Bridged-T selects filter's notch frequency and bandwidth

by P. V. Ananda Mohan
Indian Telephone Industries Ltd., Bangalore, India

If a bridged-T network is used in place of the Wien bridge in the notch filter proposed by Fellot,[1] both the bandwidth and the frequency may be independently adjusted. The bridged-T approach has been explored previously[2] as an extension of some work carried out on parallel-T notch filters, and, as illustrated here, the technique offers an excellent way of building units that are simple and versatile.

R_N and R_Q comprise the balancing arms of the bridged-T network (note A_1 is a unity-gain buffer) as seen in (a). In this configuration, the circuit's transfer function is:

$$e_o/e_i = [ns^2+\omega_o^2]/[s^2+3(1\text{-}q)s\omega_o+\omega_o^2]$$

where n and q are selected by R_N and R_Q, respectively, and $\omega_o = 1/RC$. Note that $0 \leq n,\ q \leq 1$, and that the frequency of the notch is

$$\omega_n = \omega_o/n^{1/2}$$

Therefore for this circuit ω_n will always be equal to or greater than ω_o.

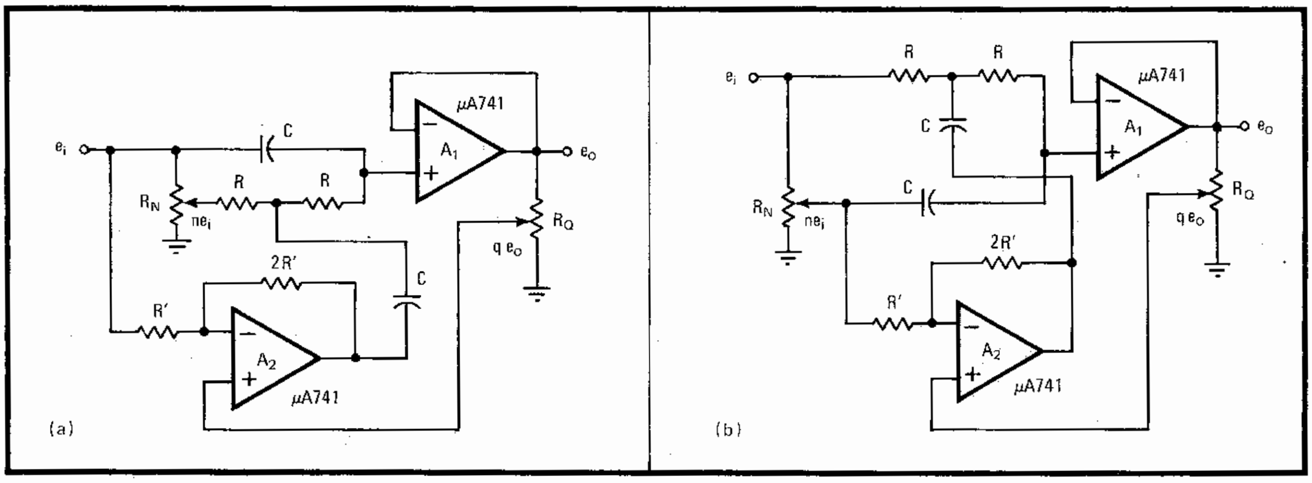
The bandwidth is adjusted with R_Q, and Qs greater than 1,000 will be realized when high-gain operational amplifiers are used. In general, Qs will be higher than can be achieved with parallel-T networks. The notch depth is at least 50 decibels throughout the operating range. R_N and R_Q must only be at least 10 times smaller than R to achieve the stated filter characteristics.

By modifying the circuit slightly, as in (b), the transfer function becomes:

$$e_o/e_i = [s^2+n\omega_o^2]/[s^2+3(1\text{-}q)s\omega_o+\omega_o^2]$$

and the notch frequency ω_n is made tunable for frequencies below ω_o, so that $\omega_n = \omega_o n^{1/2}$. □

References
1. Dominique Fellot, "Wien bridge and op amp select notch filter's bandwidth," *Electronics*, Dec. 7, 1978, p. 124.
2. "An Active RC Bridged-T Notch Filter," *Proc. IEEE*, August 1977, p. 208.

