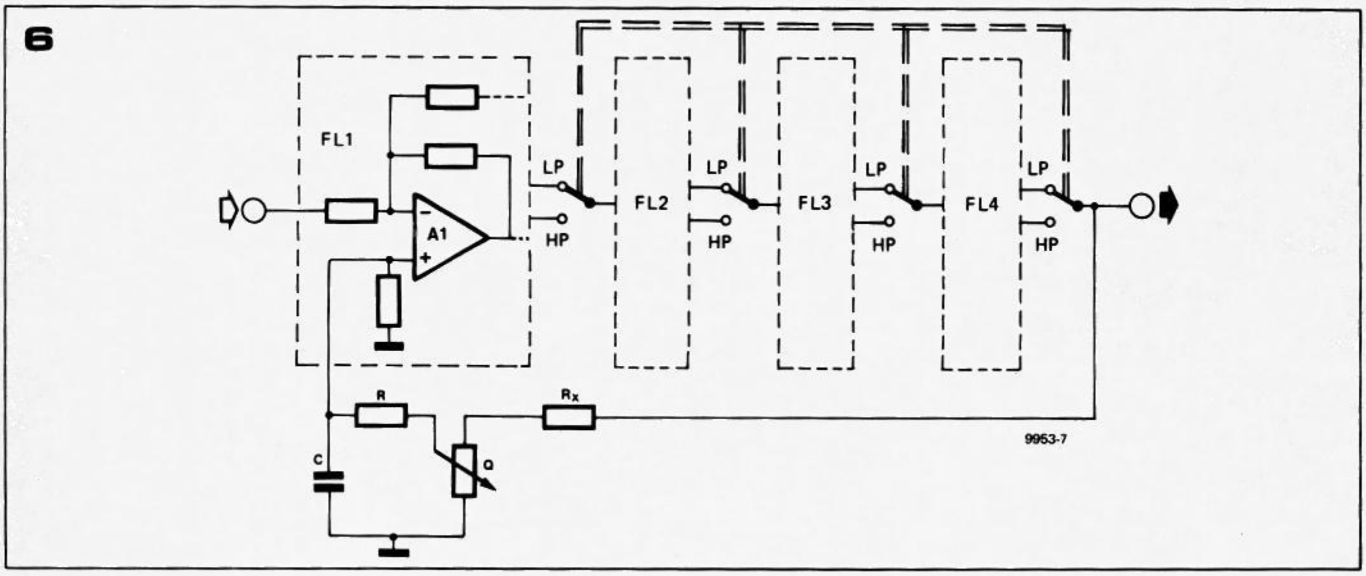


12 dB VCF this was achieved by using a discrete FET source follower and a 741 op-amp. Fortunately, relatively inexpensive quad FET op-amps such as the Texas TL084 are now available. The use of one of these ICs greatly simplifies the design and obviates the need to select FETs, which is rather a chore when one considers that the 24 dB VCF uses four integrator stages.

Highpass function

The highpass mode of the filter is achieved by connecting the 6 dB/octave lowpass section in the negative feedback loop of an operational amplifier, A1, as shown in figure 3. A highpass filter response is then available at the output of A1 whilst a lowpass response is simultaneously available at the output of A3. Of course, this arrangement gives only a 6 dB/octave slope per section, and in order to obtain a 24 dB/octave filter four filter sections, built according to the circuit of figure 3, must be cascaded as shown in figure 4. Switching at the output of each section allows selection of highpass or lowpass mode, whilst a 4-position switch allows 1, 2, 3, or 4 filter sections to be switched in to give 6-, 12-, 18-, or 24 dB/octave slopes respectively.

It is apparent that this arrangement is different from the two-integrator loop or state-variable filter which formed the basis of the 12 dB/octave filter. In the 12 dB/octave filter, lowpass, highpass, bandpass and notch modes were available simultaneously at various points in the circuit, though in fact only one function at a time could be selected at the output.

An interesting effect, shown in figure 5, can be obtained with the 24 dB VCF if a feedback loop is connected from the output of the cascaded filters to the non-inverting input of the first stage as illustrated in figure 6. Due to the phase shift around the turnover frequency this causes positive feedback, which boosts the gain of the filter around the turnover frequency as shown in figure 5. The degree of boost is adjustable by means of a 'Q' control. The choice of R_X is important as too much feedback would cause the circuit to oscillate, so the value of R_X is a compromise between stability and a reasonable degree of boost.

