chapter 6

12 dB VCF

This chapter introduces the first of the tone-shaping modules which process the 'raw' output of the VCOs to provide a wide variety of different tone colours and amplitude dynamics. The module presented here is a 12 dB per octave voltage-controlled filter (VCF) which is used to tailor the frequency spectrum of the VCO signal.

Before looking at the VCF circuit in detail, it is worth examining the ways in which the VCF is used. Four filter functions are available. A lowpass filter with a rolloff of −12 dB per octave above the turnover point, a highpass filter with a rolloff of −12 dB per octave below the turnover point, a bandpass filter with variable Q and minimum slope of −60 dB per octave on either side of the centre frequency, and a notch filter. The turnover point – or centre frequency in the case of the band filters – is the same for all four filter functions, and can be varied by the application of a control voltage.

Lowpass filter

The simplest use of the VCF is what might be called static tailoring of a VCO output using the KOV output of the keyboard to control the VCF. Suppose (to give a simple example), it is required to filter out a large proportion of the harmonics of the squarewave signal to produce a flutelike tone. The lowpass function of the VCF would be used and the turnover point would be set so that when a particular key was depressed the desired tone colour was obtained. If a higher note is depressed then the VCO pitch will increase. However, since the KOV output is also applied to the VCF the turnover point of the VCF will increase with the VCO frequency, so that it always remains in the same octave relationship to the VCO frequency. The same harmonic structure of the output waveform is thus maintained, – i.e. the VCF is being used as a tracking filter.
If the VCF is used simply as a tracking filter then the harmonic content of the output remains fixed for the duration of each note. However, dynamic variation of harmonic content during a note is also possible by controlling the VCF from the envelope shaper.
For example, to provide a good imitation of a trombone sound the note should initially start off with only a weak harmonic content. As the loudness of the note builds up the harmonic content also increases, i.e. the note becomes 'brighter'. Similarly, at the end of the note it is the harmonics which die away first.
This is achieved by using the VCF in the lowpass mode as a tracking filter with ADSR control, i.e. with inputs from KOV and from the envelope shaper. When a key is depressed the turnover point is initially determined by the KOV input, and is set so that the harmonics are filtered out. As the envelope shaper output voltage rises (attack) the turnover frequency of the VCF is increased to pass more of the harmonic content. At the end of the note (decay) the envelope shaper output falls and the turnover frequency of the VCF is reduced to filter out the harmonics once more.
These two simple examples relate to the imitative capability of the synthesiser, since most people will have a 'feel' for the sound of conventional musical instruments. However, it must once again be stressed that the synthesiser is not limited merely to an imitative role. It can also produce sounds that are unique to itself, that do not occur naturally and are totally 'electronic'.

Highpass filter

So far only the use of the lowpass filter has been discussed. The highpass filter has the opposite effect to the lowpass filter, i.e. it can be used to attenuate the fundamentals of notes while retaining the harmonics. This is obviously useful for sounds which have only a weakly developed fundamental or a bright tonal character, such as harpsichord and spinet type sounds, and certain string and brass instruments. When controlled by the envelope shaper the highpass filter can also give an 'ethereal' character to a sound.

Bandpass filter

In addition to the fundamental and harmonic series produced when a particular note of the instrument is sounded, brass and many woodwind instruments exhibit a number of fixed bandpass resonances, which are determined by the particular mechanical construction of the instrument. Use of the VCF as a bandpass filter with fixed centre frequency (KOV input switched off), together with a second VCF as lowpass tracking filter, allows these instruments to be more accurately imitated.

Pedal controlled Wa-Wa

Using the VCF in the bandpass mode with a fairly high Q-factor, A Wa-Wa effect can be obtained by controlling the VCF with a 0 to 5 V DC supply from a pedal-controlled potentiometer (such Wa-Wa pedals are available commercially or are easily home-made).

Notch filter

By sweeping the centre frequency of the notch filter up and down the spectrum, either manually using a potentiometer or automatically using a low-frequency oscillator, phaser-type sounds can be produced. If this is done using a white noise input instead of a VCO then interesting 'jet-aircraft' noises can be obtained.

Design of the VCF

As far back as 1965, R.A. Moog designed 24 dB/octave lowpass and highpass filters, and no satisfactory alternative to these was found for several years, although they were periodically 're-invented' by others. It was not until the introduction of a specific type of integrated circuit, the operational transconductance amplifier (OTA), that a viable alternative became possible.
The Formant VCF is developed from the two-integrator loop shown in figure 1. Although a complete mathematical analysis of this circuit is beyond the scope of this book (those interested are referred to the bibliography), the basic concept is fairly simple to grasp.
The two-integrator loop can be considered as an analogue computer for the solution of a second-order differential equation. If the input resistor R1 and potentiometer P_Q are removed, it can be seen that the circuit bears a remarkable resemblance to a quadrature oscillator. In fact, if the loop gain of the circuit is sufficient then it will function as an oscillator – at the frequency for which the differential equation solution holds.
P_Q provides damping so that the circuit does not oscillate, but merely acts as a filter. Highpass, bandpass, and lowpass filter functions are available simultaneously at outputs (1), (2) and (3) respectively. At the turnover or centre frequency of the filters there is 90° phase shift between the integrator inputs and outputs. Thus between point (1) and point (3) there is 180° phase shift in all. By combining outputs (1) and (3) using a voltage follower A4 a notch function can be obtained. Since the two inputs are 180° out of phase at the centre frequency there is a null at the junction of the voltage follower's two input resistors at this frequency.
Of course the centre/turnover of this filter is not voltage-controlled, but is fixed by the integrator constants R and C, so to achieve voltage control one of these elements must itself be voltage-controlled. Voltage control of capacitance is impractical in this application. Voltage controlled resistors are possible in the form of LED/LDR combinations or FETs, but unfortunately both these methods suffer from disadvantages such as unpredictable performance due to wide tolerances, small control range, poor linearity, and breakthrough of the control signal.
An alternative solution can be found by re-thinking the basic integrator design. The classic op-amp integrator consists

Figure 1. The two-integrator loop used in the Formant VCF provides 12dB/octave highpass, bandpass and with the addition of A4, a notch filter.

Figure 2. Instead of normal op-amps, OTAs are used in the Formant VCF. The output current change is g_m times the input voltage change, but g_m can be varied by feeding in a control current I_{ABC}.

Figure 3. The OTA integrator used in the Formant VCF. The integrator time constant is controlled by the current I_{ABC}. A high impedance buffer ensures that all the output current of the OTA flows into the integrator capacitor.

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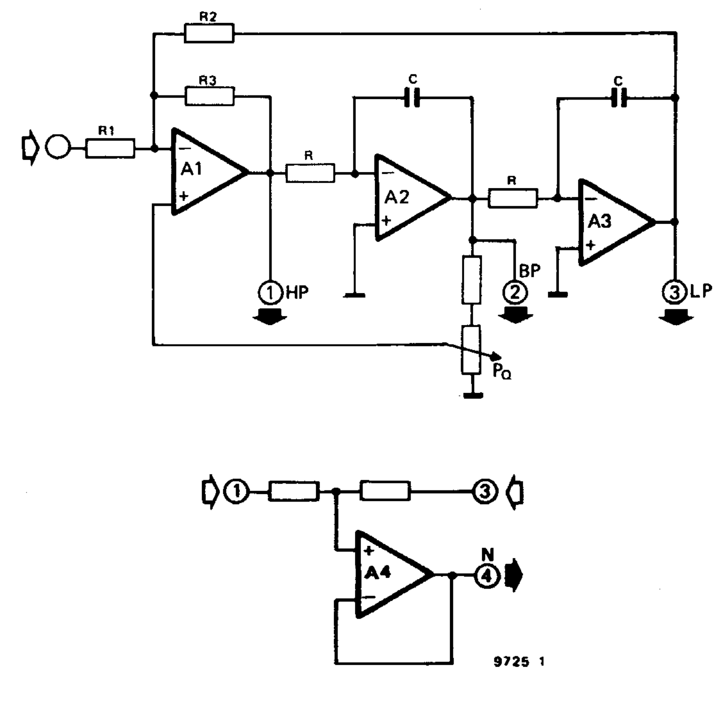

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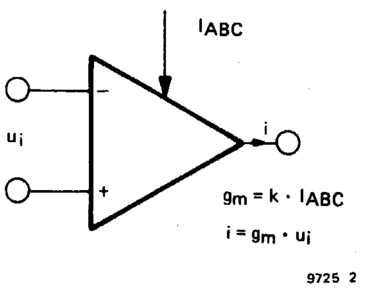

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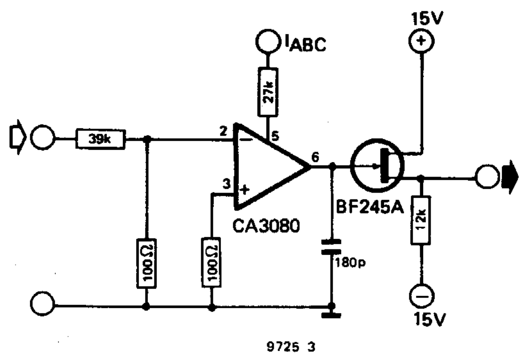

of a differential-input voltage amplifier with the non-inverting input grounded. An input resistor connected to the inverting input (which is a virtual earth point) converts the input voltage into a proportional current. Since this current cannot flow into the inverting input it must flow into the feedback capacitor, and a voltage appears across the capacitor (and hence at the op-amp output).

It is fairly obvious that the op-amp is functioning simply as a voltage-to-current converter, and an equivalent circuit for an integrator would be an amplifier with a voltage-controlled current output, with a capacitor connected, not in a feedback loop, but between the output and ground. Varying the voltage-current transconductance of the amplifier would then effectively vary the 'resistance' constant of the integrator.

A suitable device exists ready-made in the shape of the operational transconductance amplifier or OTA. This is

Hardwired inputs:

KOV	=	Keyboard Output Voltage (from interface receiver).
ENV	=	Envelope shaper control voltage (from ADSR unit).
VCO 1, 2, 3	=	From VCOs 1, 2 and 3.

Front-panel inputs:

ECV	=	External Control Voltage.
TM	=	Tone colour ('Timbre') Modulation input.
ES	=	External Signal, e.g. noise, input.

Outputs:

VCF/IOS	=	Internal Output Signal from VCF, (will be hardwired to a VCA).
EOS	=	External Output Signal from VCF (front panel output).

Front-panel controls:

OCTAVES	=	P1, coarse frequency adjustment.
ENV	=	P2, sets envelope shaper control voltage.
TM	=	P3, sets tone colour modulation level.
ES	=	P4, sets external signal level.
Q	=	P5, Q-factor adjustment.
OUT	=	P6, sets VCF/IOS output level (not EOS!).
ECV/KOV	=	S1, selects external or internal control voltage input.
HP	=	S2, selects high-pass output.
BP	=	S3, selects bandpass output.
LP	=	S4, selects low-pass output.
N	=	S2 + S4, selects notch (band-stop) output.

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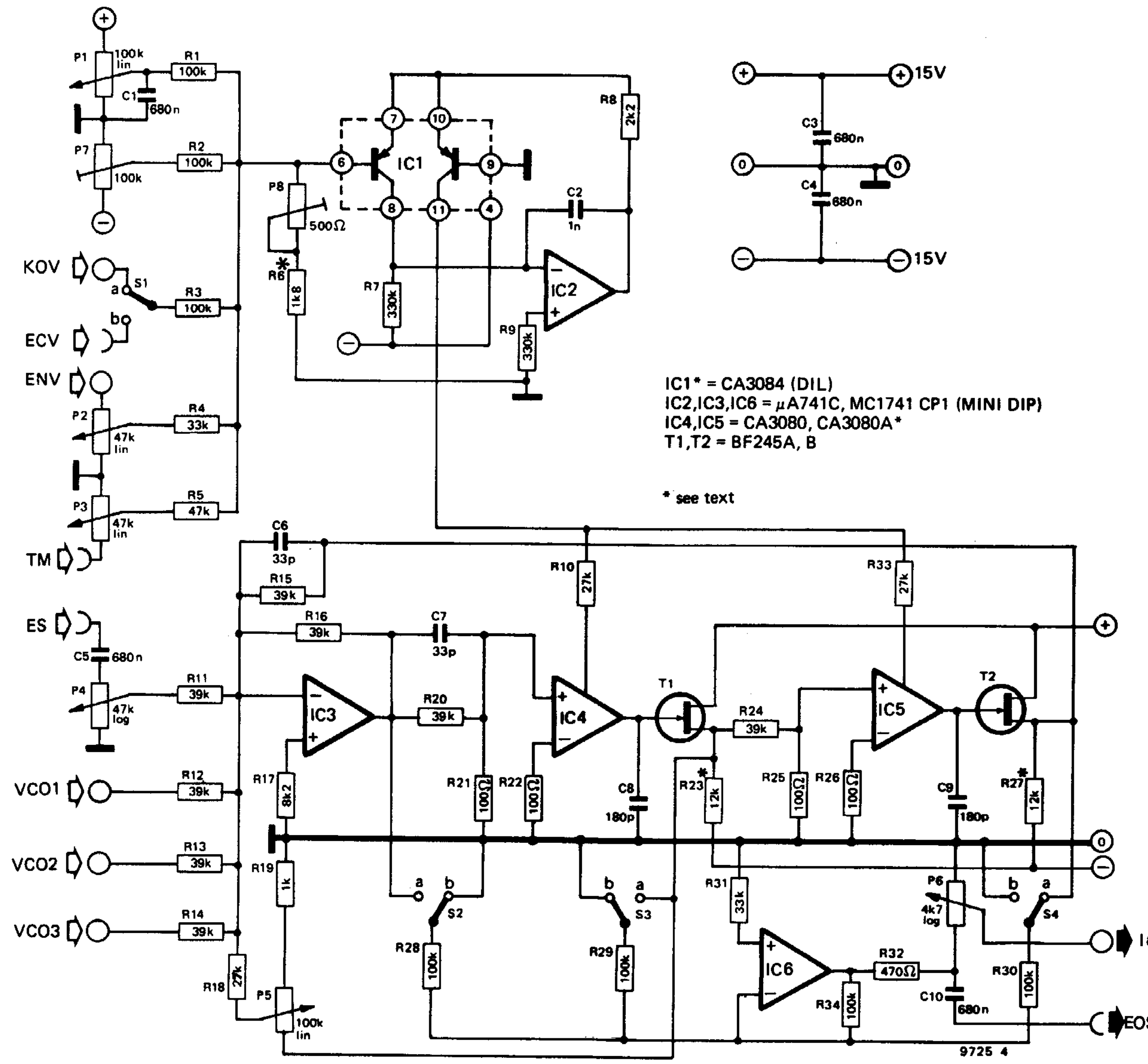


an amplifier that produces an output current which is proprotional to the input voltage, i.e. $i = g_m \cdot u_i$, where i is the output current, u_i is the input voltage and g_m is the transconductance. The feature of the OTA which makes it ideal for the VCF is that the transconductance g_m is determined by a control current I_{ABC}, thus $g_m = k \cdot I_{ABC}$, where k is a constant. This is illustrated in figure 2.

For the CA3080 OTA used in the Formant VCF the constant k is $19.2\ V^{-1}$ at an ambient temperature of 25° C, and so $g_m = 19.2 \times I_{ABC}$ mS (milliSiemens = milliamps/volt). This IC is particularly suitable because of the outstanding linearity of its transconductance characteristic over three decades of control current, and because of its relatively small tolerance in the value of 'k' (2:1 for the 3080 and 1.6:1 for the 3080A). However good linearity is achieved only for small input signals, and the input voltage must be attenuated to about ± 10 mV when used in the VCF.

Figure 3 shows the circuit of the integrator used in the Formant VCF. The input voltage is attenuated by the potential divider connected to the inverting input, and across the output is connected the 180 p integrating capacitor.

To maintain correct operation of the integrator the total output current of the OTA must flow into the integrator capacitor, which means that a buffer stage with a very high input impedance is required on the OTA output to avoid 'current-robbing'. A FET connected as a source-follower is used for this purpose.

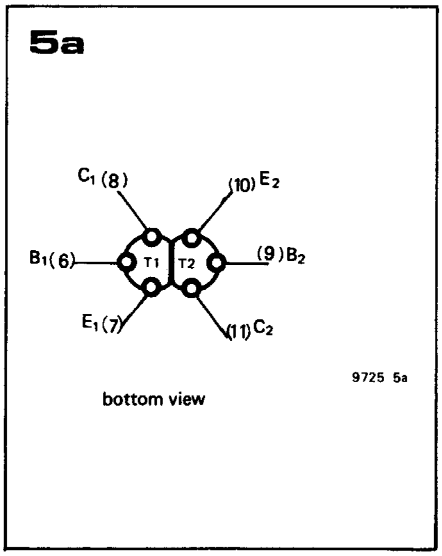
The control current I_{ABC} is fed in through a 27 k resistor. The integrator time constant is inversely proprotional to the control current, so the VCF centre/turnover frequency is directly proportional to the control current.