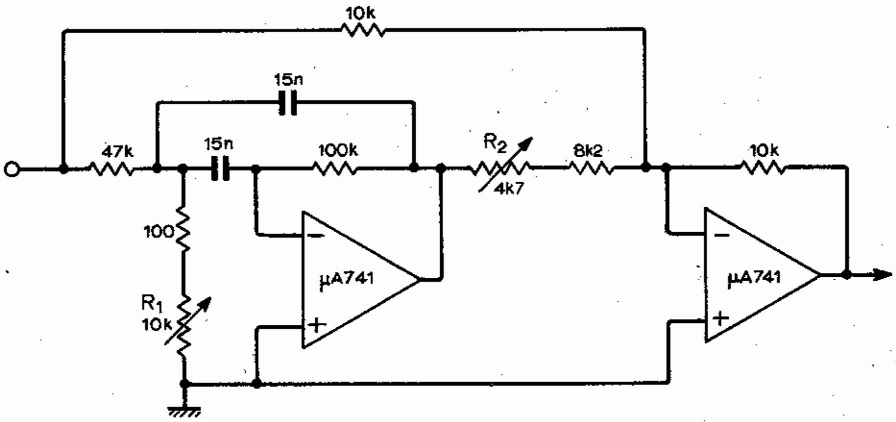


Changing tune. Using bridged-T network in place of Wien-bridge arrangement in notch filter enables independent control of filter's bandwidth and frequency. Circuit can be configured for tuning filter above (a) or below (b) its natural radian frequency $\omega_o = 1/RC$.

Circuit Ideas

Concise descriptions of new circuits are invited for Circuit Ideas, for which £5 is paid on publication. Contributors should say how their circuit is an improvement over existing circuits, preferably in the first sentence.

Simple tunable notch filter



Most active *RC* notch filters are difficult to tune because the rejection at the notch frequency depends on the accurate ganging of several potentiometers or variable capacitors (three in the case of the twin-T filter). The circuit shown has a notch frequency which can be varied by one potentiometer only, the notch rejection being independent of the setting and the bandwidth between points of 3dB attenuation on either side of the notch being independent of the frequency to which the filter is tuned. The circuit consists of a bandpass filter tuned by R_1 followed by a virtual earth summing circuit that adds the output of this filter with the input signal. Notch rejection is set to maximum by adjustment of R_2. Using the values shown the filter tunes from 170Hz to 3kHz, a bandwidth between 3dB points of 230Hz and a notch rejection of better than 40dB over the complete range. A voltage-tuned notch filter may be realised by replacing R_1 with a f.e.t. operated as a voltage-variable resistor.

R. J. Harris,
Wells,
Somerset.

Tone detector sharpens digital filter's response

by Steve Newman
Los Angeles, Calif.

Using digital techniques to set the center frequency and the passband, this circuit will serve as a precision tone detector or as a control (transmission) gate for digital bandpass or band-blocking filters. The circuit elements can be easily cascaded to provide as great a degree of selectivity as required.

The general scheme is outlined in (a). Up counters A_1 and A_2 are preprogrammed to generate a carry signal after N or M input pulses of f_{in}, respectively, where N is equal to or less than 16 and N is greater than M. The period of the reference signal f_{ref}, which is t_{ref}, determines the time each counter is enabled.