Complete circuit

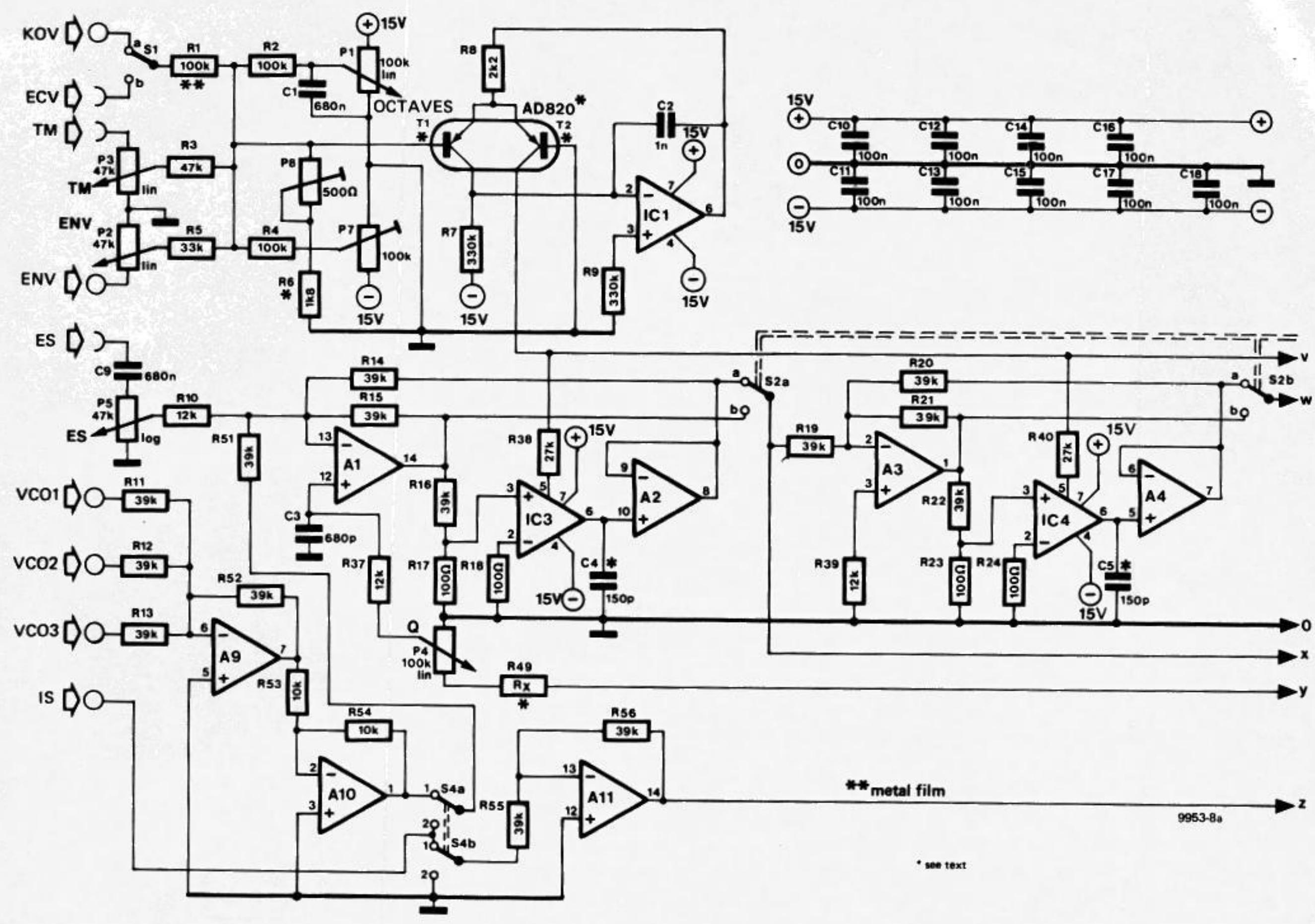
The complete circuit of the 24 dB VCF is given in figure 7. The exponential converter, constructed around T1, T2 and IC1, is identical to that used in the 12 dB VCF and gives the same 1 octave per volt characteristic to the turnover frequency of the filter. The control voltage inputs are also the same as for the 12 dB VCF, and are listed in table 1.

Since the 24 dB VCF must have the option of being connected in parallel or in cascade with the 12 dB VCF, the input switching arrangements are a little complicated. A9 and A10 form a non-inverting summing amplifier for the three VCO inputs, whilst the output of the 12 dB VCF is fed in via the IS connection. With S4 in position 2 the output of A10 is disconnected, so the VCO inputs are inhibited. The output of the 12 dB VCF is fed to the input of the 24 dB VCF via S4 and R51, so that the two VCFs are in cascade.

With S4 in position 1 the output of A10 is connected to the inputs of the 24 dB VCF, whilst the output of the 12 dB VCF is routed through A11. The output of A11 and the output of the 24 dB VCF are added together in the output summing amplifier A12, i.e. the two VCFs are connected in parallel.