Complete circuit of the VCF

Figure 4 shows the complete circuit of the VCF. The actual filter circuit has a linear frequency characteristic and is current controlled. It must therefore be preceded by an exponential converter that converts the input control *voltage* into an exponentially related control *current*, so that the VCF tracks with the same 1 octave/V characteristic as the VCOs.

The exponential converter occupies the upper portion of the circuit, and is essentially similar to that of the VCOs. However, the control characteristic of the VCF does not need to be so accurate as that of the VCO, since a small error will only introduce minor, unnoticeable errors in amplitude response, whereas the same error in the VCO characteristic would cause unacceptable tuning errors.

For this reason the VCF exponential converter is provided only with a passive input adder (cf. figure 2a of the last chapter), and temperature stabilisation of the exponentiator is dispensed with, thus saving the cost of a not in-



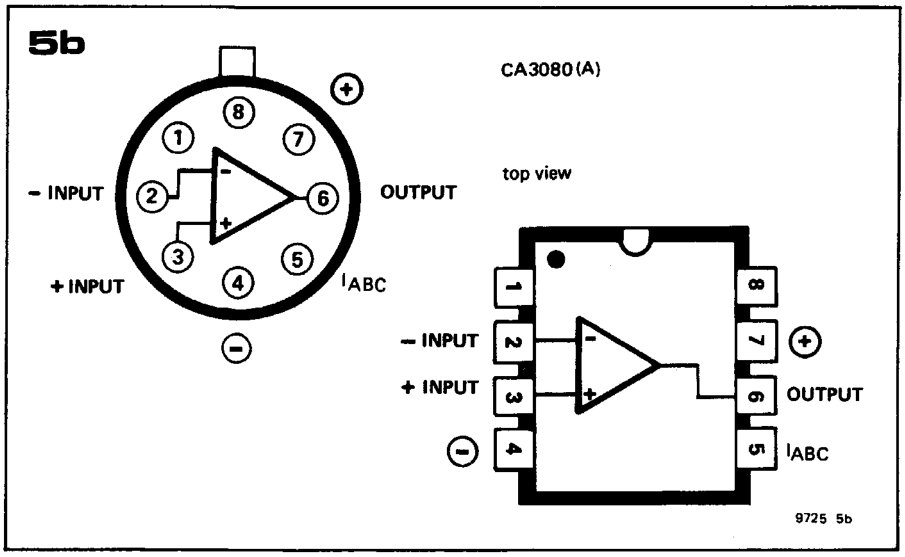


Figure 4. Complete circuit of the Formant VCF, which consists of a voltage-current exponential converter and a linear current-controlled filter.

Figure 5a. Two well-matched PNP transistors may be used in place of IC1 for greater economy. The pin numbers shown correspond to the pinout of IC1.

Figure 5b. The CA 3080 is available in two packages. If the TO- package is used the leads must be bent to fit the DIP layout on the p.c.b.

expensive μA726 IC. However, temperature compensation is retained in the form of a matched transistor pair. The circuit differs here from the VCO since the exponentiator must source current into the OTAs rather than sinking it as in the VCO, so PNP transistors are used.
Since temperature stabilisation is not used, a number of options are open for the choice of the matched transistor pair. Those who have access to a good transistor tester or curve tracer can select a matched pair of any small signal medium gain ('B' spec) transistors such as the BC 179B, BC 159B, BC 557B etc. These are then glued together with epoxy adhesive for good thermal tracking as shown in figure 5a, taking care that there is no electrical contact between the cases if metal-can types are used. (Note that the pin numbers given in figure 5a correspond to the IC pinning in figure 4).
The preferred solution is to use a CA 3084 transistor array, which is what was used in the prototype, but if this is difficult to obtain then almost any dual PNP transistor, such as the Analog Devices AD 820 . . . AD 822, Motorola 2N3808 . . . 2N3811 or SGS-ATES BFX 11, BFX 36, will do.
Note that the value shown for R6 (1k8) is correct when using the CA 3084. If a dual transistor is used, it is advisable to reduce the value of R6 to 1k5.
The current-controlled filter consists of IC3, IC4 and IC5. It will be noted that the integrators IC4 and IC5 are non-inverting. This does not affect the operation of the circuit, since non-inversion has the same effect as the double inversion that takes place in figure 1. However, it does ensure that the three outputs of the filter are in the same sense, whereas in figure 1 the bandpass output is inverted with respect to the other two outputs.
IC6 functions as an output buffer, and also as a summing amplifier for the highpass outputs to provide the notch function. By setting S2, S3 or S4 in position 'a', highpass, lowpass or bandpass functions respectively may be selected. By setting both S2 and S4 in position 'a' the notch function is obtained. Since IC3 is connected as an inverting amplifier and IC6 also inverts, this double inversion means that the output signal is non-inverted with respect to the input signals. The overall gain of the VCF (in the passband) is x 1 (0dB).

Inputs, controls and outputs

The exponential converter section is equipped with a coarse octave tuning control P1 (note the absence of a fine control as compared with the VCO) and two presets P7 and P8 to adjust the offset and octave/V characteristic.
KOV and ECV control inputs are provided, as for the VCO. The input for envelope shaper control (ENV) is adjustable by means of P2. The tone colour modulation input controlled by P3/(TM) is analogous to the FM input of the VCO, i.e. it allows the centre/turnover frequency of the VCF to be modulated. There are four signal inputs, three internally-wired VCO inputs and one external signal (ES) input, whose amplitude can be controlled by P4. The Q-factor of the filter is controlled by P5.
Switches S2 to S4 select the desired filter type, as has already been described. Two outputs are provided, an uncontrolled output EOS which is brought out to a front-panel socket, and an internal output IOS, which is controlled by P6.

Construction

A printed circuit board and component layout for the VCF are given in figure 6. The same considerations of component quality apply to the VCF that apply to all parts of the synthesiser. As mentioned earlier, two basic versions of the CA 3080 are available. The CA 3080A has better specifications as regards tolerance, and extended temperature range, but the basic CA 3080 is quite adequate (assuming that the synthesiser is not to be used in Antarctic blizzards).
The CA 3080 is available in two packages, TO- can and mini-DIP, both of which are shown in figure 5b. The p.c. board is laid out for the mini-DIP version, but the TO- version can easily be accomodated by splaying out the leads to conform with the mini-DIP pinning (in fact some TO- package 3080s are supplied with this already done).
The FETs T1 and T2 must be tested as detailed in chapter 3 and their source resistors R23 and R27 selected in accordance with Table 1 of that chapter.
A front panel layout for the VCF is given in figure 7, and a wiring diagram for the front-panel mounted components is shown in figure 8.

Testing and adjustment

During assembly, it is convenient to use IC sockets so that the current-controlled fiter section of the circuit can be tested independently of the exponential converter. To test the CCF, IC1 is removed and a 100k log potentiometer is connected 'back-to-front' between ground and -15V (i.e. so that the end of the track approached by clockwise rotation of the wiper is connected to ground).
A multimeter set to the 100 μA range is

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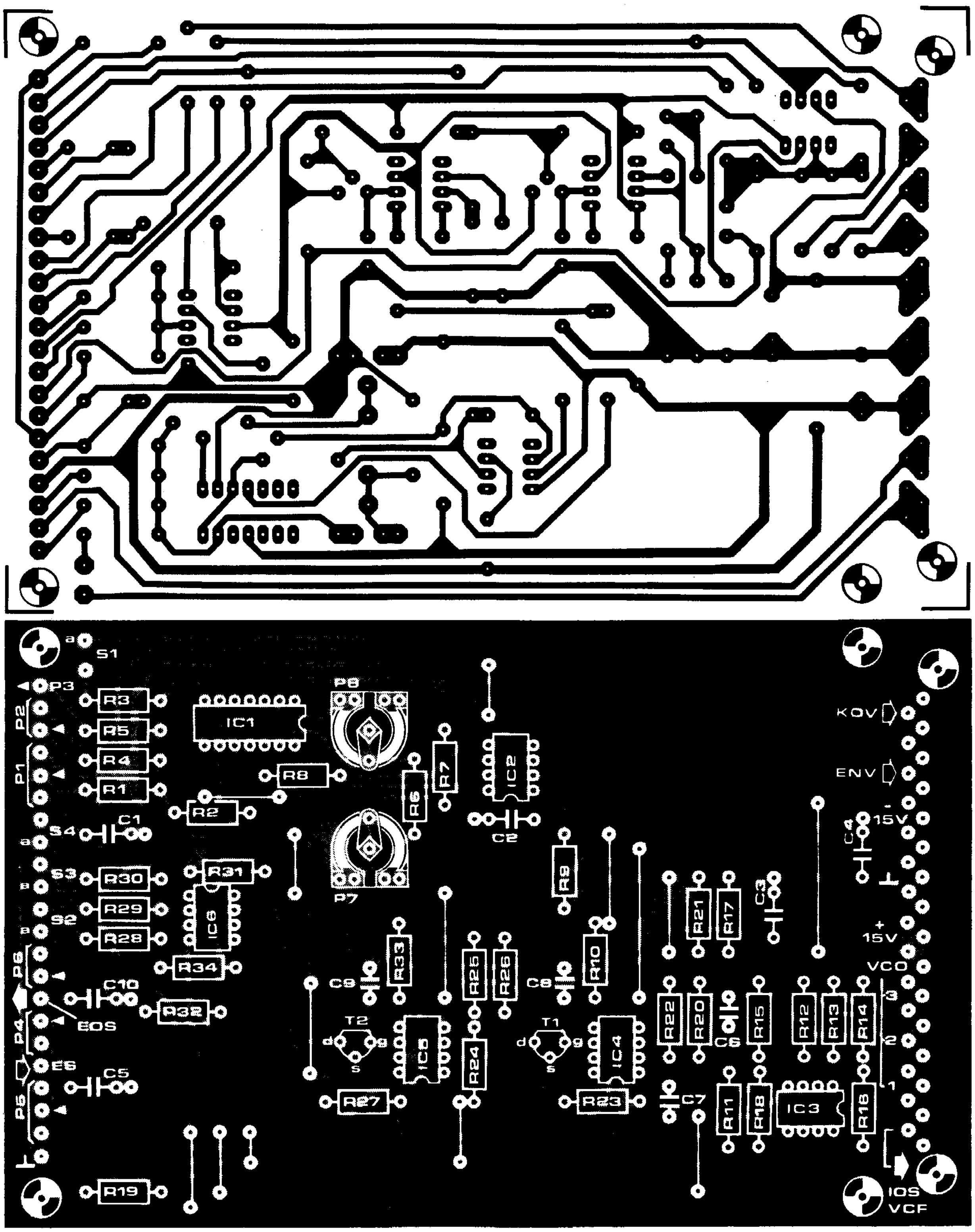

Parts List

Resistors:
R1, R2, R28, R29,
R30, R34 = 100 k
R3 = 100 k (1% metal oxide)
R4 = 33 k
R5 = 47 k
R6 = 1k8 (see text)
R7, R9 = 330 k
R8 = 2k2
R10, R33 = 27 k
R11, R12, R13, R14,
R15, R16, R20, R24 = 39 k
R17 = 8k2
R18 = 22 k
R19 = 1 k
R21, R22, R25, R26 = 100 Ω
R23, R27 = 12 k (nominal value, see text)
R31 = 33 k
R32 = 470 Ω

Potentiometers:
P1, P5 = 100 k lin
P2, P3 = 47 k (50 k) lin
P4 = 47 k (50 k) log
P6 = 4k7 (5 k) log

Presets:
P7 = 100 k
P8 = 470 Ω (500 Ω)

Capacitors:
C1, C3, C4, C5, C10 = 680 n
C2 = 1 n
C6, C7 = 33 p
C8, C9 = 180 p

Semiconductors:
IC1 = CA 3084 (DIL) see text.
IC2, IC3, IC6 = μA 741 C (Mini DIP), MC1741 CP1 (Mini DIP).
IC4, IC5 = CA 3080 (A)
T1, T2 = BF 245a, b.

Miscellaneous:
31-way plug (DIN 41617)
S1 – S4 = miniature SPDT toggle switch

Figure 6. Printed circuit board and component layout for the VCF. (EPS 9724-1).

connected between the wiper of the potentiometer and the junction of R10 and R33, an input signal is provided to the VCF from a sinewave generator or from the VCO, and the Bandpass output is monitored on an oscilloscope. The test then proceeds as follows:
1. Set the Q-factor of the filter to maximum (wiper of P5 turned towards R19).
2. By means of the 100k log potentiometer set the control current to 50 μA on the meter.
3. Slowly increase the generator frequency from about 300 Hz to 1500 Hz; somewhere in this range the VCF output should peak as its resonant frequency is reached (i.e. there will be a sharp increase in output at a particular frequency with a fall-off on each side). Note the frequency at which resonance occurs.
4. Increase the control current to 100 μA and check that resonance now occurs at twice the previously noted frequency.
Note. Tests 2 to 4 are intended to check the linearity of the filter frequency v. control current characteristic. The tolerance in the absolute value of filter frequency for a given control current is due to OTA tolerances and is unimportant provided linearity is maintained i.e. the filter frequency doubles for each doubling of control current.
5. Set the generator to about 50 Hz and check that it is possible to obtain resonance at this frequency by varying the control current with the 100 k potentiometer. Repeat this test at 15 kHz.
The exponential converter can now be tested after inserting IC1 and removing IC4 and IC5. A multimeter set to the 100 μA range is connected from the bottom end of R10 to -15V and the wiper voltage of P1 is monitored with a voltmeter.
The test and adjustment now proceed as follows:
1. Set P8 to its mid-position, and turn P1 fully anticlockwise so that its wiper voltage is zero. Adjust P7 until the mircroammeter reading is 50 μA.
2. Turn P1 clockwise until its wiper voltage is 1V, then adjust P8 until the microammeter reads 100 μA.
3. Repeat the procedure for 2V, 3V, 4V etc. on the wiper of P1, checking that the exponentiator output current doubles for every 1V increase.

Offset adjustment

Now that the two sections of the VCF have been checked, IC4 and IC5 can be re-inserted so that the entire VCF can be checked as a functional unit, as follows:
1. A squarewave with 50% duty-cycle at a frequency of about 500 Hz is fed to one of the filter inputs. P1 is turned fully clockwise and P7 is turned anticlockwise.
2. The lowpass output of the VCF is

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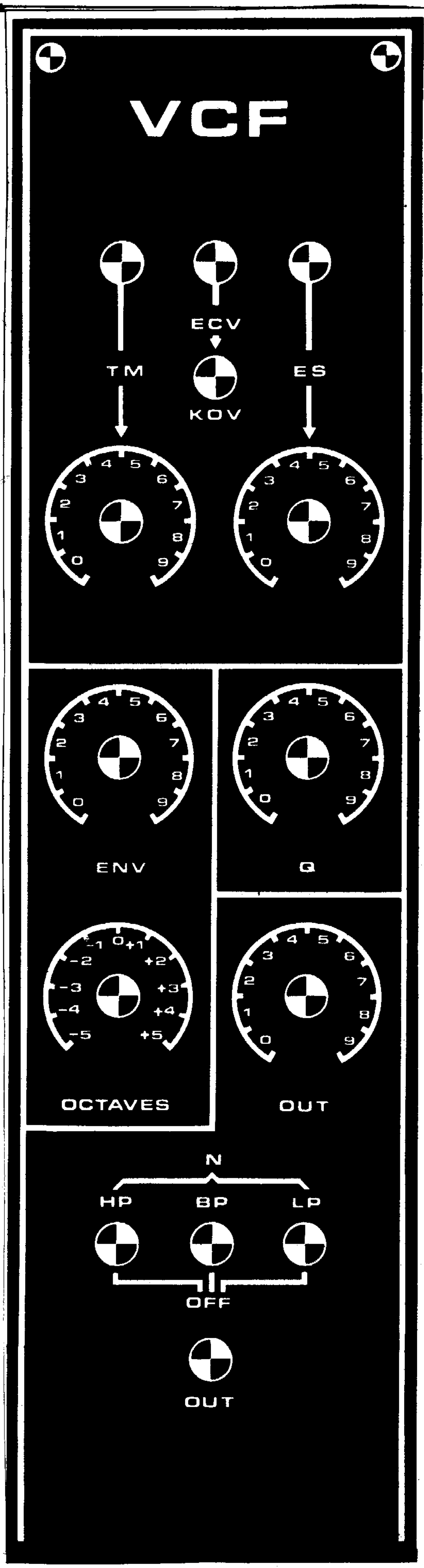

Figure 7. Front panel layout for the VCF.