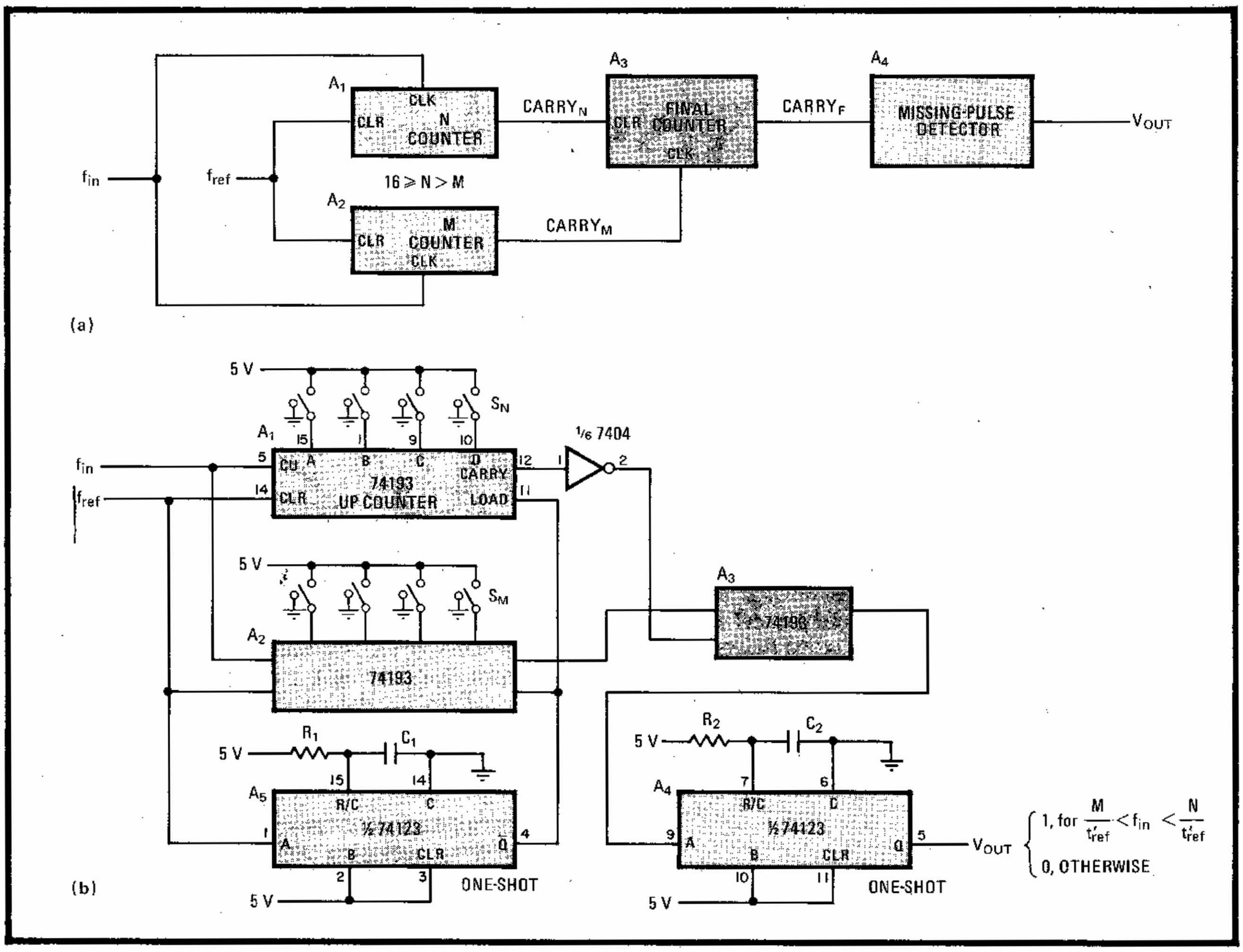
The output of the M counter clocks the presettable counter A_3, while the output of the N counter is connected to its clear input. Thus if M or more pulses occur during the time $\frac{1}{2}t_{ref}$, hereafter called t'_{ref}, the final counter generates carry pulses periodically (this circuit is intended to detect tones of fairly extended duration only). As a result, the output of the missing-pulse detector, A_4, can be forced high.

If, on the other hand, the incoming frequency is greater than N/t'_{ref}, presettable counter A_3 is always reset before it can produce a carry pulse. Thus A_4, which requires a steady stream of pulses to keep it active, is forced low. If f_{in} is less than M/t'_{ref}, A_3 again cannot produce a carry pulse because there are no clock pulses from the M counter. Thus, V_{out} will be high only for the case where $M/t'_{ref} < f_{in} < N/t'_{ref}$.

The actual circuit is shown in (b). A_1 and A_2 are enabled at the instant f_{ref} moves low. A_3 and A_4 assume the same functions described in (a). If the detection process must be sped up, A_3 can be preset so that it generates a carry for every Pth pulse from A_2, in the range 1 to 15. In general operation, however, A_3 is not preset.

At the start of the measurement cycle, f_{ref} initiates the process whereby one shot A_5 presets A_1 and A_2 to their switch-programmed values (switches S_N and S_M are active low). The reference frequency is selected in conjunction with S_M and S_N to provide almost any desired passband.

For example, if a frequency between 9.5 and 10.5 kilohertz (f_{in}) must be detected, M can be arbitrarily selected to be 9 and N = 10. A t'_{ref} = 0.95 millisecond is then required (f_{ref} = 526 hertz at 50% duty cycle). Note the time constant R_1C_1 must be much less than t_{ref}.



Logical boundaries. Tone detector sets limits of desired frequency range digitally. Using counting technique, circuit rejects tones whose frequencies are too high or low. Detector generates output only for $2M/t_{ref} < f_{in} < 2N/t_{ref}$, where f_{in} is input frequency, M and N are the Mth and Nth pulses that generate a carry from the M or N counters, respectively, and $t_{ref} = 1/f_{ref}$, where f_{ref} is the reference frequency.