The four 6 dB/octave filter sections

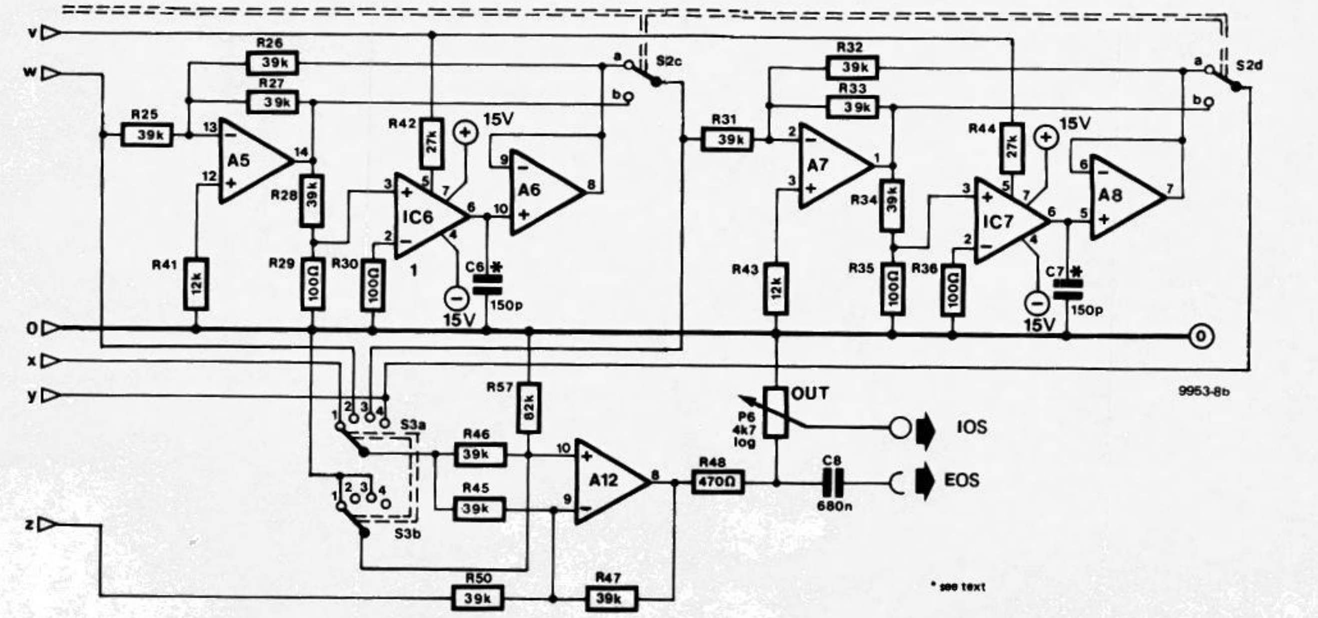
7a



7b

A1 + A2 + A3 + A4 = IC2 = TL 084
A5 + A6 + A7 + A8 = IC5 = TL 084
A9 + A10 + A11 + A12 = IC8 = TL 084

IC1 = µA 741 Minidip
IC3 = CA 3080*
IC4 = CA 3080*
IC6 = CA 3080*
IC7 = CA 3080*



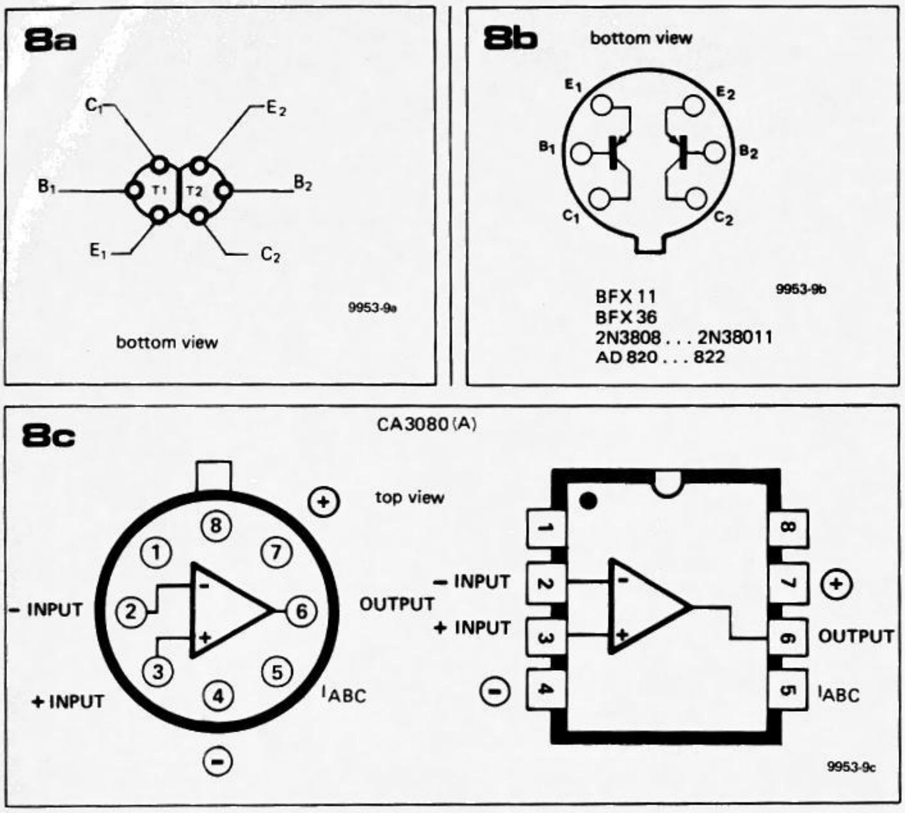


Figure 8. Pinouts for the dual transistors and CA3080.

Figure 9. Printed circuit board and component layout for the 24 dB VCF. (EPS 9953-1).