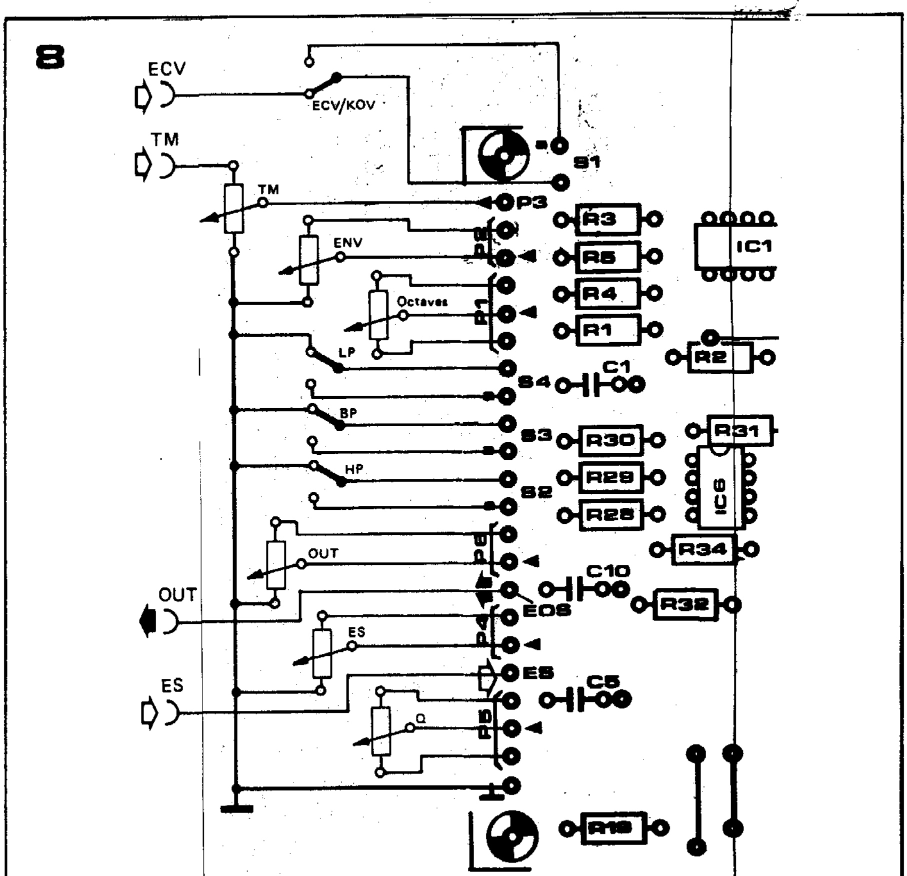

Figure 8. Wiring diagram for the panel mounted components.

monitored on an oscilloscope, and at this stage should appear at the output without degradation.
3. If the wiper of P7 is now turned slowly clockwise the leading edge of the squarewave will start to be rounded off as the turnover point of the filter is reduced. To carry out the offset adjustment with P7 its wiper is turned as far clockwise as is possible without significantly degrading the square waveform (just a slight rounding of the top corner is acceptable, but this adjustment does not have to be particularly precise).

Octaves/Volt adjustment

The octave/V characteristic of the VCF can be adjusted by seeing how well it tracks against a previously calibrated VCO. To do this, the KOV input is connected to the VCO and the VCF, and the sine output of the VCO is connected to the VCF input. The adjustment procedure is as follows:
1. Switch off the main tuning of the keyboard, depress top C of the keyboard and use the octaves control of the VCO to set its frequency to about 500 Hz.
2. Set the Q control, P5, of the VCF to maximum, monitor the bandpass output of the VCF and adjust P1 until the VCF output peaks. As the filter is easily over-loaded at high Q-factors it may be necessary to reduce the VCO output voltage.
3. Depress the key two octaves lower and adjust P8 until the VCF output again peaks.
4. Depress top C again and if necessary readjust P1 so that the output peaks.
5. Repeat 3 and 4 until no further readjustment is necessary for the output to peak when changing from one note to the other.
6. The offset adjustment may have been disturbed, so check this and if necessary readjust P7 as described in the offset adjustment procedure.
7. Repeat 3 onwards until no further improvement can be obtained.

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chapter 7

24 dB VCF

Because of the greater range of tonal possibilities they offer, VCFs with an extremely steep slope seem to have a particular appeal for most synthesiser enthusiasts. The design presented here is for a VCF offering a choice of lowpass or highpass functions and a filter slope of 6, 12, 18 or 24 dB per octave.

New possibilities

It should be stated at the outset that the 24 dB VCF is not intended to replace the 12 dB design. 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

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.

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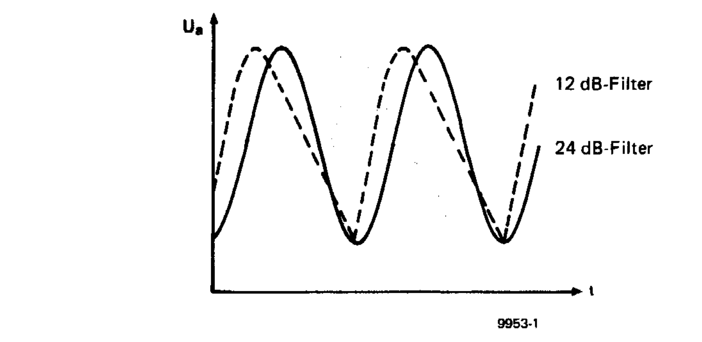

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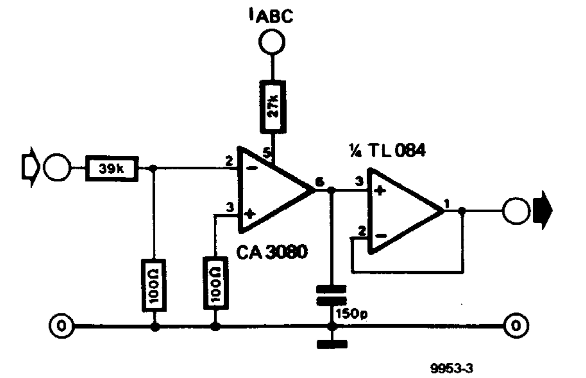

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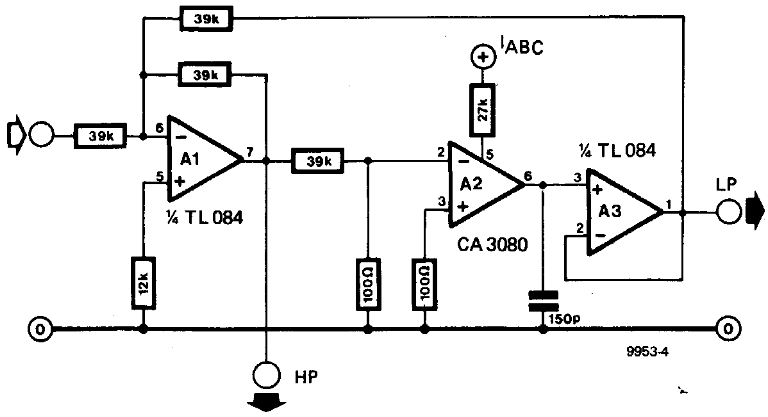

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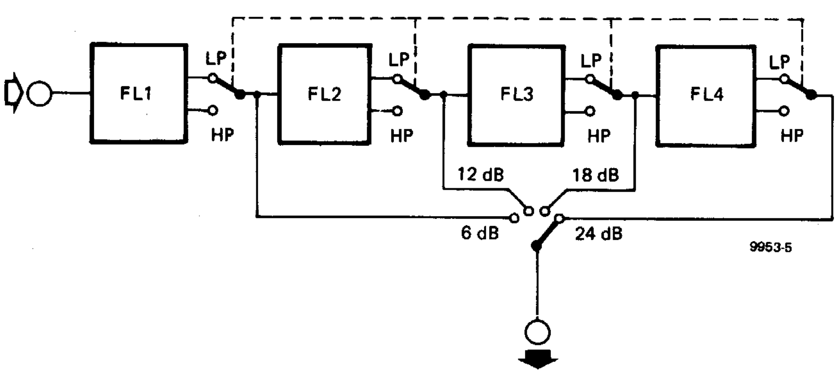


VCF (continuous line) when fed with a sawtooth waveform. It is apparent that, due to the almost complete removal of the harmonics of the sawtooth, the output of the 24 dB VCF is practically a sinewave, whereas the original waveform is still apparent at the output of the 12 dB VCF since the harmonics 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.

Design of the 24 dB VCF

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 in the previous chapter. However, advantage has been taken of recent developments in FET op-amp technology to simplify the design slightly. As has been explained, 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

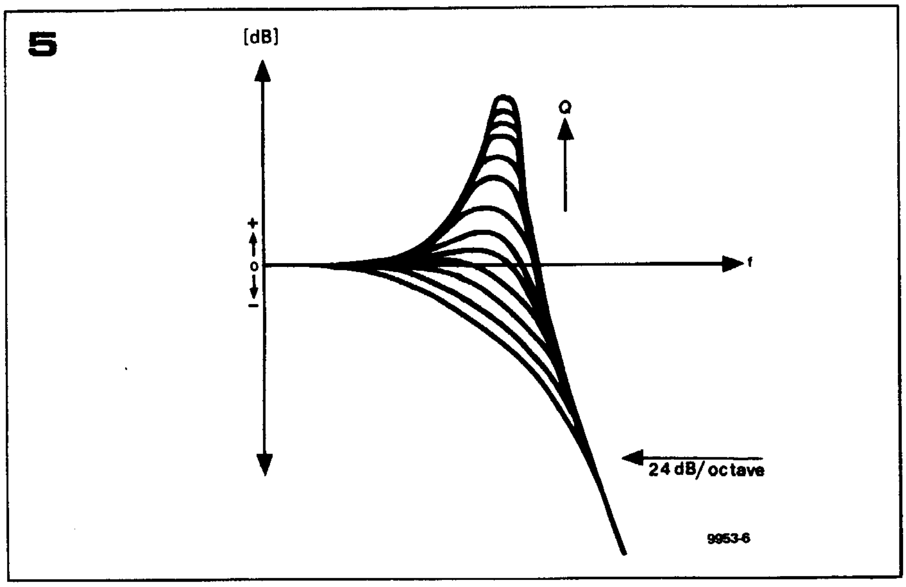
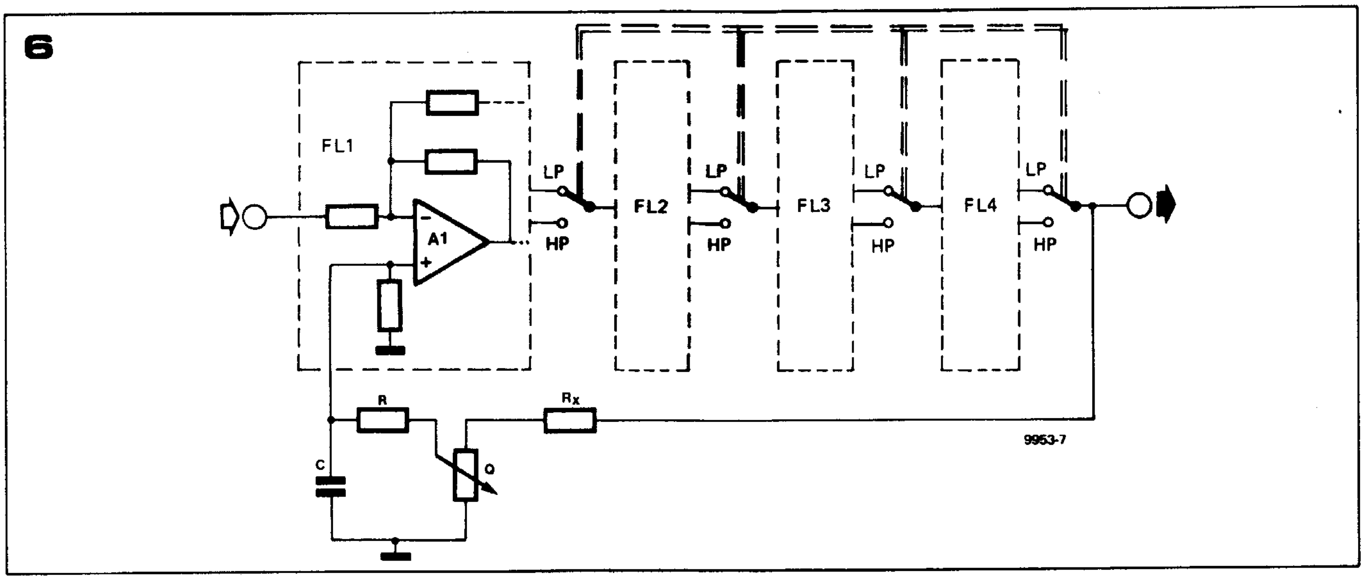


Figure 5. Positive feedback around the entire filter allows the response to be boosted about the turnover frequency. The degree of boost can be varied by a 'Q' control.

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.



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 available. The use of one of these ICs simplifies the design and obviates the need to select FETs, which becomes something of 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

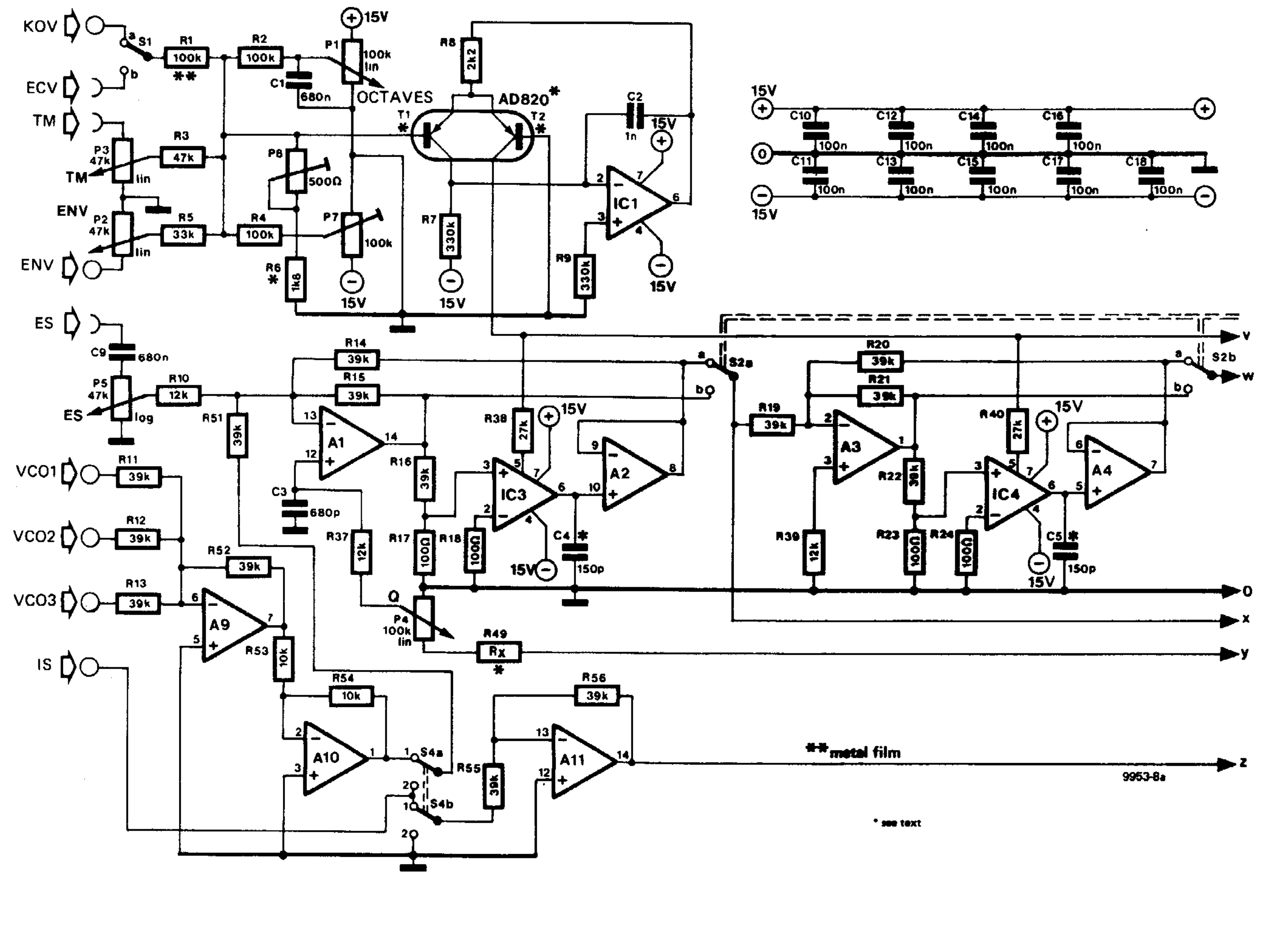
The complete circuit of the 24 dB VCF is given in figure 7. The exponential converter, constructed around T1, T2 and IC1, is identical to that used in the 12 dB VCF and gives the same 1 octave per volt characteristic to the turnover frequency of the filter. The control voltage inputs are also the same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