R_2C_2 must just exceed $240/f_{in}$ to cover the case where $M = 15$ and $N = 16$, whereupon as many as 240 input pulses are required to detect the tone. Under these conditions, however, the bandpass will be only 3% of the center frequency, f_o.

If greater selectivity is desired, a pair (or more) of 74193 up counters can be simply cascaded without upsetting the basic operation of the M and N counter chains. With two counters in each chain, the range of M and N can be expanded to 256, whereupon a bandpass of only 0.2% of f_o can be selected. □

Designer's casebook is a regular feature in *Electronics*. We invite readers to submit original and unpublished circuit ideas and solutions to design problems. Explain briefly but thoroughly the circuit's operating principle and purpose. We'll pay $50 for each item published.

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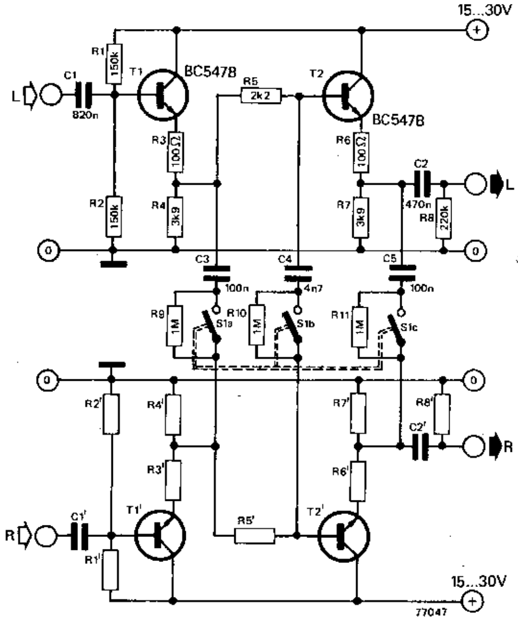
stereo noise filter

The signal-to-noise ratio of an FM broadcast received in stereo is considerably worse than that of the same broadcast received in mono. This is most noticeable on weak transmissions, when switching over from stereo to mono will considerably reduce the noise level. This noise reduction occurs because the left-channel noise is largely in anti-phase to the right-channel noise. Switching to mono sums the two channels and the anti-phase noise signals cancel.

By summing only the high-frequency components of the signal it is possible to eliminate the annoying high-frequency noise without destroying the stereo image since channel separation is still maintained at middle and low frequencies.

Each channel of the circuit consists of a pair of emitter followers in cascade, with highpass filters comprising R3 to R7 and C3 to C5 that allow crosstalk to occur between the two channels above about 8 kHz when switch S1 is closed. When S1 is open the two channels are isolated, but resistors R9 to R11 maintain a DC level on C3 to C5 so that switching clicks do not occur when S1 is closed.

The stereo-mono crossover frequency can be increased by lowering the values of C3 to C5 or decreased by raising them.



Synchronous noise blanker cleans up audio signals

by M.J. Salvati
Sony Corp. of America, Long Island City, N.Y.

Fluorescent lights, gas rectifiers, neon lamps, SCRs, and triacs all produce a substantial rf signal that often radiates through their power-line connections and interferes with nearby communications receivers. This type of radio interference desensitizes the receiver and makes the recovered audio signal very difficult to understand.

The circuit shown here significantly improves the audio intelligibility of a receiver by eliminating the noise pulses generated by a single dominant nearby noise source. The noise pulses are removed from the audio signal with only slight distortion. Moreover, since this noise-blanking circuit is not internally connected to the receiver, it can be moved from one receiver to another as needed.