Table 1. Summary of the control functions and input/output connections of the 24 dB VCF.

comprise A1 to A8 and IC3 to IC7. The four poles of switch S2 select between highpass and lowpass modes, while S3 selects the filter output and hence the slope. The reason that S3 is a two-pole switch may not be immediately apparent, but is easily explained. Ignoring the phase shift introduced by the action of the filter, i.e. considering only signals in the filter passband, each filter section inverts the signal fed to it, since A1, A3, A5 and A7 are connected as inverting amplifiers. This means that the outputs of alternate filter sections are either in phase or inverted with respect to the input signal. To ensure that the filter output is in the same phase relationship to the input signal whatever filter slope is selected, S3b is arranged to switch A12 between the inverting and non-inverting modes to cancel the inversions produced by the filter sections.

Like the 12 dB VCF, the 24 dB VCF has two outputs, a hardwire output connection IOS and an uncommitted output, EOS, which is connected to a front panel socket.

Construction

As far as the choice of components for the 24 dB VCF goes, the same general comments apply that were made about the 12 dB VCF and the Formant synthesiser in general. All components should be of the highest quality; resistors should be 5% carbon film types except where metal oxide or metal film types are specified; capacitors should preferably be polyester, polystyrene or polycarbonate, and must be these types where specified. Semiconductors should be from a reputable manufacturer.

As with the 12 dB VCF the dual transistor may be any of the types specified in

Table 1

a) hardwired inputs (not on the front panel)	
KOV	= Keyboard Output Voltage (from interface receiver)
ENV	= Envelope shaper Control Voltage (from ADSR unit)
VCO 1,2,3	= Signals from VCOs 1, 2, 3
IS	= Internal signal from the 12 dB VCF
b) external inputs (sockets on front panel)	
ECV	= External Control Voltage (for exponential generator of the VCF)
TM	= Tone Colour Modulation input
ES	= External Signal (from e.g. noise module)
c) outputs	
IOS	= Internal Output Signal (from VCF to VCA)
EOS	= External Output Signal (socket on front panel)
d) controls	
TM	= P3; sets tone colour modulation level
ES	= P5; sets external signal level
ENV	= P2; sets envelope shaper control voltage
OCTAVES	= P1; coarse frequency adjustment
Q	= P4; sets level of peak boost around turnover frequency
OUT	= P6; sets IOS output level
e) switches	
ECV/KOV	= S1; selects external or internal control voltage input

Parts list to figures 8 and 10

Resistors:
R1 = 100 k metal oxide
R2,R4 = 100 k
R3 = 47 k
R5 = 33 k
R6 = 1k8
R7,R9 = 330 k
R8 = 2k2
R10,R37,R39,R41,R43 = 12 k
R11 . . . R16,R19 . . . R22, R25 . . . R28,R31 . . . R34,R45, R46,R47,R50,R51,R52,R55, R56 = 39 k
R17,R18,R23,R24,R29,R30, R35,R36 = 100 Ω
R38,R40,R42,R44 = 27 k
R48 = 470 Ω
R49 = 100 k (see text)
R53,R54 = 10 k
R57 = 82 k

Potentiometer:
P1,P4 = 100 k linear
P2,P3 = 47 k (50 k) linear
P5 = 47 k (50 k) logarithmic
P6 = 4k7 (5 k) logarithmic
P7 = 100 k preset
P8 = 470 Ω (500 Ω) preset

Capacitors:
C1,C8,C9 = 680 n
C2 = 1 n
C3 = 680 p (polystyrene, not ceramic)
C4,C5,C6,C7 = 150 p (polystyrene, not ceramic)
C10 . . . C18 = 100 n

Semiconductors:
IC1 = 741
IC2,IC5 = TL 084, TL 074
IC8 = TL 084, TL 074, LM 324
IC3 . . . IC6 = CA 3080, CA3080A (MINIDIP or TO; see text)
T1,T2 = AD 820 . . . 822, 2N3808 . . . 3811, BFX 11, BFX 36 (see text) or 2 x BC 557B

Miscellaneous:
31-pin DIN 41617 connector or terminal pins
S1 = SPDT
S2 = 4-pole double throw
S3 = 2-pole 4-way; index angle approx. 30°
S4 = DPDT
4 minature sockets, 3.5 mm dia.
7 13 . . . 15 mm collet knobs with pointer (to match existing synthesiser modules).

the parts list, or may be home-made by gluing together two normal transistors, though in this case thermal tracking will not be quite so good. The CA3080 should preferably be in a MINIDIP package to fit the hole spacings on the p.c. board, though the metal can type can be made to fit by splaying the leads. The pinouts for the dual transistors and the CA3080 are given in figure 8.

Although not absolutely necessary, it is a good idea to select OTA's with approximately the same transconductance, since the four sections of the filter will then have almost the same turnover frequency. The CA3080 is available in two versions, the standard version, in which the ratio between the maximum and minimum g_m is 2:1, and the CA3080A, in which the spread in g_m is only 1.6:1. A test circuit and test procedure for selecting ICs with similar g_m are given at the end of the article, and it is certainly worthwhile buying a few extra OTAs and selecting the four with the most similar g_m. The 'reject' devices are perfectly acceptable for use in the 12 dB VCF or VCA, and need not be wasted.