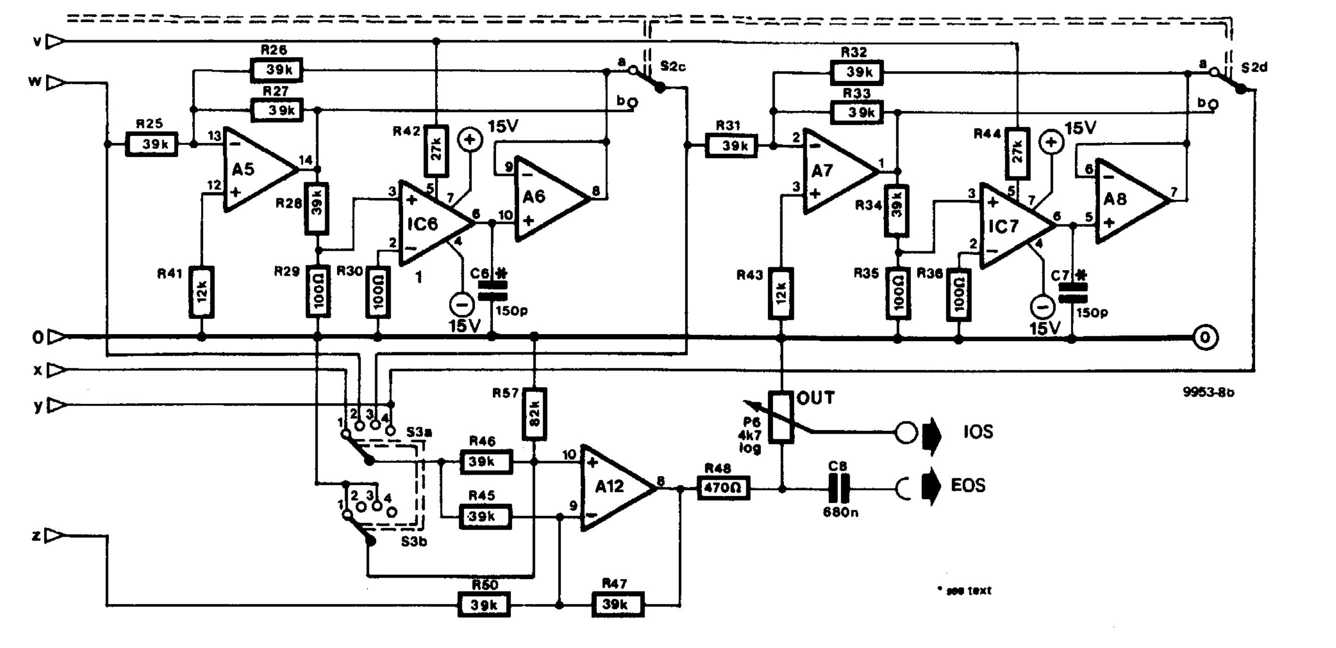
7a

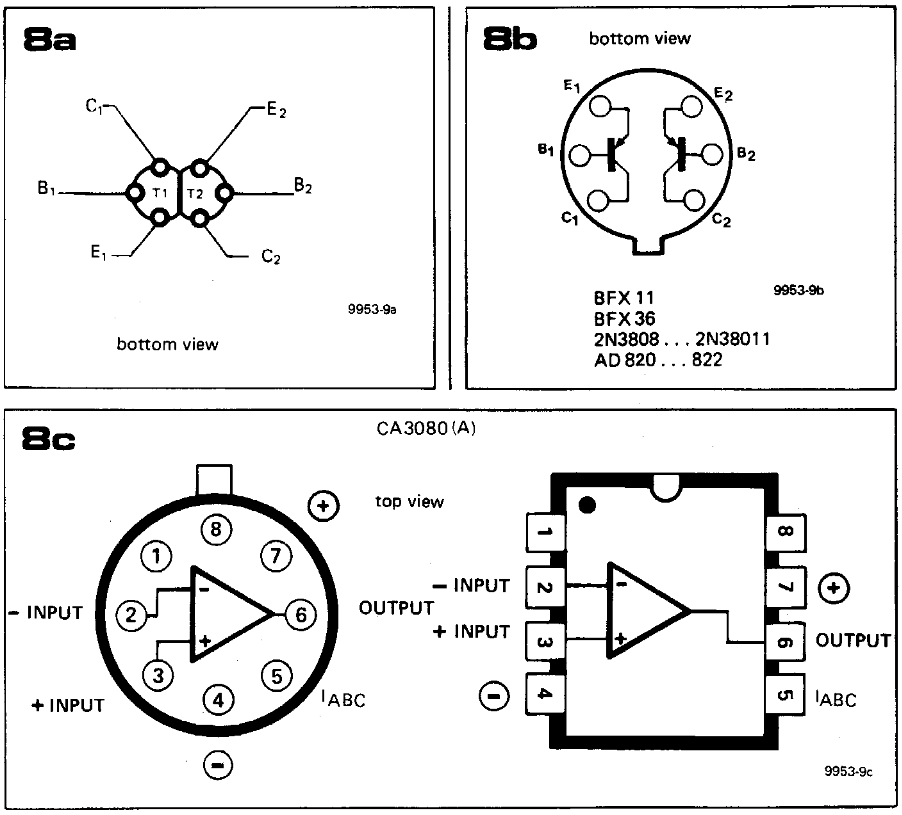


7b

A1 + A2 + A3 + A4 = IC2 = TL084
A5 + A6 + A7 + A8 = IC5 = TL084
A9 + A10 + A11 + A12 = IC8 = TL084

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





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

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.

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 = TL084, TL074
IC8 = TL084, TL074, LM 324
IC3 . . . IC6 = CA 3080, CA3080A (MINIDIP or TO; see text)
T1,T2 = AD820 . . . 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 chapter 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 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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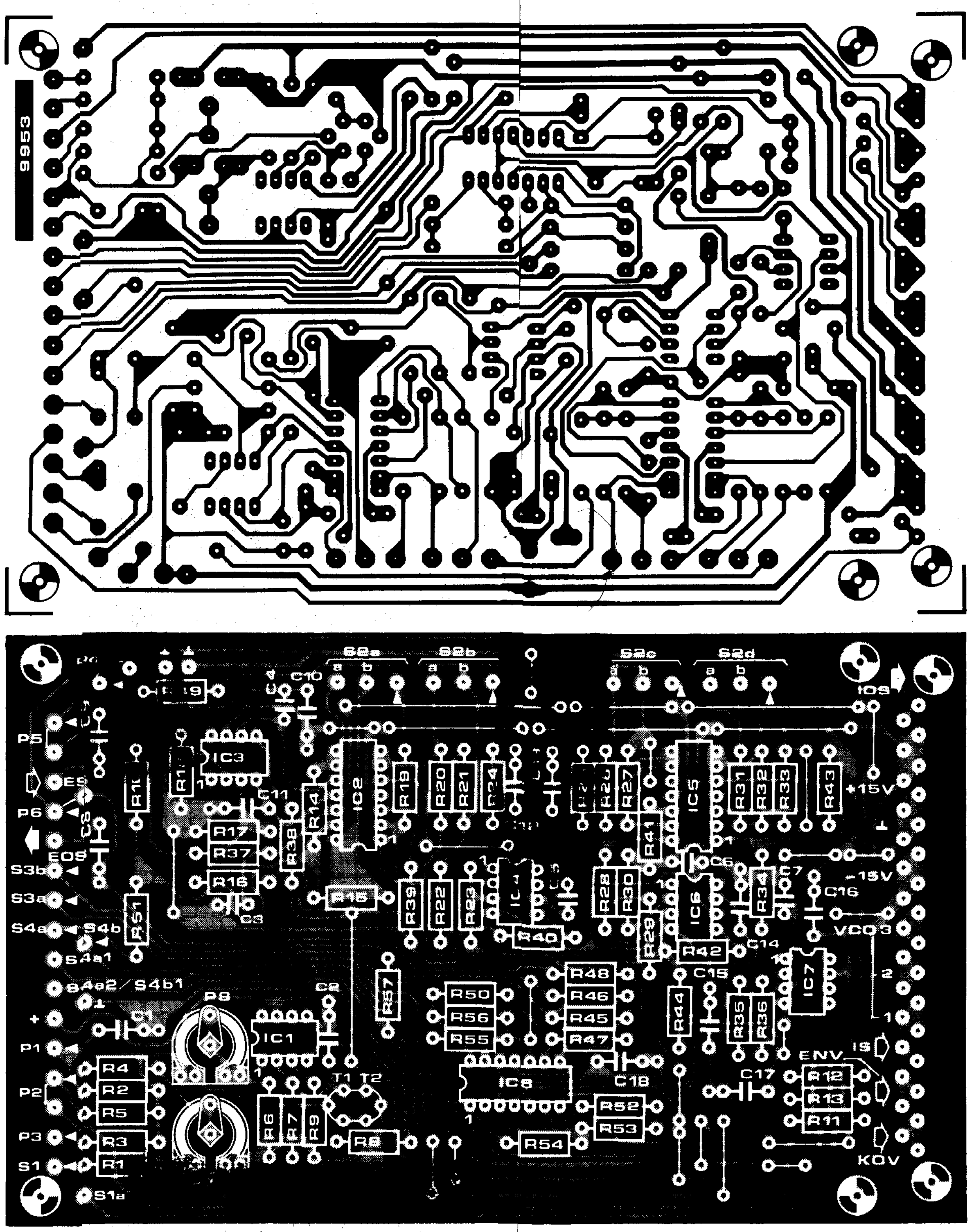

given in figure 9, while a front panel layout is given in figure 10.

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 an 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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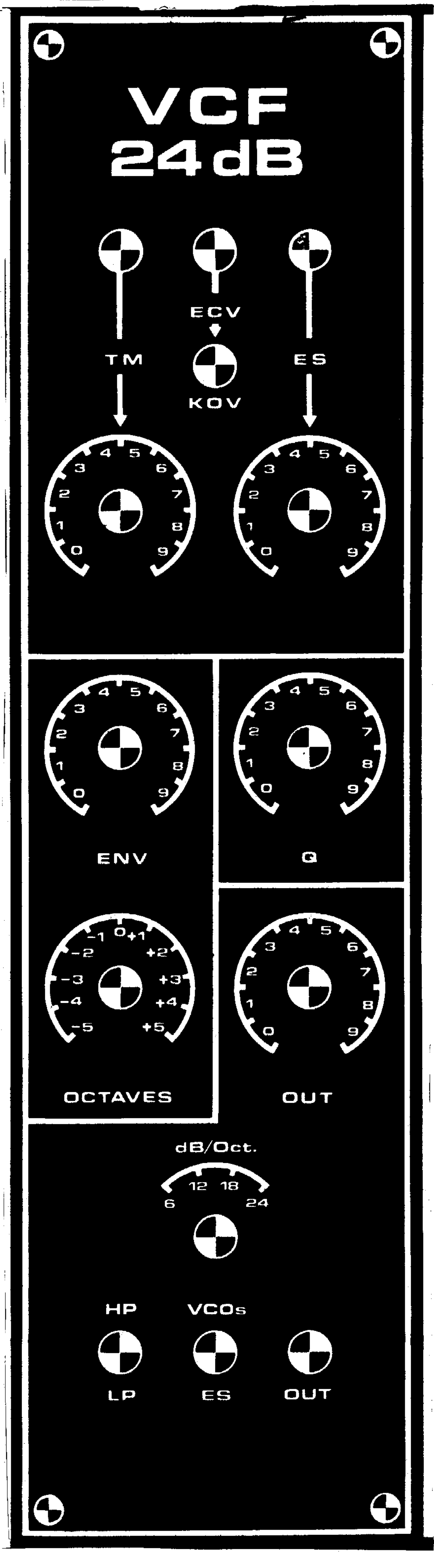

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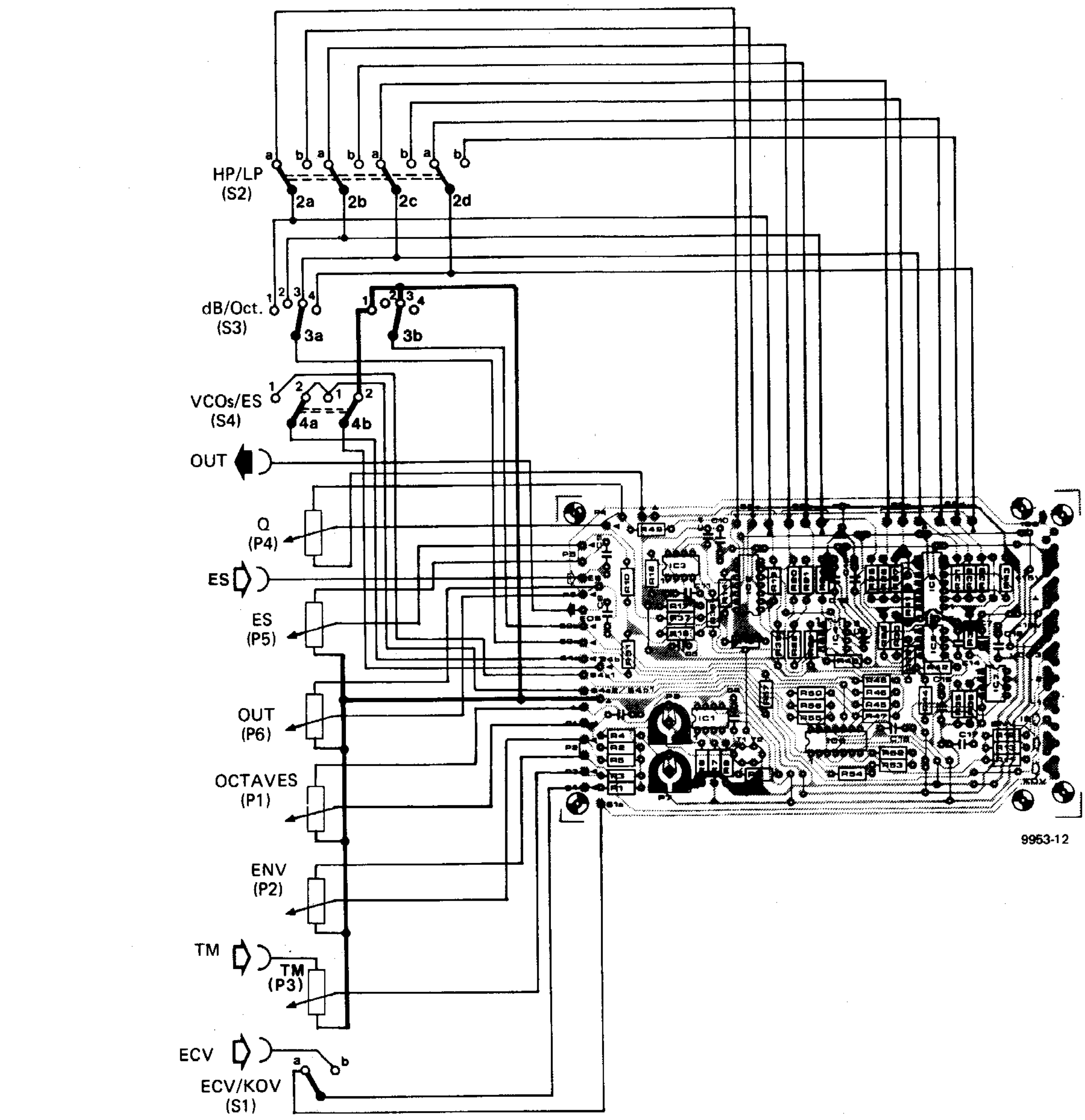

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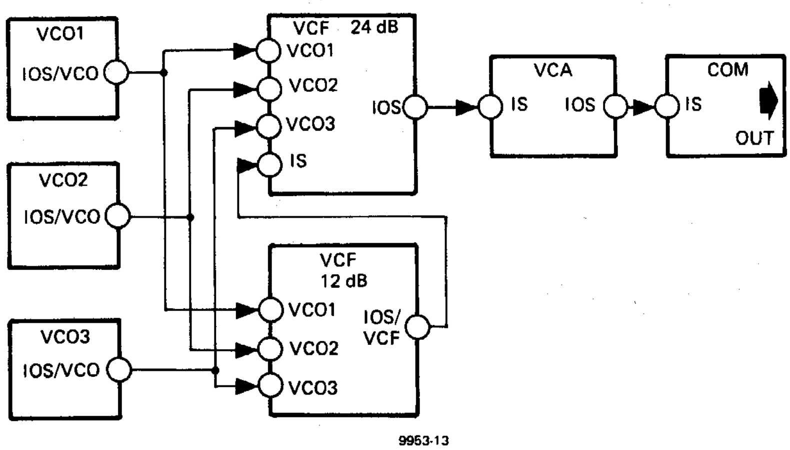

Figure 10. Front panel layout for the VCF. (EPS 9953-2).