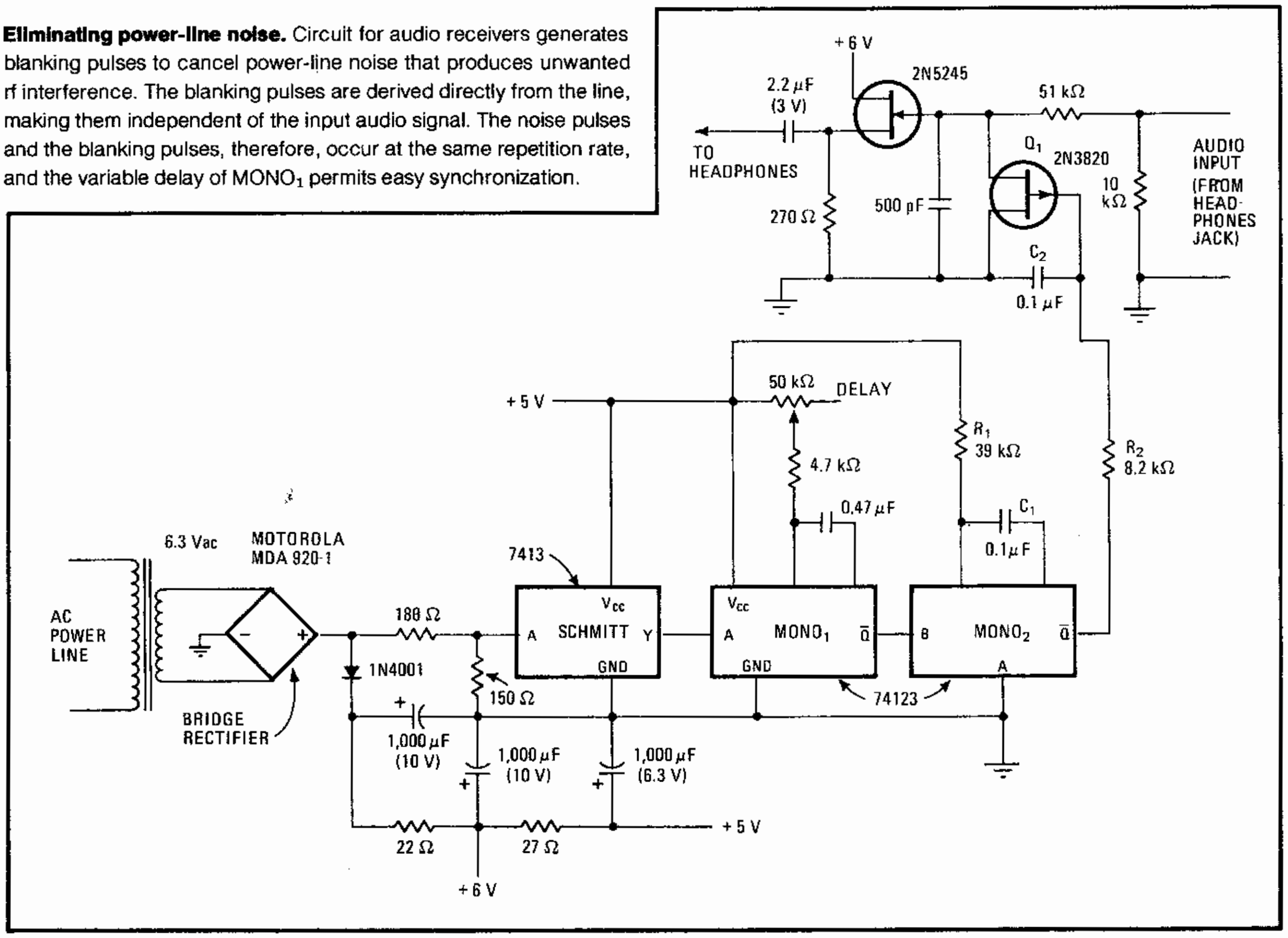
The noise pulses produced by power-line radiation occur at a repetition rate of twice the local power-line frequency. Since the noise-blanking circuit is driven by the same power utility as the noise source, the output signal from the bridge-rectifier section of the noise blanker will have the same rate as the noise pulses.

The source of the blanking pulses, therefore, is independent of the input audio signal. The blanking pulses cause the FET gate (transistor Q_1) to conduct to silence the receiver. Since the blanking pulses are not derived from the input signal, their timing does not depend on the shape and rise time of the noise pulses, nor is it affected by the modulation characteristics of the desired signal.

The output from the bridge rectifier is shaped by a Schmitt trigger that drives a dual monostable multivibrator. The first monostable ($MONO_1$) delays the blanking pulse, which is produced by the second monostable ($MONO_2$), relative to the rectifier's output. The delay is variable so that the blanking pulse can be positioned to coincide with the noise pulse.

The width of the blanking pulse is determined by resistor R_1 and capacitor C_1. The fast rise time of the blanking pulse (from $MONO_2$) is slowed down by the low-pass filter formed by resistor R_2 and capacitor C_2, thereby minimizing the distortion of the recovered audio signal □

Eliminating power-line noise. Circuit for audio receivers generates blanking pulses to cancel power-line noise that produces unwanted rf interference. The blanking pulses are derived directly from the line, making them independent of the input audio signal. The noise pulses and the blanking pulses, therefore, occur at the same repetition rate, and the variable delay of $MONO_1$ permits easy synchronization.



Logic-gate filter handles digital signals

by Andrzej M. Cisek
Electronics for Medicine, Honeywell Inc., Pleasantville, N. Y.