The other ICs in the circuit should all be TL074 or TL084 quad BIFET op-amps, although for IC8 it is only permissible to use an LM324. Thanks to the use of quad op-amps it is possible to accomodate the 24 dB VCF on a standard Eurocard-size (160 mm x 100 mm) p.c. board, although the control connections are not all on the front edge of the board. The printed circuit pattern and component layout for this board are

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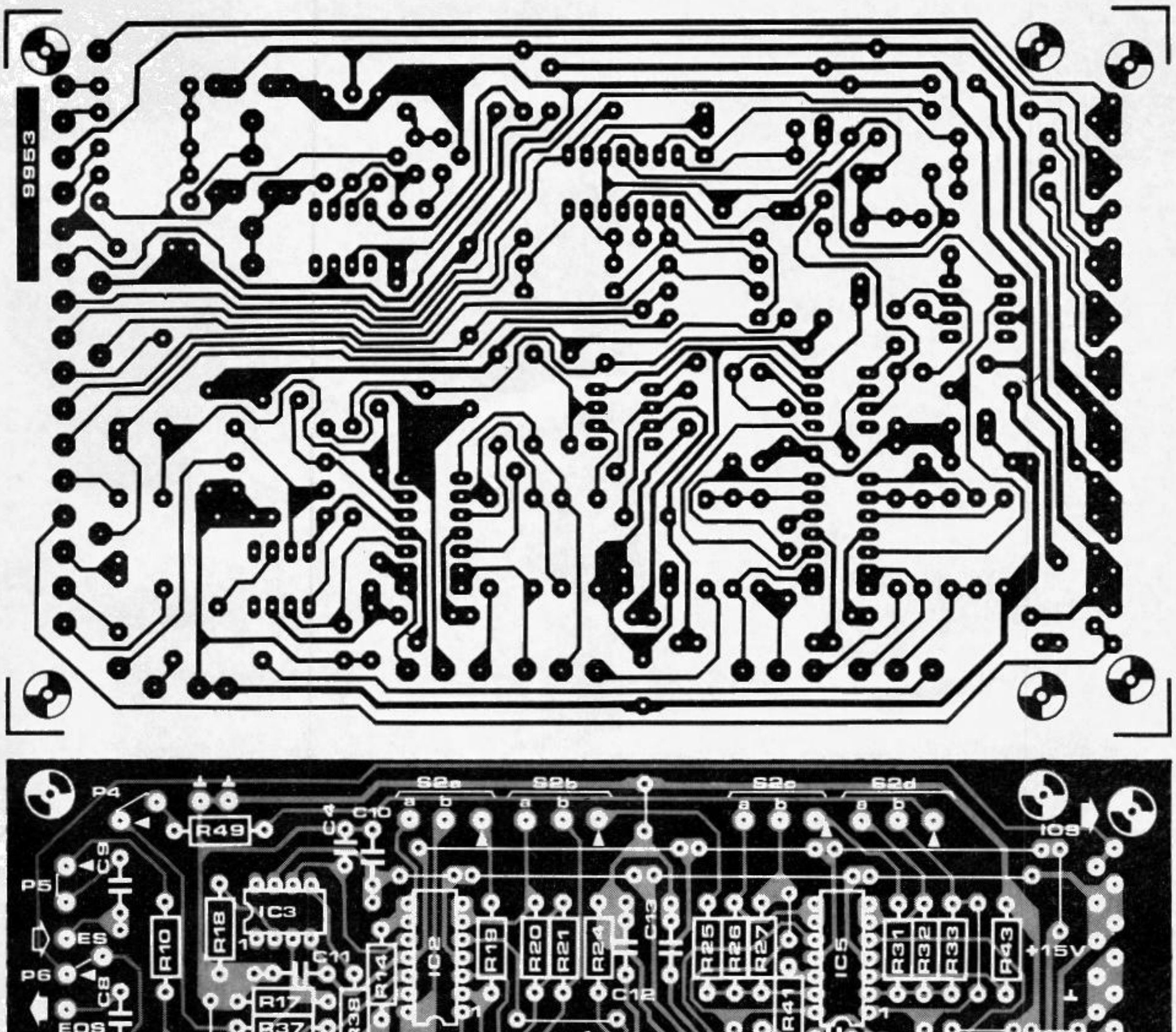

given in figure 9, while a front panel layout is given in figure 10.

Test and adjustment

To enable the exponential converter and the filter section to be tested separately they are joined by a wire link which runs across the board from T2 to a point adjacent to R15. This link should be omitted until the VCF has been tested.