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.

procedure given in the previous chapter. The offset and octaves per volt adjustments can also be carried out using the procedure given there. 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

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 goes to the IC 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.

Figure 13. Test circuit for the selection of OTAs.

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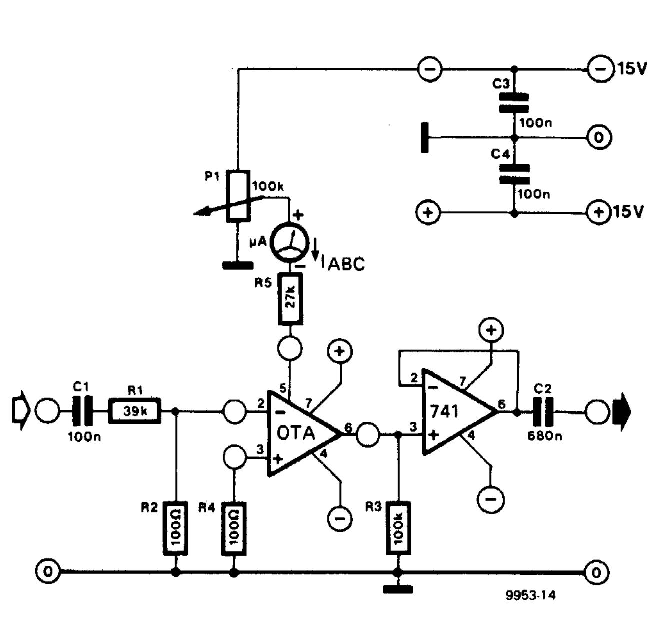

OTA selection procedure

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.

chapter 8

resonance filter module

In addition to an almost limitless variety of non-natural, wholly 'electronic' sounds, the Formant music synthesiser can, of course, be used to imitate the voicing of conventional (mechanical) musical instruments. The filter module described in this article is designed to allow more realistic simulation of natural musical instruments by providing the fixed bandpass resonances which are an important determining factor in the timbre of mechanical tone generators.

Although music synthesisers are capable of producing the most 'wierd and wonderful' electronic effects, it is a fact that they are frequently employed to imitate the sound of traditional acoustic instruments. Many commercially available synthesisers, for example, are provided with preset facilities for various common instrumental voices, whilst special units such as 'string-synthesisers', which are designed solely to reproduce the sound of a string section, are becoming increasingly popular.

As has already been explained, basic factors influencing the characteristics of a muscial note are pitch, dynamic amplitude, and dynamic harmonic content. As the reader will be aware, pitch and dynamic amplitude characteristics are controlled by the VCO (Voltage Controlled Oscillator) and VCA + ADSR (Voltage Controlled Amplifier and Attack - Decay - Sustain - Release) modules in the Formant synthesiser, whilst the VCF (Voltage Controlled Filter) is used to vary the harmonic content of the signal.
However, in the case of mechanical tone generators, for example brass and woodwind instruments, an additional consideration is the existence of resonant areas in the instrument which possess free vibration periods of their own. These resonances, which are known as *formants* (whence the name for the Elektor music synthesiser!) are determined by the shape and mechanical construction of the particular instrument (the wooden back and belly of a violin, the pipes of an organ etc.). Unlike the variable pitch of, say, a violin string, they tend to reinforce the same harmonics, whatever the pitch of the note being played. The nature of the formants in an instrument is in fact one of the factors which govern its quality.
It should thus be apparent that, in order to realistically simulate the tonal characteristics of traditional instruments, one must be able to tailor the static harmonic content of the note accordingly. What is required is a number of resonant filters with independently variable centre frequency, gain and Q-factor. These features are present already in the state variable VCF of the Formant; however, that is only one filter, and more to the point, in this particular application there is no need for the filters to be voltage-controlled, since the filter parameters will be preset to suit whatever musical instrument is being imitated. This explains the reason for the separate manually-controlled resonance filter module described in this article.

Figure 1. The fundamental frequency range of a number of traditional musical instruments, with reference to that of a grand piano. *(From: 'Elektronik Taschenbuch, Band 1', Ferd. Dümmlers Verlag, Bonn; with kind permission from the publishers.)*

The uses of resonance filters

The effect of resonance filters can best be heard on 'bright' sharp VCO waveforms which have a high proportion of fairly intense upper harmonics. The effect on vocal sounds can be illustrated by taking a suitable signal with a frequency of around 200 Hz, setting the Q of the filter to a mid-value, and varying the centre frequency from minimum to maximum. At first 'dark' sounding tones, largely devoid of higher harmonics will be obtained; as the centre frequency is increased however, one by one the various vowel sounds can be distinguished until, at high centre frequencies, reedy flute-like sounds are produced. The higher the Q of the filter, the more pronounced the above effects – and vice versa.
All bandpass resonances of musical importance lie between roughly 100 and 2000 Hz. Table 1 lists the main fixed resonances of a number of common musical instruments and also indicates which VCO waveform is best suited to imitate the instrument in question. This table is, of course, merely intended as a rough guide, in the final instance the decision should rest with one's own ears. Unless otherwise indicated, the Q-control should be set to the mid-position. As a further aid, figure 1 shows the fundamental frequency ranges of vari-

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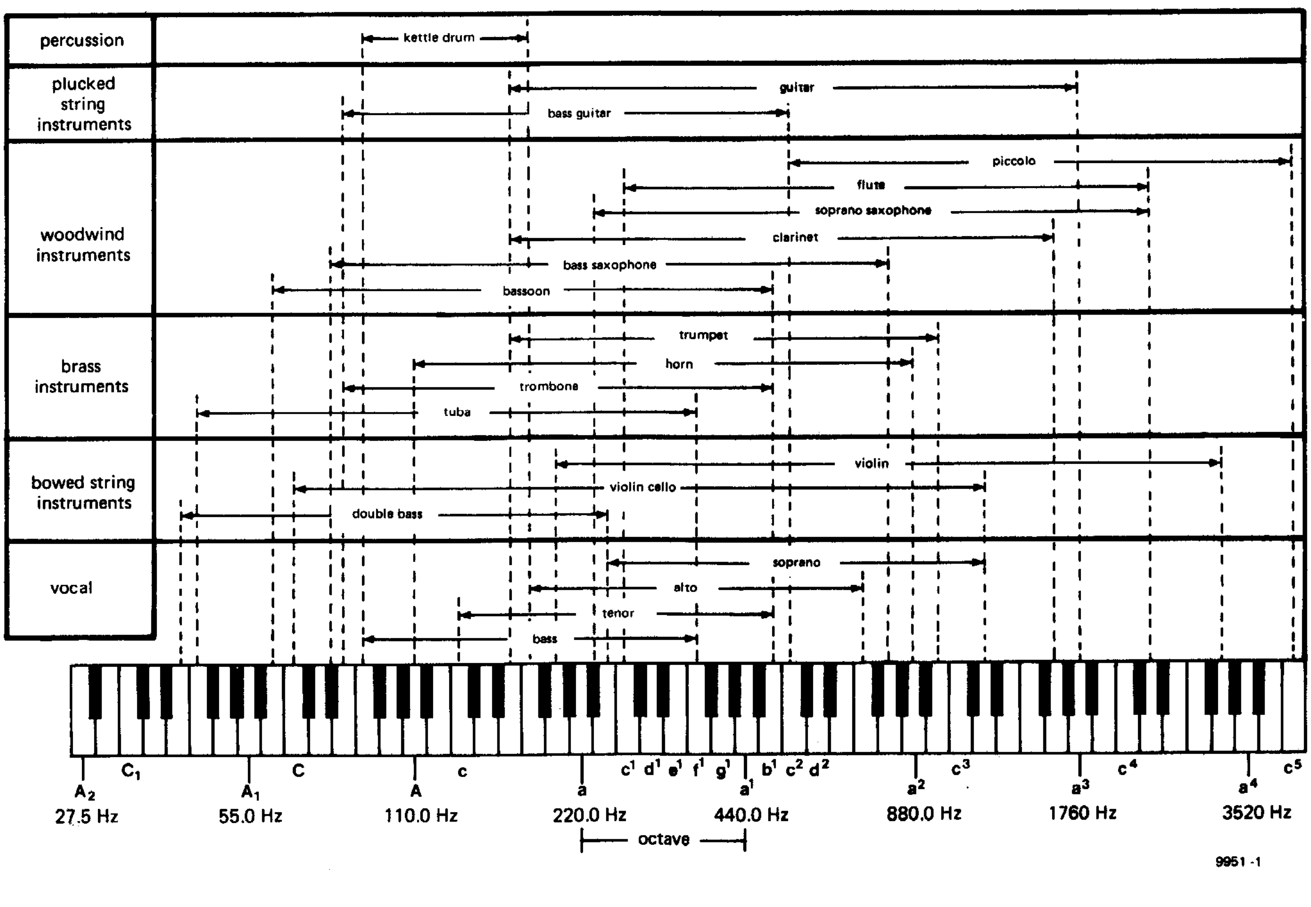
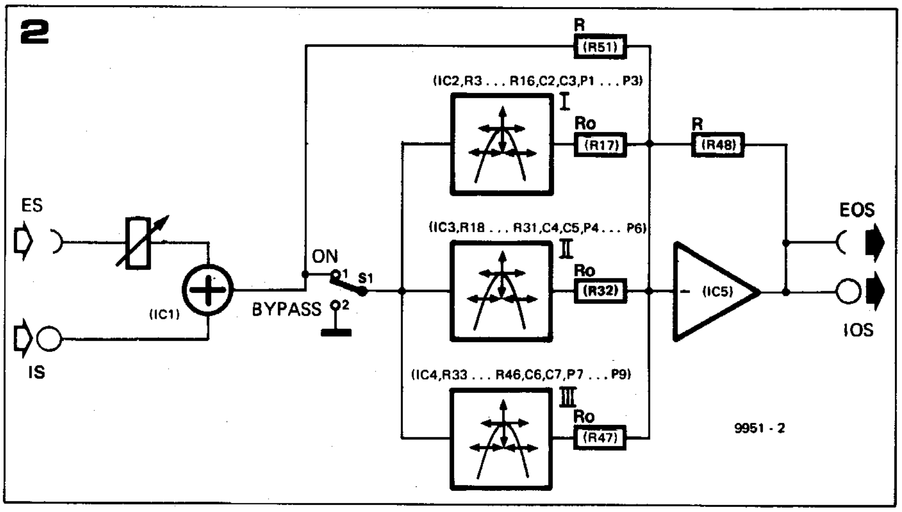




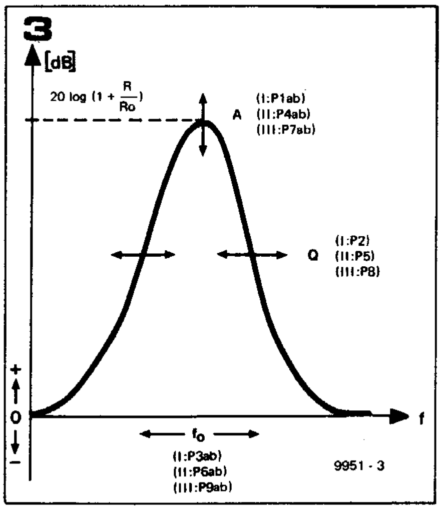


Table.

instrument	main resonance at	VCO signal
flute	approx. 800 Hz	fairly asym. squarewave
clarinet	1 . . . 2 kHz*	sym. squarewave
oboe	1300 . . 1700 Hz*	heavily asym. squarewave (pulse)
bassoon	approx. 440 Hz*	heavily asym. squarewave (pulse)
trumpet	approx. 1500 Hz	'spaced' sawtooth
bugle	approx. 1000 Hz*	sawtooth
trombone	approx. 600 Hz	'spaced' sawtooth
French horn	approx. 400 Hz*	sawtooth
tuba	approx. 250 Hz	sawtooth
violin	approx. 4000 Hz**	'spaced' sawtooth, sawtooth or heavily asym. squarewave (pulse)
cello	approx. 200 Hz**	
double bass	approx. 100 Hz**	

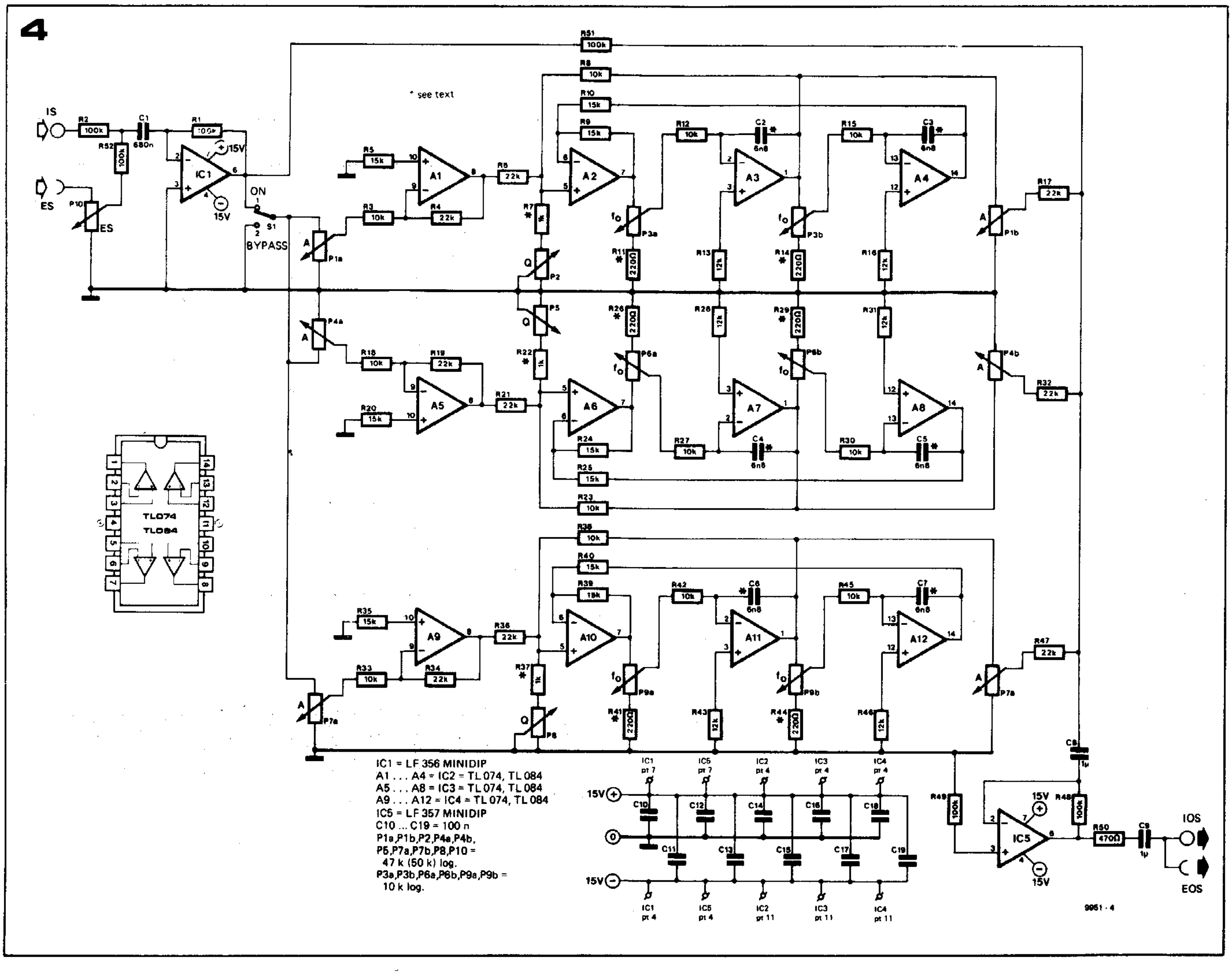
NB:
* with increased Q
** if possible, use several resonant filters (or a comb filter)

Figure 2. Block diagram of the resonant filter module. As can be seen, it possesses three independently variable filter sections.