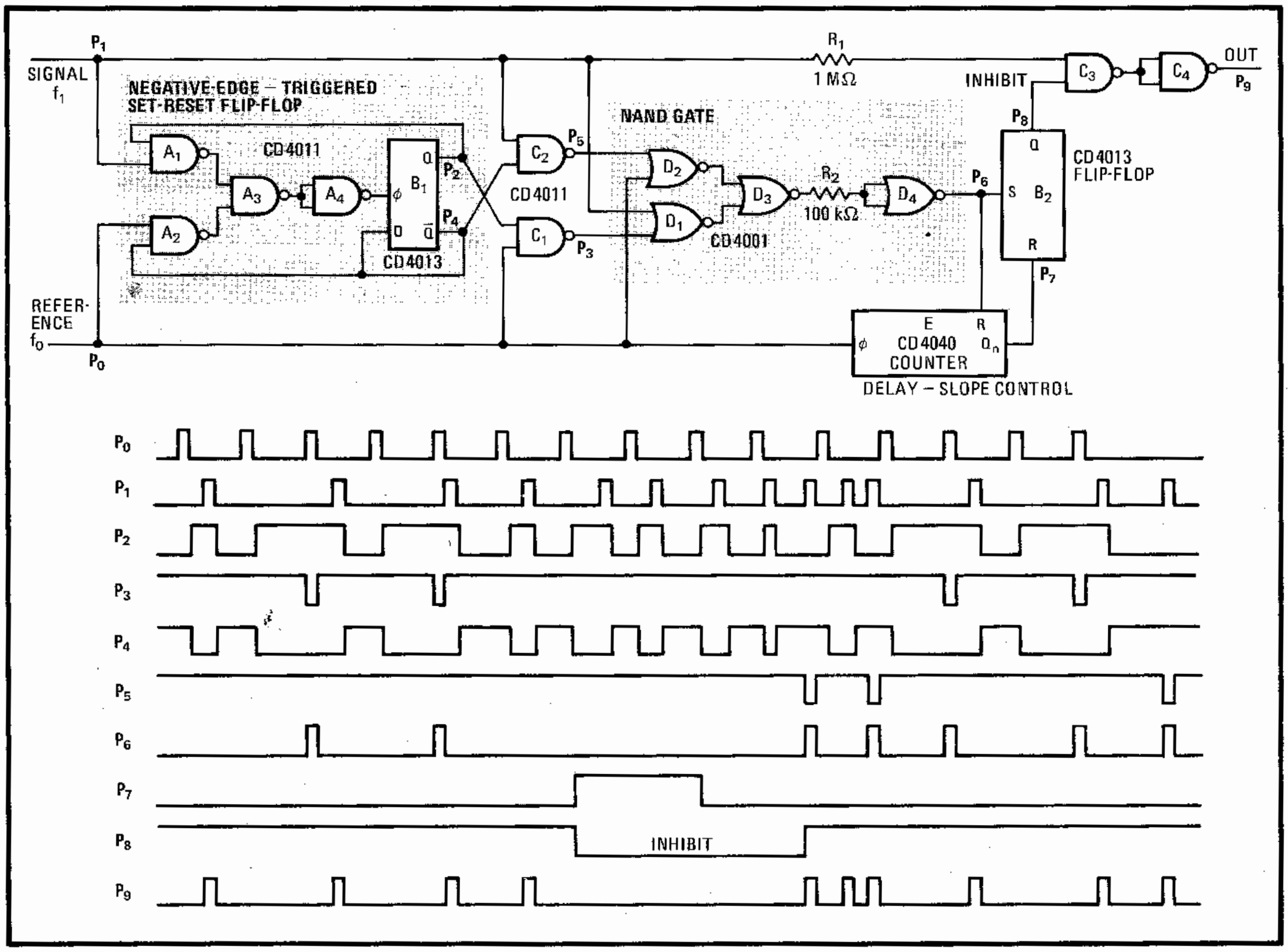
Performing the digital counterpart of electric-wave filtering in the analog domain, this unit can function as a low-pass, high-pass, bandpass, or band-reject filter of a square-wave pulse train. No RC integrating networks or comparators are needed, the all-digital filter being tuned simply by adjusting the reference frequency. Built originally for biomedical applications, it can find much broader use in the field of communications.