To test the filter section it is necessary to provide a temporary control current. This is done by connecting a 100 k log potentiometer between −15 V and ground, with its wiper linked to the junction of R39 and R4 via a multimeter set to the 100 μA DC range. The test then proceeds as follows:

1. Turn the wiper of P4 fully towards ground, select 24 dB slope with S3 and adjust the control current to 100 μA.
2. Feed a sinewave signal into the ES socket and adjust either the sinewave amplitude or P5 for 2.5 V peak-to-peak measured on a oscilloscope at the wiper of P5.
3. Monitor the filter output on the 'scope and check the operation of the filter by varying the sinewave frequency and checking that the signal is attenuated above the turnover frequency in the lowpass mode and below the turnover frequency in the highpass mode.
4. The function of S3 should now be checked. Set S3 to the 6 dB position and S2 to the LP position. Increase the frequency of the input signal until the output of the filter is 6 dB down on (i.e. 50% of) what it was in the passband where the response was level. Now switch to 12 dB, 18 dB and 24 dB and check that the response is respectively 12, 18 and 24 dB down, i.e. is reduced to 25%, 12.5% and 6.25% of its original value. The exact results of this test will depend upon the matching of the OTAs.
5. Set the Q control, P4, to its maximum value, when the circuit should show no sign of oscillation. If the circuit does oscillate it will be necessary to increase the value R49. If it does not oscillate then the Q range can be increased by decreasing R49, taking care that instability does not occur.
6. Finally, the linearity of the turnover frequency v. control current characteristic should be checked. Adjust the input frequency until the response is a convenient number of dB down (say 6 dB). Double the control current then double the input frequency and the response should still be 6 dB down.
7. To check the exponential converter connect a 27 k resistor in series with a multimeter set to the 100 μA range between the collector of T2 and the −15 V rail. Then follow the test

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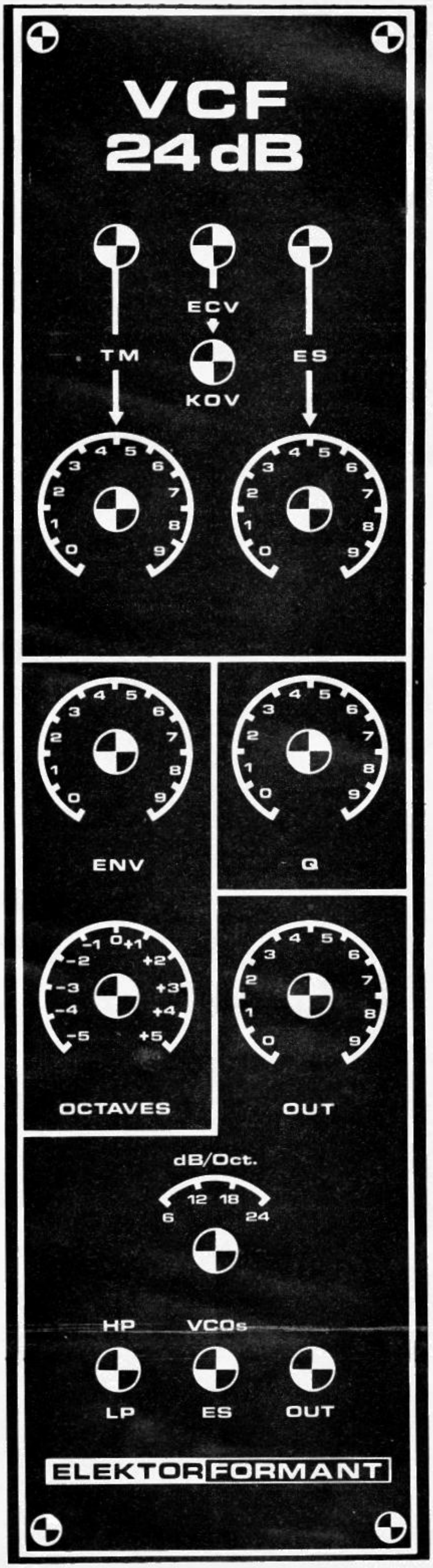