Figure 3. The frequency response of one of the three filter sections contained in the resonant filter module. The figure illustrates how the filter parameters can be independently varied by means of the control potentiometers.

Figure 4. Detailed circuit diagram of the filter module.

Figure 5. Track pattern and component layout of the filter module p.c.b. (EPS 9951-1).



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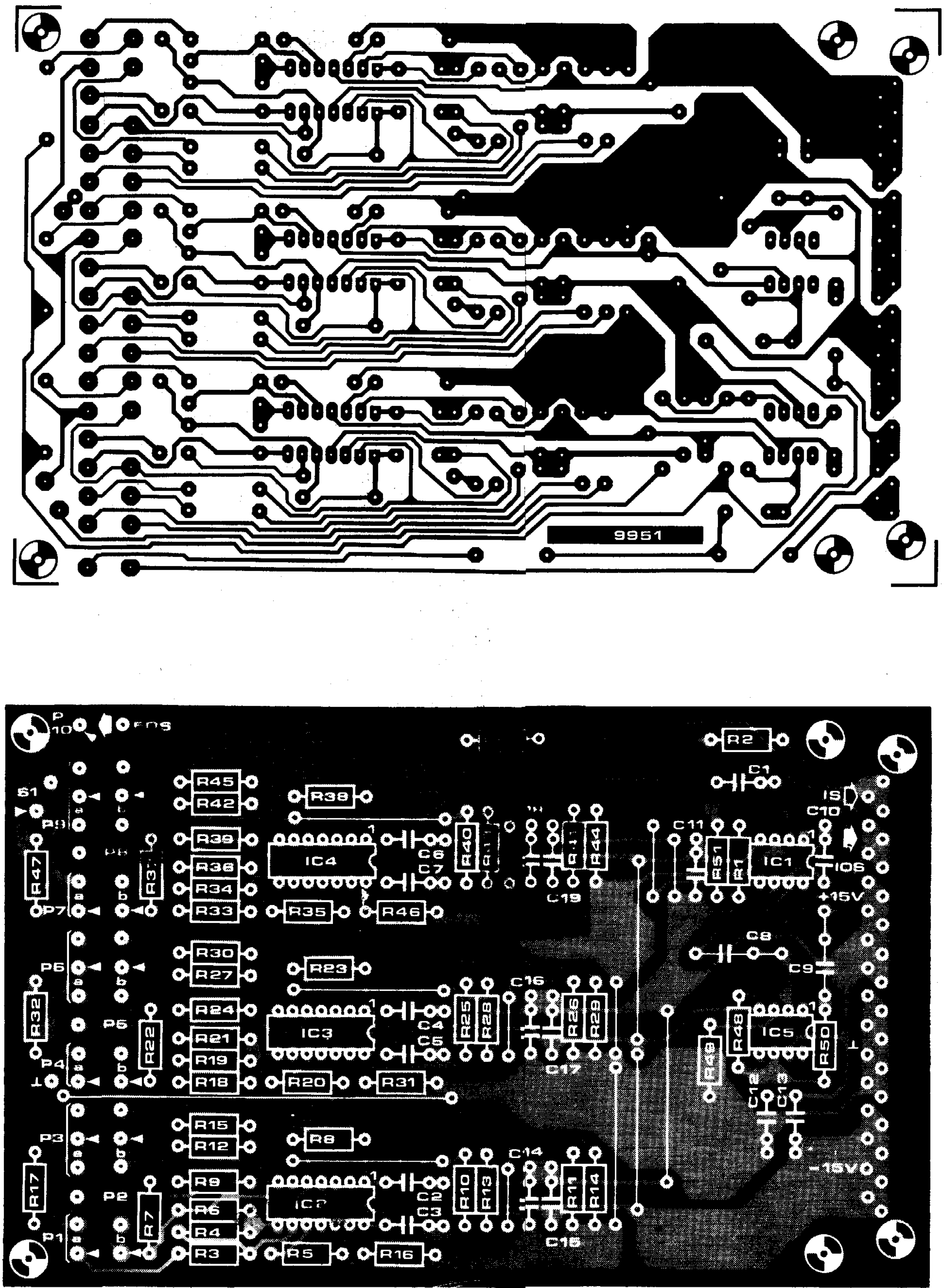
9951
R45
R42
R38
R39
R36
R34
R33
R35
R46
R47
IC4
C6
C7
R40
R44
C19
R2
C1
IS
C10
C11
R51
R1
IC1
IC6
+15V
C8
C9
R30
R27
R23
R24
R21
R19
R18
R20
R31
R32
R22
IC3
C4
C5
R25
R28
C16
R26
R29
C17
R48
IC5
R50
C12
C13
R15
R12
R8
R9
R6
R4
R3
R5
R16
R17
R7
C2
C3
R10
R13
C14
R11
R14
C15
-15V
P1
P2
P3
P4
P5
P6
P7
P9
S1

Parts list to figure 4 and 5.

Resistors:
R1,R2,R48,R49,R51,R52 = 100 k
R3,R8,R12,R15,R18,R23,
R27,R30,R38,R42,R45 = 10 k
R4,R6,R17,R19,R21,
R32,R34,R36,R47 = 22 k
R5,R9,R10,R20,R24,
R25,R35,R39,R40 = 15 k
R7,R22,R37 = 1 k (see text)
R11,R14,R26,R29,
R41,R44 = 220 Ω (see text)
R13,R16,R28,R31,R43,R46 = 12 k
R50 = 470 Ω

Potentiometers:
P1,P4,P7 = 47 k (50 k) logarithmic, stereo, dia 4 mm
P2,P5,P8,P10 = 47 k (50 k) logarithmic; dia 4 mm
P3,P6,P9 = 10 k logarithmic, stereo; dia 4 mm

Capacitors (all Siemens MKM, MKH or other polycarbonate/polyester type)
C1 = 680 n
C2,C3,C4,C5,C6,C7 = 6n8 (see text)
C8,C9 = 1 µ
C10 . . . C19 = 100 n

Semiconductors:
IC1 = LF 356 (National Semiconductors), Mini DIP
IC2,IC3,IC4 = TL 084, TL 074 (Texas Instruments)
IC5 = LF 357 (National Semiconductors) Mini DIP

Miscellaneous:
31-way DIN 41617 edge connector or terminal pins
S1 = miniature SPDT
2 miniature sockets 3.5 mm dia.
10 x 10 mm collet knobs (with pointer)
1 front panel

ous traditional instruments, with reference to a piano keyboard.

Circuit

The basic requirements of the filter circuit are, independently variable centre frequency, Q and gain. Since the function of the filter is essentially to enhance a particular band of frequencies (corresponding to the formants of the instrument in question), the circuit is of the boost-only type, i.e. provides selective gain. Without the need to provide a selective cut (below the 0 dB line) the circuit design is considerably simplified. A total of three resonant filters forms an acceptable compromise between the number of settings required for reasonably realistic imitation and the constraints of space and economy. Of course, it is quite possible to double the range of control facilities by connecting a second filter module in cascade with the first.

Block diagram

The block diagram of the resonant filter module is shown in figure 2. The figures

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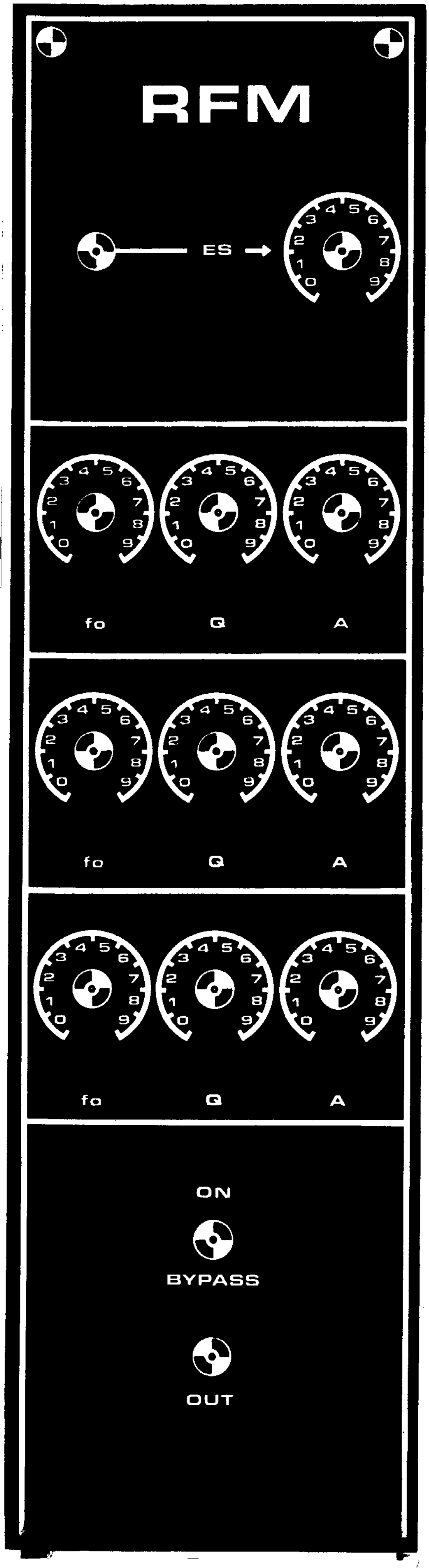

in brackets indicate which components in the final circuit are associated with the different sections of the circuit. Signals can be fed in via the panel-mounted socket (ES) or via the hard-wired input (IS). A portion of the signal is fed direct to the output summing amplifier via R (R51 in the complete circuit) and the input signal is also fed to three bandpass filters whose gain, centre-frequency and Q can all be varied. The outputs of these filters are also summed in IC5 via resistors R_O. The output of the filter module will thus consist of a portion of the original input signal plus signals boosted around the centre frequencies of the three filter stages. Two outputs are provided from the filter module, an internal hardwired output (IOS) and an output to a front panel socket (EOS). A bypass switch is provided, which allows the three filter sections to be switched out, in which case only the original signal appears at the output, and the gain is frequency independent, being unity.
The amount of boost that can be provided by a filter section relative to the gain obtained in the 'bypass' condition is determined by the gain of the filter sections and the ratio R/R_O. If it is assumed that the filter gain can be varied between zero and one then the maximum amount of boost (in dB) is $20 \log (1 + \frac{R}{R_O})$.
The frequency response of a filter section is shown in figure 3. The figures in parentheses indicate which controls in the complete circuit vary the different parameters of the filter.
The complete circuit of the filter module is shown in figure 4. IC1 sums and inverts the two input signals, whilst the three filter sections are of the state-variable type. The resonant gain of the filters is set by means of P1, P4 and P7 respectively. One gang of the pots is connected at the input, the other at the output of the filter. This has the effect of improving the dynamic range, since it means reduced noise and less chance of overloading. Finally, there is the inverting summing amplifier round IC5, which also cancels the phase shift introduced by IC1.

With the values for R and R_O given in the circuit diagram, the maximum gain of the filter is approx. +15 dB. The quality factor, Q, can be varied by P2 (P5, P8) between roughly 0.8 and 5. The centre frequency can be varied between approx. 50 and 2300 Hz, which is more than sufficient for normal use. The frequency range can, however, be modified by altering the value of a number of components; the necessary changes are detailed in the appendix.
Maximum Q is obtained for the minimum resistance of the Q-potentiometer. The maximum Q can therefore be increased by reducing the value of R7 (R22, R37); in this way a Q of between 20 and 30 can easily be obtained. A high Q is useful when processing waveforms such as squarewaves, which have very steep edges. These tend to set the filters 'ringing' at their resonant frequencies, and produce percussive effects. For R7 (R22, R37) = 470 Ω, a Q of 11.3 is obtained; R7 = 330 Ω gives a Q of 15.8, and R7 = 220 Ω a Q of 23.4. The higher the Q, the more pronounced the percussive effect.

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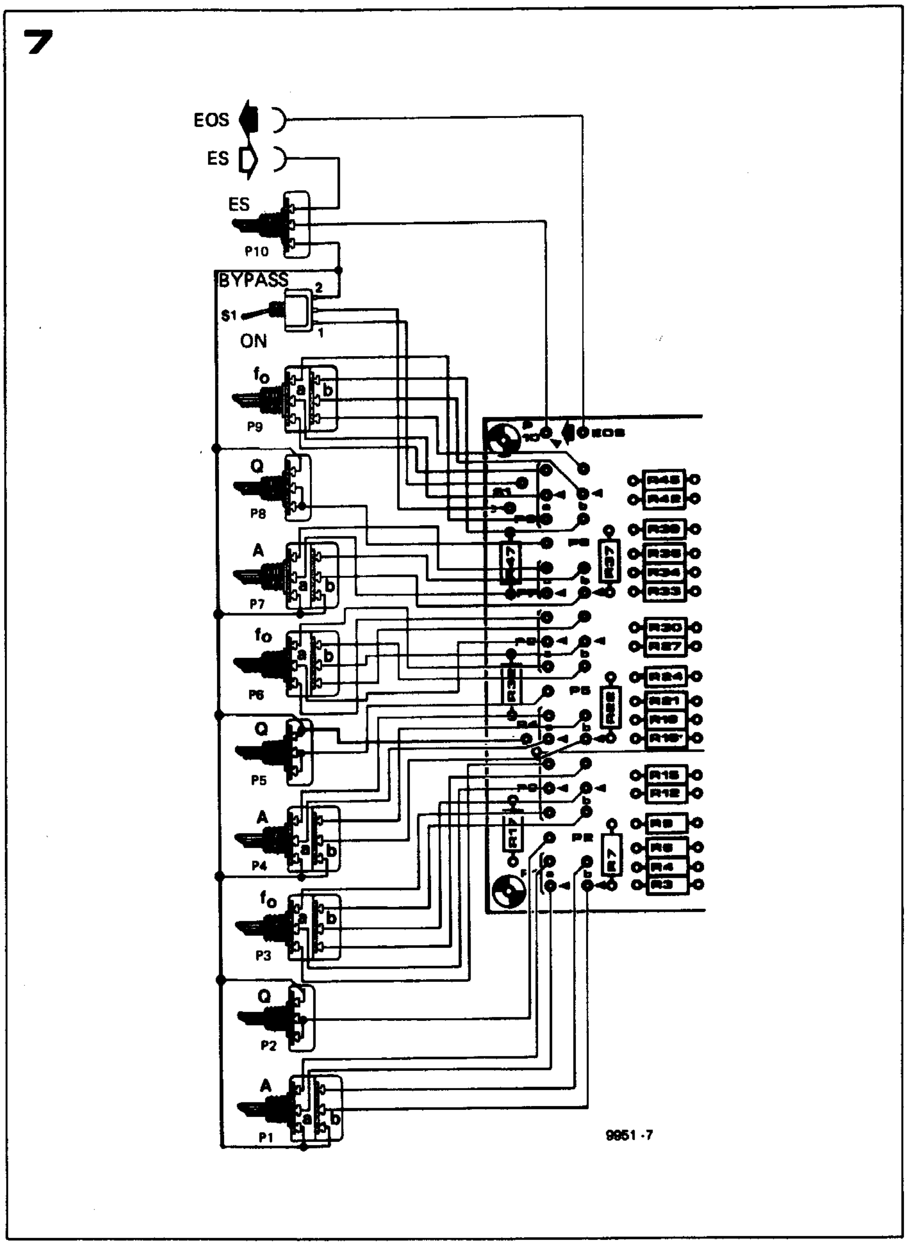

Figure 6. Because of the large number of controls, the front-panel for the resonant filter module is clearly different from the other Formant front-panels (EPS 9951-2).

Figure 7. Wiring diagram for the components mounted on the front-panel.

Construction

The printed circuit board for the resonant filter module is shown in figure 5.

As far as the selection of components is concerned, the usual criteria apply. The only difference is that in view of the large number of front-panel controls (10 potentiometers) it is strongly recommended that miniature components (miniature pots with 4 mm diameter spindles) be used. In this way the controls can be arranged in functional groups of three to a row.
The front panel for the filter module is shown in figure 6, and the details of the wiring for the front-panel controls are illustrated in figure 7. In contrast to the

other Formant modules, the resonant filter module requires no calibration or adjustment procedure. The operation of the circuit can be checked by feeding in a white noise input from the noise module. Varying the three filter parameters should produce clearly audible changes in the resulting sound. It will also be apparent that rapid variation of the Q- and f_0 controls produces effects similar to phasing, thus the filter module can be used to provide manual phasing.
The scale on each of the f_0 potentiometers on the front panel is calibrated with five nominal frequencies. The three middle settings in particular should be viewed as rough guidelines, since the resistance curve of logarithmic potentiometers can exhibit fairly wide tolerances.
The filter module should be placed between the COM-module and the power amp. However, if one wishes to use the headphone output on the COM-module, the resonant filter module can be connected directly before the latter.