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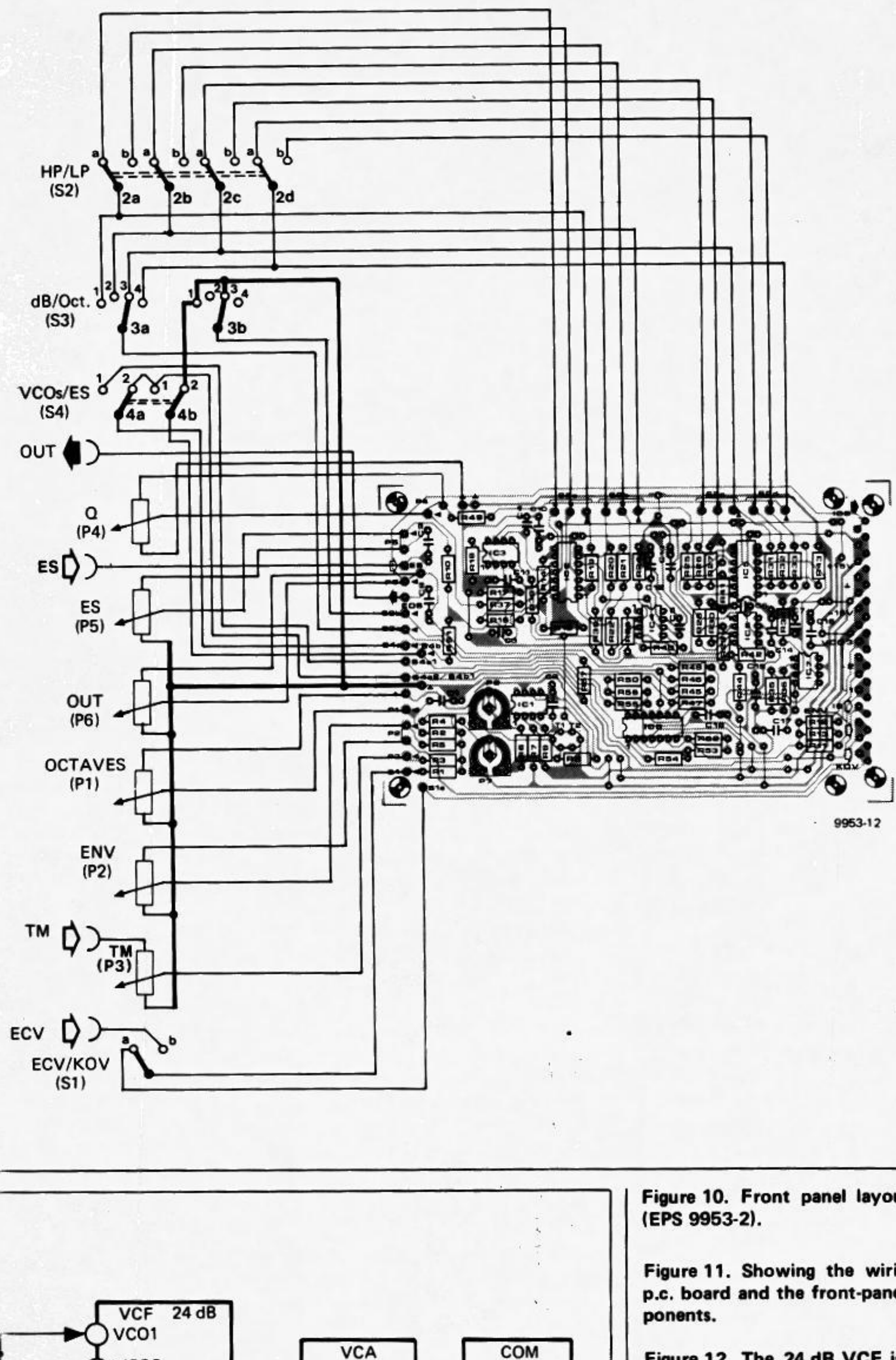

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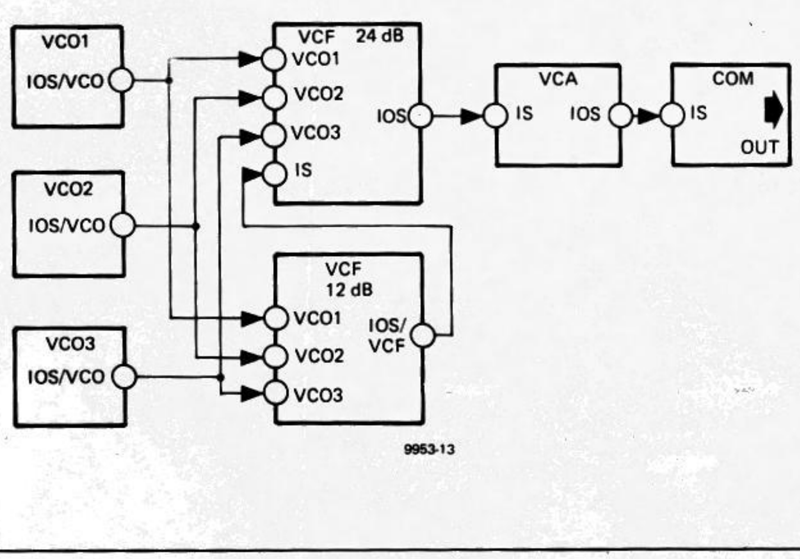
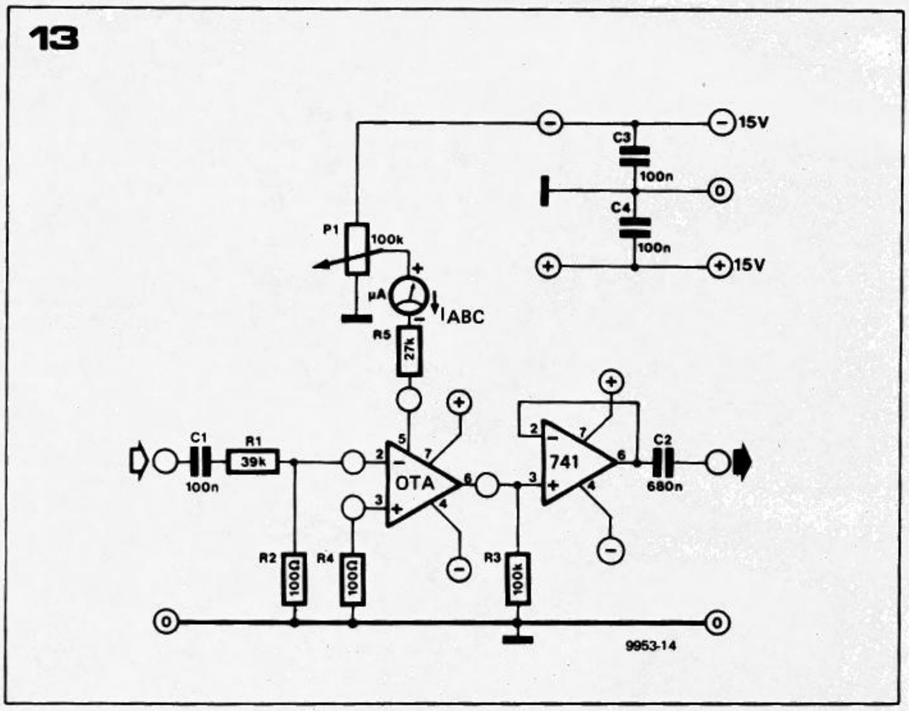