Appendix

With the component values given in the circuit diagram, the centre frequency of the filters can be varied between roughly 50 and 2300 Hz. To calculate the correct values for higher frequencies than this, the procedure is as follows:
Firstly, the desired maximum frequency of f_0 can be used to calculate the value of C2 = C3 = C4 = C5 = C6 = C7 = C from the following equation:

$$C = \frac{16}{f_0 \max}$$

where C is in nanofarads and f_0 in kHz.
Secondly the value of resistor R (see figure 2) can be determined on the basis of the desired minimum centre frequency f_0 min:

$$R = \frac{16}{C \cdot f_0 \min}$$

where C is in nanofarads, R is in kΩ, and f_0 in kHz
The value of R_0 = R11 = R14 = R26 = R29 = R41 = R44 can be calculated from:

$$R_0 = \frac{10}{R - 2}$$

where R and R_0 are in kΩ. These equations can be used to check the values of figure 4.

chapter 9

ADSR

The ADSR (Attack-Decay-Sustain-Release) shaper is used to control the VCF and VCA modules and thereby determine the dynamic harmonic structure and dynamic amplitude characteristic of the VCO signals.

It is often not realised, even by musicians, how much the character of an instrument is determined by the dynamic amplitude and harmonic behaviour, rather than by the steady-state harmonic content of the instrument. If the attack and decay periods of a note are artificially modified, then the whole character of the sound is altered. An interesting and amusing experiment is to record the sounds of several musical instruments, but to remove the attack and decay periods by bringing up the recording level after the note starts and fading it down before the note ends. Then ask some musical friends to identify the instruments. They will no doubt be amazed how characterless the sound of an instrument becomes when robbed of its particular amplitude envelope.
On the other hand, starting with a single basic waveform such as the triangle output of the Formant VCO, a whole range of instrument sounds can be produced simply by varying the amplitude envelope, ranging from 'soft' sounds such as flute and some organ voices, to 'hard', percussive sounds such as piano and xylophone.
Envelope control of the harmonic content using the VCF allows even greater variation in the character of the sound.

Types of envelope curves

The envelope shaper of the synthesiser must be able to simulate the envelope contour of conventional musical instruments when the synthesiser is used in an imitative capacity, and also to produce envelopes that are purely synthetic in character (i.e. not found in sounds produced by normal acoustic methods). Fortunately, there are relatively few types of envelope contour that are musically important, and these are all fairly easy to generate electronically.

1. Attack/decay contour

The simplest type of envelope curve is that consisting only of attack and decay periods. The envelope contour rises to a peak when the note is played, and begins to decay immediately the peak is passed (see figure 1). By varying the attack and decay times a wide variety of sounds can be produced.
For example, if a rapid attack and slow decay is applied to the VCA control, then a percussive sound like a piano results. Applied to the VCF in the low-pass mode, the same envelope contour can produce very bright, metallic sounds, depending on the input waveform.
If the attack period is made long and the decay period short, then applying this to the VCA will produce completely synthetic 'fantasy' sounds similar to those obtained by playing a recording backwards. Applying this type of envelope contour to the VCF can produce sounds similar to those of a brass instrument played staccato.
However, the main use of this type of envelope curve is for the production of percussive sounds such as xylophone, marimba, glockenspiel, bells and gongs, cymbals, and struck or plucked strings such as guitar, banjo, harp, other string instruments played pizzicato, harpsichord, and of course, piano.

2. Attack-sustain-release contour

The attack/decay characteristic previously described is typical of instruments where the sound is initiated by a short pulse of energy (e.g. by striking or plucking a string), after which the sound dies away since there is no further excitation to sustain it. The envelope contour shown in figure 2 is typical of instruments in which a note is sounded and sustained, such as a pipe organ, woodwind instruments, and bowed string instruments. In a pipe organ the note builds up fairly rapidly after a key is depressed as standing wave modes are established in the pipe, and the note is sustained by virtue of the fact that air is continuously blown into the pipe. When the supply of air stops on releasing the key the note terminates more or less rapidly.
The same basic contour applies to woodwind instruments and to string instruments played with a bow, since the note is here again sustained by blowing or bowing. However, with such instruments much greater expression can be obtained by modulation of the

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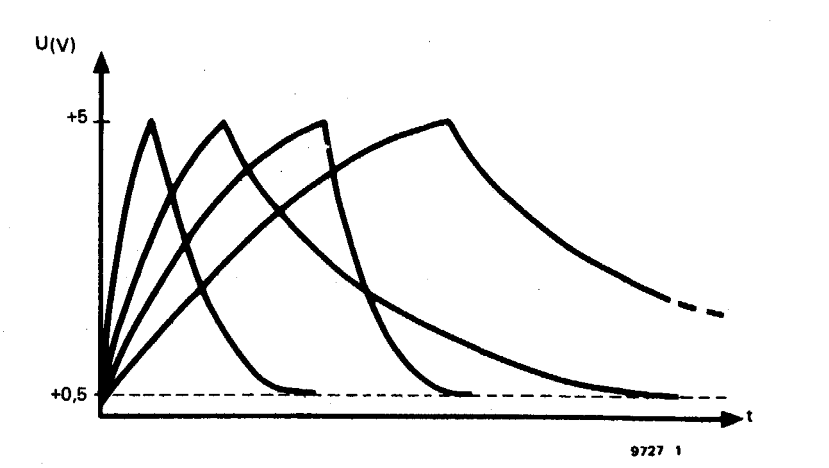

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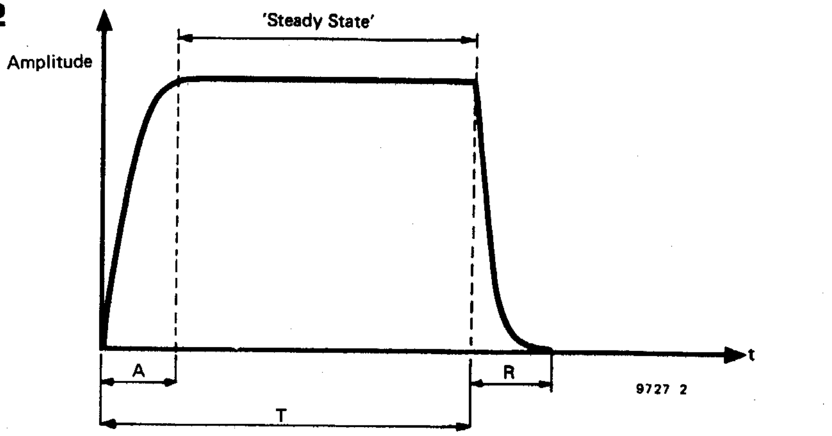

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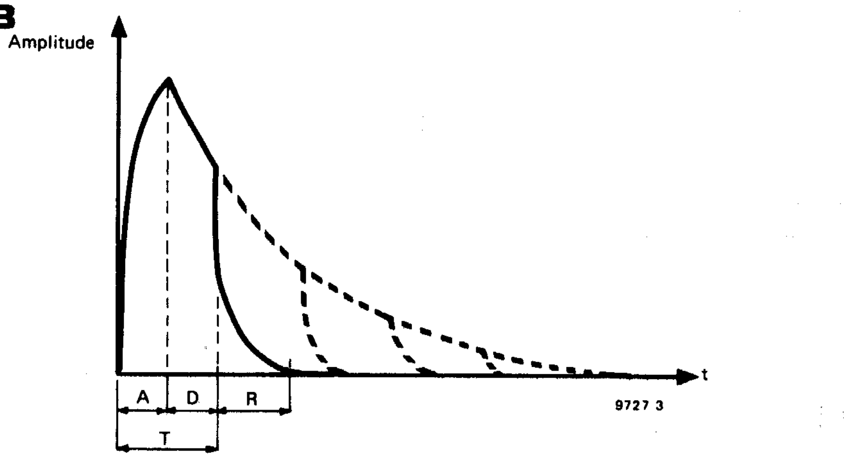

Figure 1. The attack-decay envelope contour is the simplest contour found in music.

Figure 2. The attack-sustain-release contour is used to simulate instruments where the note can be sustained at a constant level, such as organ, woodwind, and bowed string instruments.

Figure 3. Instruments such as the piano can be simulated using the attack-decay-release contour. As long as the key remains depressed the decay path is followed, but once the key is released the note is ended more abruptly, following the release contour.

steady-state level, since this is determined by the player, and not by a mechanical blower as is the case with a pipe organ.
With a synthesiser, a degree of expression can be obtained by modulating the VCA using the low-frequency oscillators or noise source.

3. Attack-decay-release contour

A variation on the attack-decay contour is shown in figure 3. Here the slow decay is allowed to continue for only a certain time, and the note is then terminated by a more rapid release. The most common example of this type of contour is provided by our old friend, the piano. When a note is sounded and the key remains depressed, then the damper is held off the string and the note decays over a period of a few seconds. If, however, the key is released after playing a note, the felt damper contacts the string and the note terminates after about 500 ms.

4. Attack-decay-sustain-release contour

Most of the examples given so far relate to envelope control of the VCA, since the amplitude contour of a sound is somewhat easier to visualise than its dynamic tone colour behaviour. However, the most complex envelope contour, shown in figure 4, is a good illustration of envelope control of the VCF.
Many brass instruments, such as the trumpet, are characterised by a rapid build-up of harmonics during the attack period of the note, which gives the instrument a very strident sound. Once the note is established, however, the harmonics die away somewhat, and the tone is much more mellow during the steady state period. Finally, during the release period at the end of the note, the note dies away fairly rapidly.
This type of characteristic can be obtained by using the VCF in the low-pass mode and controlling it with an envelope contour similar to that shown in figure 4. As the control voltage rises during the attack period, so the turnover frequency of the VCF increases, passing more harmonics. During the decay period the VCF turnover frequency falls until the steady-state value is reached, and finally, during the release period the VCF turnover frequency drops very rapidly.

Envelope shaper requirements

It is apparent from figure 5 that the envelope contours shown in figures 1 to 3 are merely special cases of the more general attack-decay-sustain-release contour illustrated in figure 4. Any of the four contours can be generated by an envelope shaper having the following four functions:

- variable attack time (A)
- variable decay time (D)
- variable sustain level (S)
- variable release time (R)

These four parameters can be preset manually using the ADSR controls of the envelope shaper. The envelope shaper is controlled by the gate pulse output of the keyboard. When a key is depressed the gate output goes high and this initiates the attack-decay sequence. The output of the envelope shaper then remains at the sustain level until the key is released, when the release period begins.

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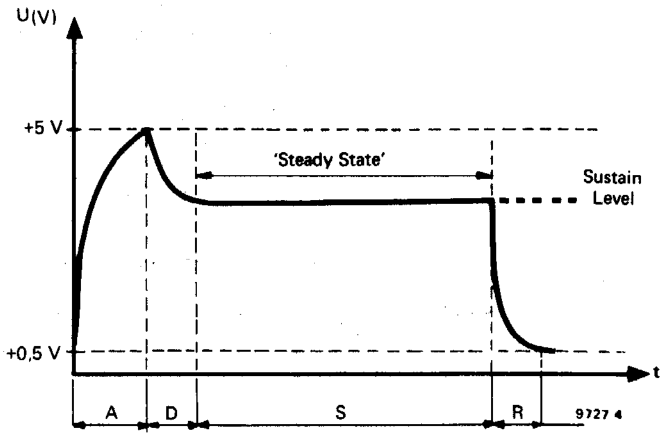

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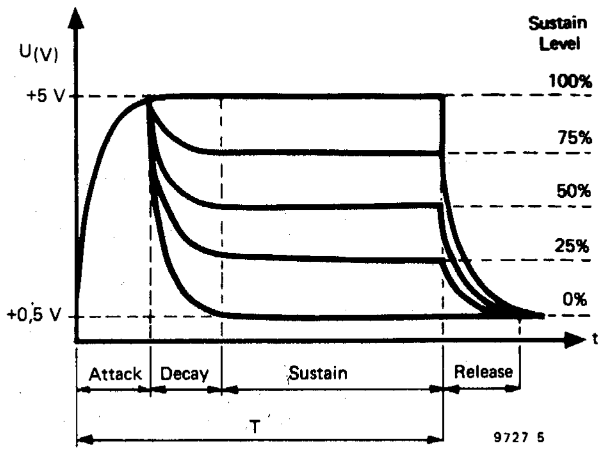

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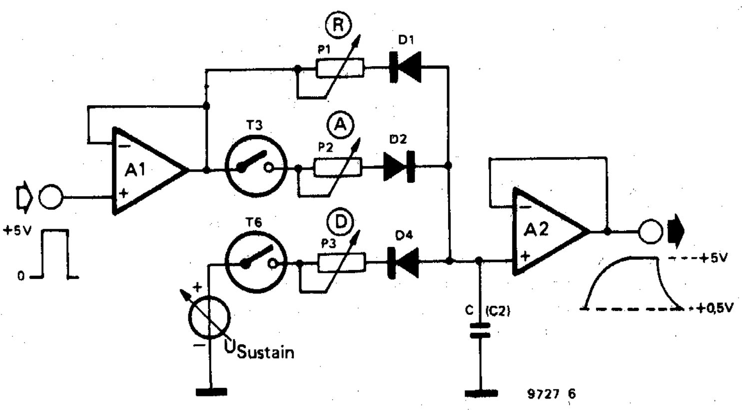


Block diagram

The required exponential attack, decay and release characteristics are easily obtained by charging and discharging a capacitor through resistors, and the sustain level by clamping the capacitor voltage to a preset D.C. level during the sustain period. The basic principle of the envelope shaper is illustrated in figure 6. The gate pulse is fed to a voltage follower A1, and when the gate pulse is high C charges exponentially through P2 and D2 (and T3).

At the end of the Attack period, 'switch' T3 is opened and T6 is closed. Capacitor C now discharges through D4 and P3 (Decay), until the Sustain level is reached. This level is maintained until the gate pulse finishes, either when the key is released or when a preset time has elapsed.

When the gate pulse finishes, the output of A1 goes to zero volts, and C discharges through D1 and P1 (Release). The capacitor cannot discharge fully,

ADSR adjustment ranges:

Attack period	(A)	10 ms . . . 20 s
Decay period	(D)	10 ms . . . 20 s
Sustain level	(S)	0.5 V . . . 5 V
Release period	(R)	10 ms . . . 20 s

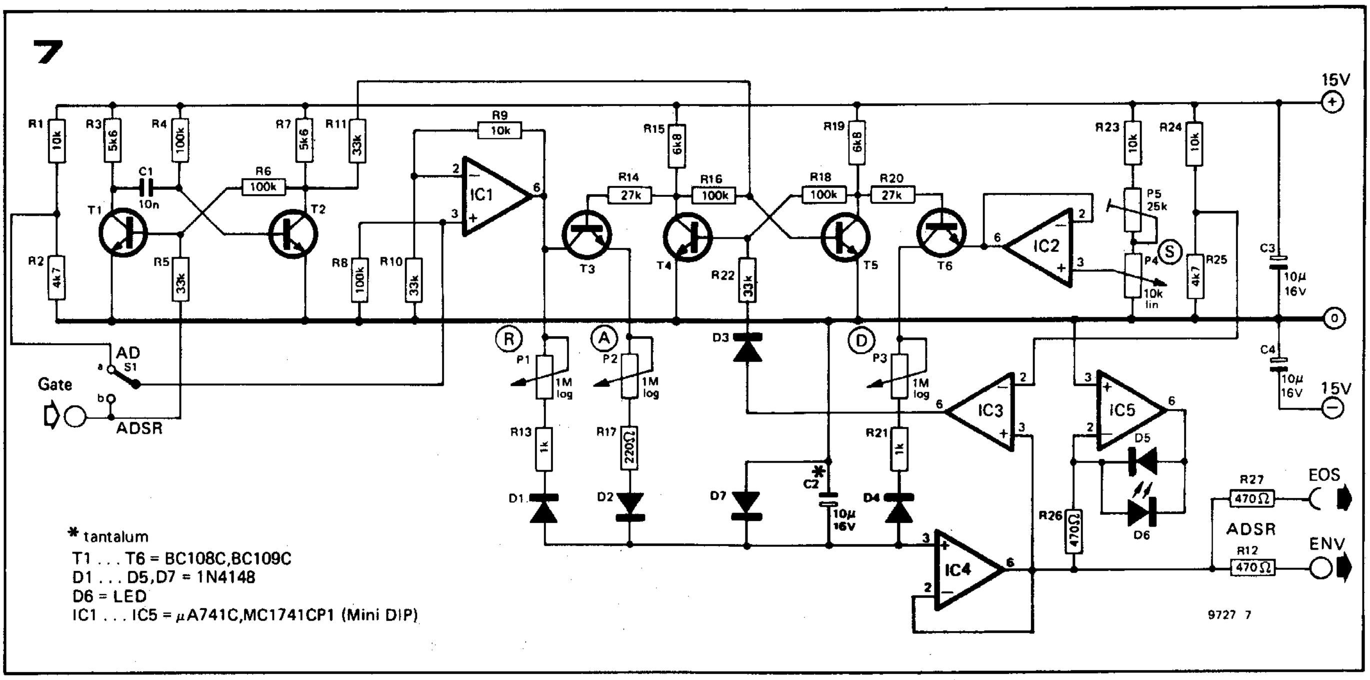
since D1 ceases to conduct once the voltage on C has fallen to about 0.5 V, but this is not important as it merely constitutes a D.C. offset which can be compensated for. The attack, decay and release times may be adjusted by means of P2, P3 and P1.