Figure 10. Front panel layout for the VCF. (EPS 9953-2).

Figure 11. Showing the wiring between the p.c. board and the front-panel mounted components.

Figure 12. The 24 dB VCF is connected into the Formant system between the 12 dB VCF and the VCA.

Figure 13. Test circuit for the selection of OTAs.



procedure given on page 12-33 of Elektor 32, column 1 line 51. The offset and octaves per volt adjustments can also be carried out using the procedure given in this issue. During the offset adjustment P4 should be set to minimum and S3 should be set to the 24 dB position. During the octaves/volt adjustment of P8 the Q control, P4, should be set to maximum, as with the 12 dB VCF.

Using the 24 dB VCF

Before the 24 dB VCF can be put to work it must first be connected into the Formant system. Fortunately, as far as the signal paths go this involves changing only two connections and adding three more. As can be seen from figure 12, the 24 dB VCF is connected between the 12 dB VCF and the VCA, so that the IOS output of the 12 dB VCF now goes to the IS input of the 24 dB VCF instead of to the VCA, whilst the VCA receives its input from the IOS output of the 24 dB VCF. The 24 dB VCF also has inputs from the three VCOs.

In addition to the signal connections the 24 dB VCF must also be provided with supply to the VCF module in accordance with the standard practice for Formant. Provision of control voltage inputs from the ADSR envelope shapers will be discussed later.

For satisfactory operation of the 24 dB VCF the correct setting of the input level is important, even more than in the case of the 12 dB VCF. On the one hand, the input level should not be so large that distortion occurs, but on the other hand it should not be so small that the signal-to-noise ratio is degraded. The 24 dB VCF is designed so that the optimum input level is obtained using three VCOs set to maximum output, with one waveform selected per VCO. If more than three VCOs are in use or more than one output waveform is selected from each VCO then the VCO output levels must be reduced. On the other hand, if only one VCO is used then the signal level may be too low. In this case it is best to patch the EOS socket of the VCO to the ES input of the VCF, since this input has approximately three times the sensitivity of the hardwired VCO inputs.

The 24 dB VCF is capable of the same basic functions as the 12 dB VCF; driven by the KOV control voltage it will operate as a tracking filter, whilst the ENV and TM inputs allow dynamic modulation of the harmonic content of the VCF output. Due to the greater slope of the 24 dB VCF the setting of the ENV level control is more critical than with the 12 dB VCF, but if correctly adjusted then subtle nuances in the tonal character of the output signal are possible.

The question arises as to which ADSR envelope shaper should be used to control the 24 dB VCF, since only two are built into the basic Formant system, and control the VCA and 12 dB VCF respectively. Because of the modular construction of Formant it is, of course, perfectly feasible to build a third envelope shaper, which is the most versatile arrangement. The alternatives are to patch one of the other ADSR outputs to the TM input of the 24 dB VCF, or to hardwire the ENV input of the 24 dB VCF to the output of the envelope shaper that controls the 12 dB VCF. This latter arrangement is probably preferable, as it allows the ADSR signal to be fed to one or both VCFs by suitable adjustment of their ENV controls and also allows the possibility of patching the output of the other envelope shaper into the TM input of either VCF.

Appendix

OTA selection procedure

Although not absolutely essential, it is well worth selecting OTAs with closely matched transconductance characteristics to ensure that the four filter sections track accurately.

A test circuit for the OTAs is given in figure 13. This should be fed with a sinewave signal of about 2 V peak-to-peak (or 0.7 V measured on an AC voltmeter) from a signal generator or from one of the VCOs. The output should be monitored on a 'scope or AC voltmeter. With a control current of 100 μA, measured on the multimeter in series with R5, the output voltage should be between 0.7 V and 1.3 V peak-to-peak. Without changing the input level or control current the OTAs to be tested should be plugged into the circuit one at a time and the output level for each OTA noted. The four OTAs whose output levels are most similar should be used in the VCF.

The circuit can also be used to check the linearity of the transconductance v. control current characteristic of the OTAs, e.g. doubling the control current should double the output of the test circuit and halving the control current should halve the output.

Dynamic Noise Limiter