Complete circuit

The complete circuit, which is shown in figure 7, is, of course, more complicated. The envelope shaper has two modes of operation, ADSR and AD, which are selected by means of S1. With S1 in position 'b' (ADSR) the circuit operates as follows:

When a key is depressed the gate pulse output goes to +5 V. IC1 has a gain slightly greater than unity, so about +6 V appears at its output.

The leading edge of the gate pulse also triggers mo.iostable T1/T2, which pro-



duces a short pulse to set flip-flop T4/T5 (T5 turned on and T4 turned off). The collector voltage of T4 thus rises, turning on T3 and allowing C2 to charge from the output of IC1 through T3, P2, R17 and D2. This is the attack period.

The voltage on C2 is fed to voltage-follower buffer IC4, which is connected to the outputs EOS and ENV and also to the non-inverting input of IC3. This IC functions as a comparator, with its inverting input held at about 4.7 V by R24 and R25. When the voltage on C2, and hence at the output of IC4, exceeds this value, the output of IC3 swings positive, resetting flip-flop T4/T5, turning off T3 and terminating the attack period. T6 is turned on, initiating the decay period when C2 discharges through D4, R21, P3 and T6 into the output of IC2 until the sustain level, set at the output of voltage follower IC2 by P4, is reached.

The output of the envelope shaper then remains at the sustain level until the key is released, when the output of IC1 goes to zero volts and C2 discharges through D1, R13 and P1 (release period).

Diode D7 protects C2 in the event of the output of IC1 going negative for any reason, when the voltage across C2 is clamped to a maximum of −0.7 V.

A LED indicator constructed around IC5 allows visual monitoring of the envelope contour. The brightness of the LED follows the envelope voltage.

Two outputs are provided from the envelope shaper; an external output to a front panel socket (EOS), and an internally wired output (ENV).

The full ADSR envelope contour is, of course, produced only if the key is depressed for a period longer than the attack plus decay time, and if the sustain level is greater than 0%. If the key is released before the sustain level is reached then the release period is initiated prematurely, and either AR or ADR curves may be produced. If the

Figure 4. The attack-decay-sustain-release contour is the most complex envelope shape provided by the Formant envelope shaper. When applied to the VCF it is useful for imitating brass instruments, where the harmonic content of the note rises initially to a large value, then reduces to a lower level during the steady-state part of the note.

Figure 5. By varying the sustain level the envelope contour can be changed from an AD contour at 0% sustain, through various ADSR contours to an ASR contour at 100% sustain. T is the time for which the key remains depressed.

Figure 6. This simplified diagram illustrates the basic principle of the envelope shaper. C charges through D2 and P2 during the attack period. It then discharges through D4 and P3 to the (adjustable) sustain level; finally, it discharges through D1 and P1 during the release period. P1, P2 and P3 can be used to vary the release, attack and decay times.

Figure 7. Complete circuit of the Formant envelope shaper.

sustain level is 0% then only AD or ADR curves may be produced, depending on when the key is released. If the sustain level is 100% then, of course, only AR or ASR curves may be produced, depending on when the key is released, since the decay period is inhibited.

Triggered AD mode

It is sometimes useful to be able to produce AD envelope contours that are unaffected be releasing the key, that is to say, once the key is depressed, a fixed attack-decay sequence is initiated, which is completed whether the key is released or not. This triggered AD contour is obtained by setting S1 to position 'a' and selecting 0% sustain level. The input of IC1 is now connected to the junction of R1 and R2, so its output is permanently at about +6 V, irrespective of the gate input.

When a key is depressed, the gate signal triggers the monostable, setting the flip-flop and turning on T3. At the end of the attack period, comparator IC3 resets the flip-flop, turning on T6 and initiating the decay period. C2 will now discharge through D4, R21, P3 and T6 to the 0% level (sustain is set at 0%).

Even if the key is released before this sequence is complete, the release period is inhibited since the output of IC1 is permanently at +6 V, so C2 cannot discharge through D1, R13 and P1.

Construction

There are no special requirements with regard to resistor tolerances in the envelope shaper circuit, and ordinary, good-quality 5% carbon film components are quite adequate; C2 should be a tantalum electrolytic capacitor for low leakage, and C1 the usual

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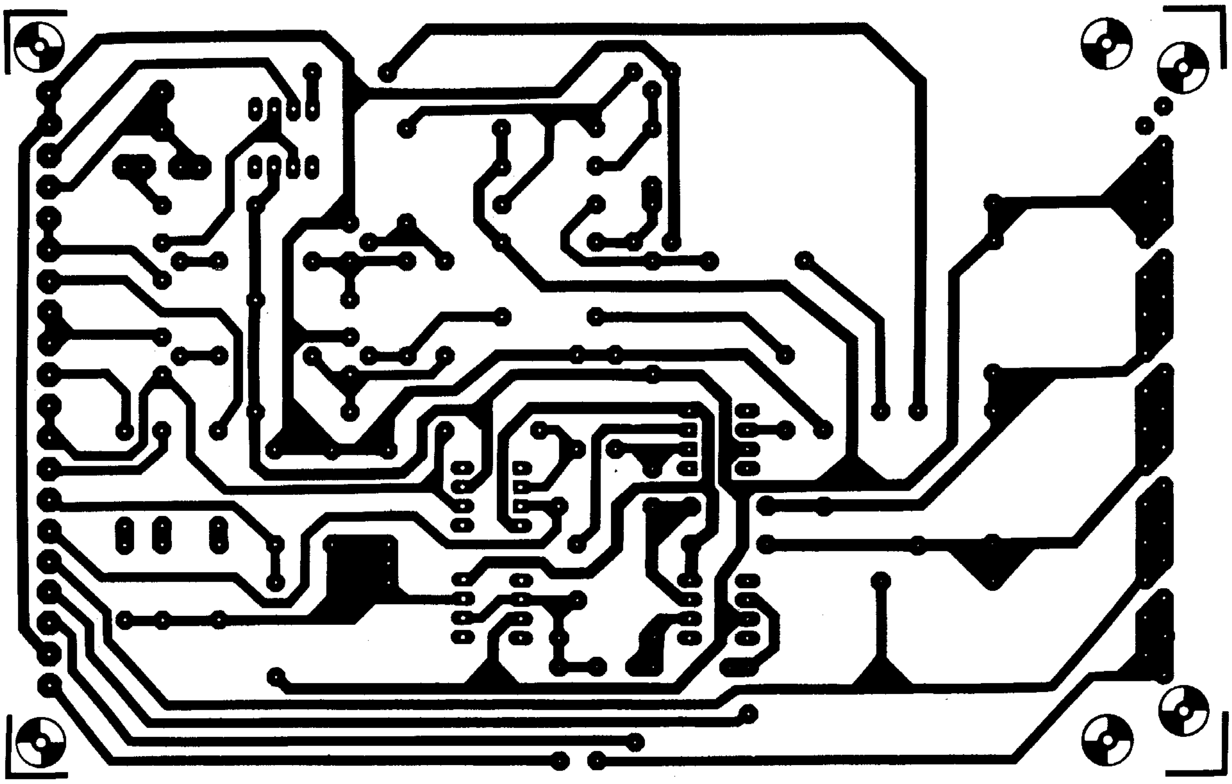

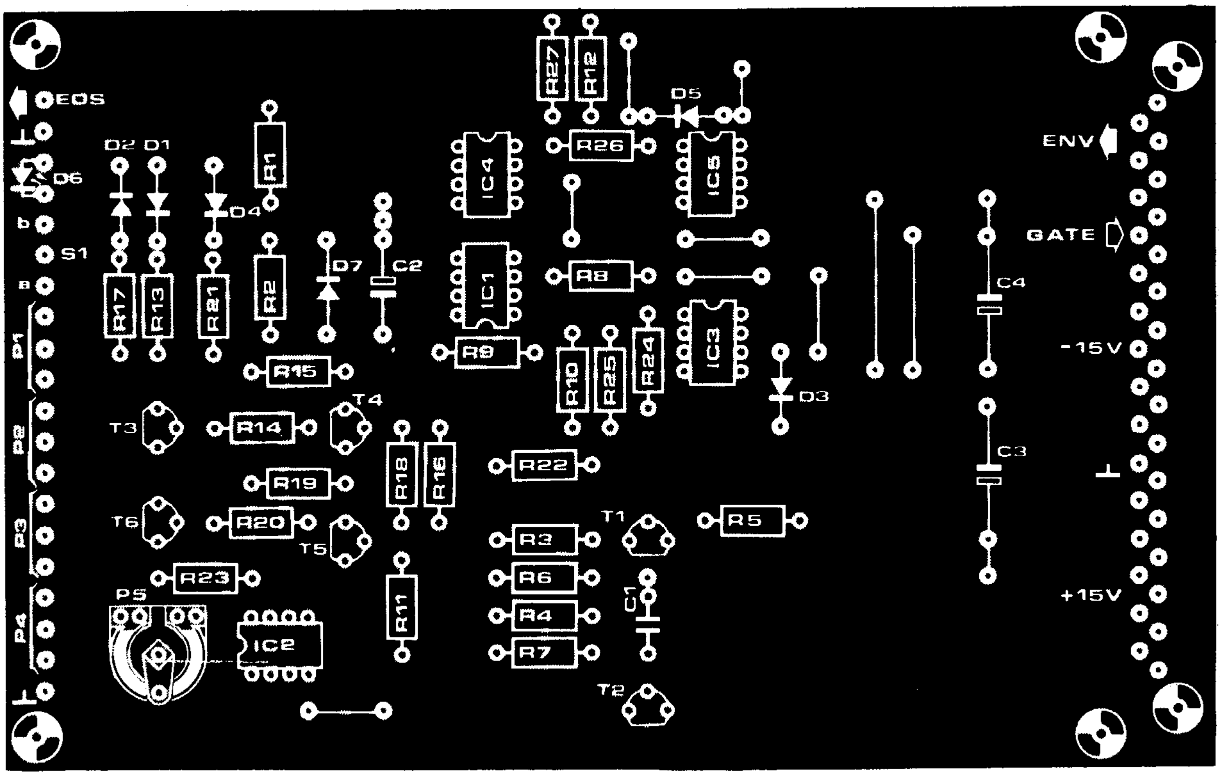
EOS
D6
b
S1
a
D2 D1
D4
D7
C2
P1
P2
P3
P4
R1
R2
R17
R13
R21
R15
T3
R14
T4
R18
R16
R19
T6
R20
T5
R23
P5
IC2
R11
R27
R12
D5
R26
IC4
IC5
IC1
R8
IC3
R9
R10
R25
R24
D3
R22
R3
R6
R4
R7
T1
R5
C1
T2
C4
C3
ENV
GATE
-15V
+15V

Parts list for figures 7 and 8

Resistors:
R1,R9,R23 = 10 k
R2,R25 = 4k7
R3,R7 = 5k6
R4,R6,R8,R16,R18 = 100 k
R5,R10,R11,R22 = 33 k
R12,R26,R27 = 470 Ω
R13,R21 = 1 k
R14,R20 = 27 k
R15,R19 = 6k8
R17 = 220 Ω

Potentiometers:
P1,P2,P3 = 1 M log.
P4 = 10 k lin.
P5 = 25 k preset

Semiconductors:
T1 . . . T6 = BC108C, BC109C or equivalent
D1 . . . D5,D7 = 1N4148, 1N914
D6 = LED (TIL209 or similar)
IC1 . . . IC5 = μA 741C, MC1741 CP1 (MINI DIP)

Capacitors:
C1 = 10 n
C2 = 10 μ/16 V tantalum
C3,C4 = 10 μ/16 V

Miscellaneous:
31-way Euro connector (DIN41617)
1 x 3.5 mm jack socket
4 x 13 . . . 15 mm collet knobs with pointer

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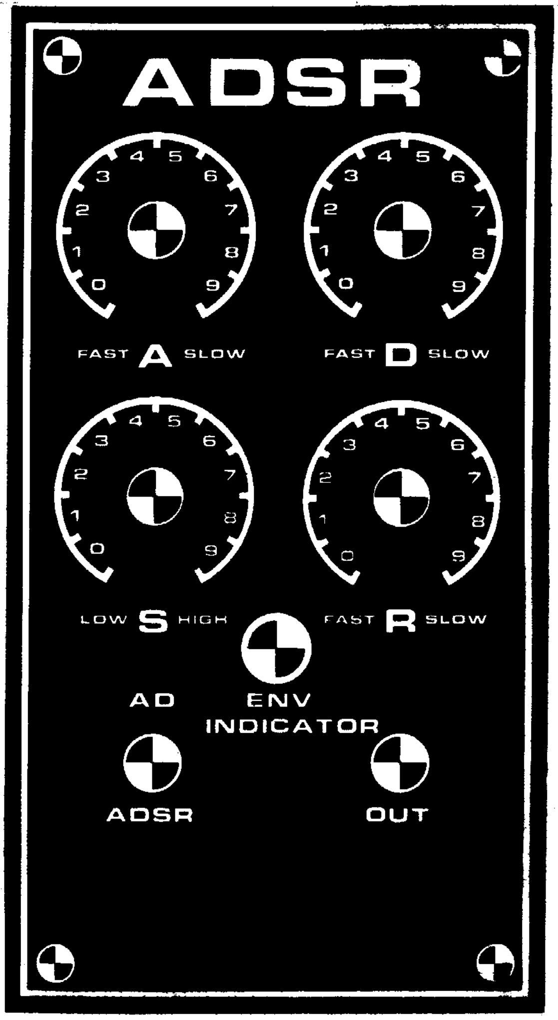

polyester or polycarbonate type. It is a good idea to test T3 and T6 for leakage, using the method detailed in chapter 5.
A printed circuit board and component layout for the envelope shaper are given in figure 8, and a front panel layout is shown in figure 9. Connections to the front panel are fairly simple, the only front panel-mounted components being the four potentiometers for attack time, decay time, release time and sustain level, switch S1, the external output socket and the envelope indicator LED.

Testing and adjustment

To test the envelope shaper a gate pulse must be available from the 'GATE' output of the interface receiver board. The EOS output of the envelope shaper is monitored on an oscilloscope with the Y sensitivity set to about 1 V/div and the timebase set to about 10 ms/div.
For the first test, the sustain level is set to zero, S1 is set to the 'AD' position and the attack and decay potentiometers are set to 'fast'. The release potentiometer has no effect during this test. If a key is depressed at short intervals then a short AD envelope curve will be seen, which rises and falls between about 0.5 V and 5 V. The output of IC3 can also be monitored, to check that it swings briefly between −15 V and +15 V when the peak of the attack curve is reached.
The only adjustment required to the envelope shaper is to set the 100% sustain level, using P5, to correspond with the voltage on C2 at the end of the attack period. If it is too low, then there will always be a decay, even at 100% sustain level; if it is too high then the calibration of P4 will be inaccurate, since 100% sustain will be reached before maximum rotation of the potentiometer.
To make the adjustment, the sustain level is set to 100% and medium attack and decay times are selected. Preset P5 is then adjusted until there is just no decay after the attack period (i.e. the attack period blends into the sustain level with no dip). The adjustment can be checked by turning P4 slightly to the left, when a slight dip after the peak of the attack period should be noted. As P4 is turned further anticlockwise then the decay down to the sustain level will become greater and greater, until finally, at 0% sustain level, pure AD curves will be produced.
The envelope shaper is now ready for use.

Figure 8. Printed circuit board and component layout for the envelope shaper (EPS 9725-1).

Figure 9. Front panel layout for the envelope shaper module.