Consider the case of the band-reject filter shown in (a) in the figure. As seen with the aid of the timing diagram, the Q output of the edge-triggered set-reset flip-flop formed by the gates of the CD4011, A_1–A_4, and the 4013 D flip-flop (B_1) is brought high by the training edge of reference frequency f_0 and brought low by the falling edge of signal f_1. The combined output of the flip-flop and f_0 appears at gate C_1, moving low if $f_0 > f_1$. Similarly, C_2 moves low if $f_1 > f_0$. Therefore, signals from the output of the NAND gate formed by NOR gates D_1–D_4 appear whenever $f_0 \neq f_1$. Each pulse sets flip-flop B_2 high if it is not already so, permitting signal f_1 to pass through to the output.

Meanwhile, the 12-stage 4040 ripple counter advances on each pulse from f_0. The counter will reach the Q_n state if $f_0 = f_1$, because no reset pulse can emanate from gate D_4 under that condition. These events will disable gate C_3 and prevent f_1 from reaching the output.

The steepness of the filter's roll-off characteristic will be determined by which stage of the counter resets flip-flop B_2. The filter's reaction time to changes in the input and reference frequencies will vary accordingly—that is, the steeper the slope, the longer the response time, this delay being the major drawback of the filter. The corner frequencies are $f_{1\,min} = (N-1)f_0/N$ and $f_{1\,max} = (N+1)f_0/N$, where N is the number of pulses of f_0 required for the counter to produce a reset pulse. The quality factor is $Q = f_0/\Delta f_1 = N/2$.

The stop-band filter can be easily modified to a bandpass type if the $\overline{Q}$ output of flip-flop B_2 is wired to serve as the inhibit line. To convert the filter for low-pass duties C_2 is removed and both inputs of D_3 are connected to D_1. In like fashion, C_1 is removed and both inputs of D_3 are connected to D_2 if a high-pass response is desired. Note that the NOR-gate circuitry is required to avoid any ambiguity of output state when pulses of input and reference signals overlap. Also, resistors R_1 and R_2 neutralize the effect of variable propagation-time differences of f_0 and f_1 through the gates □



Digital damping. To perform band-reject function, this combinational logic circuit ascertains the frequency relationship of two square-wave signals. Tuning is done by adjusting the reference frequency. Selectivity is determined by tap position Q_n of the counter. Waveforms for given points in circuit show timing relationships. With minor changes, the filter is easily adapted for high-pass, low-pass, and bandpass duties.

Designer's Notebook
Switched Capacitor Filters

Switched capacitor filters might appear to be unsuitable devices for anyone who isn't an expert. In Part One, Tim Orr looks at them and finds that they are really much easier to use than conventional filters.

MOST COMPLEX FILTER designs require a large number of precision resistors, inductors and capacitors. Resistors and capacitors are relatively easy to integrate, and inductors can be synthesized using an op amp plus capacitors and resistors. Thus it would seem possible to produce a monolithic active filter, but there is one major problem. The accuracy of monolithic capacitor values is typically fairly low, and constructing multi-pole filters requires good matching between stages: a sixth order low-pass filter would typically require a component tolerance of 1 or 2%. Additionally, monolithic capacitors and resistors are limited to fairly low values, though this could be designed around.

Switched capacitor techniques get over these problems by using a capacitance to synthesize a resistance; see Fig. 1. The resistance is synthesized by switching charge into and out of the capacitor C_R, as the MOSFET switches open and close in antiphase; the average current passing from the input to the op amp (and hence the apparent resistance) depends on the switching frequency. This helps because the *ratio* of the values of two monolithic capacitors on the same IC can be accurately controlled (to 1% or better). In the simple example of a low-pass filter shown in Fig. 1, the break frequency is proportional to the switching frequency and the ratio of the capacitor values. Switched capacitor filters are normally designed this way (even the very complicated ones). The switching frequency is usually arranged to be around either 50 or 100 times the break frequency of the filter, and therefore only very simple anti-aliasing and recovery filtering are required.

Several manufacturers produce switched capacitor filter ICs and all the classic filter structures are available. These devices offer several advantages over conventional passive and active filters:

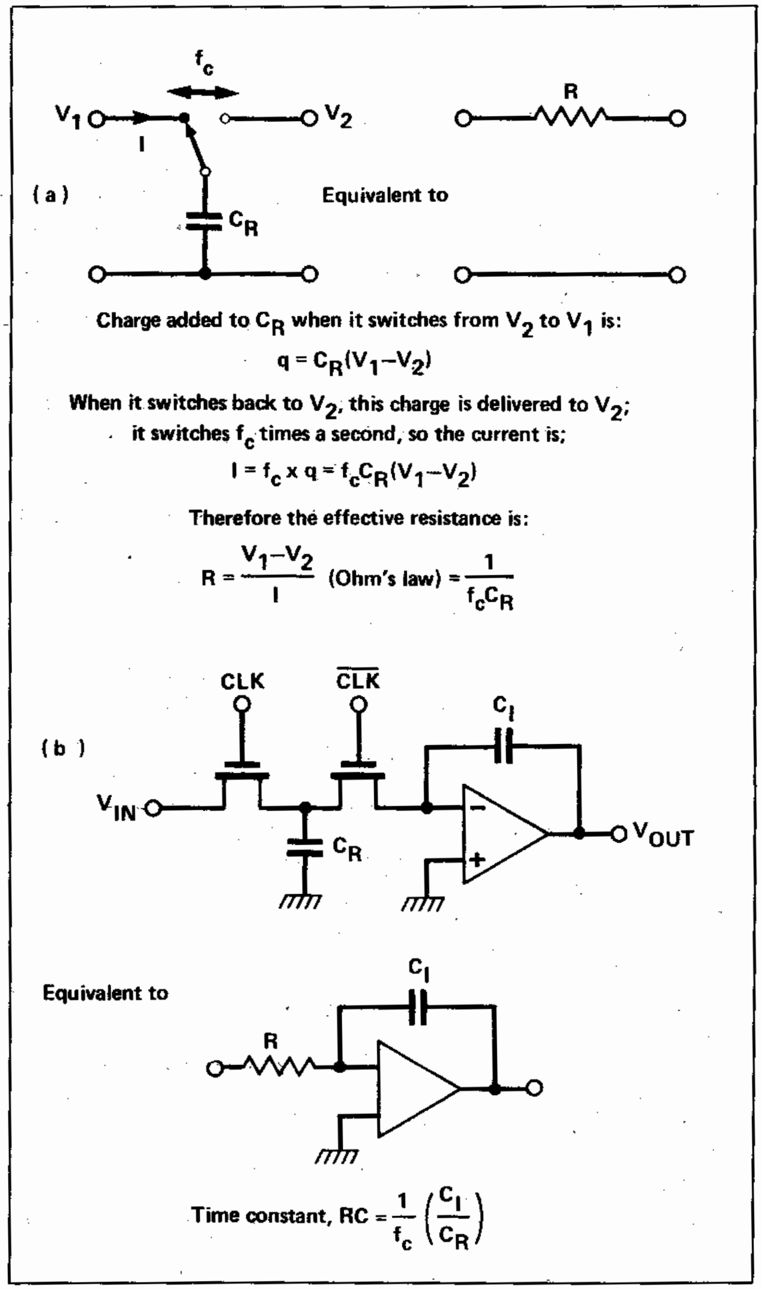


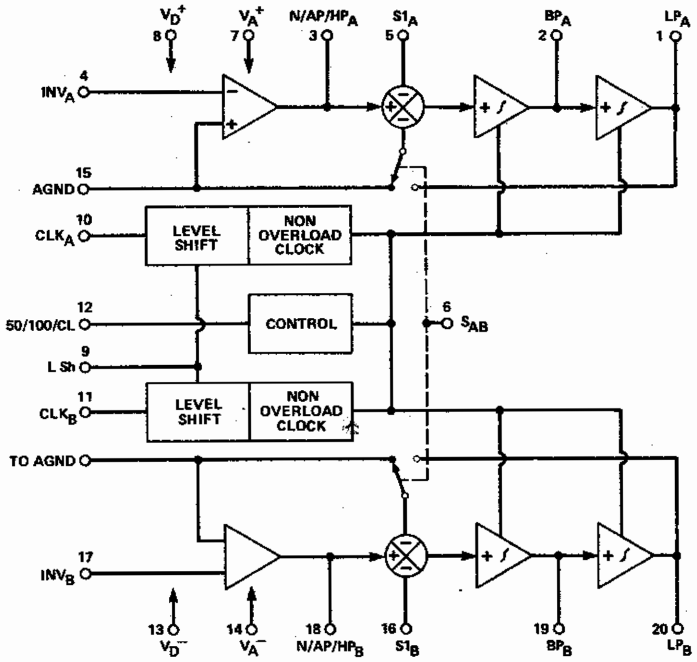
Fig. 1 Basics of switched capacitor filters. Synthesising a resistance (a); a simple low-pass filter using a synthesised resistance (b).

- filters can be made very compact
- very few external components are needed
- the filters are tunable by adjusting the externally generated sampling frequency — so no re-alignment is necessary when break-frequencies are changed
- circuit calculations are made very easy

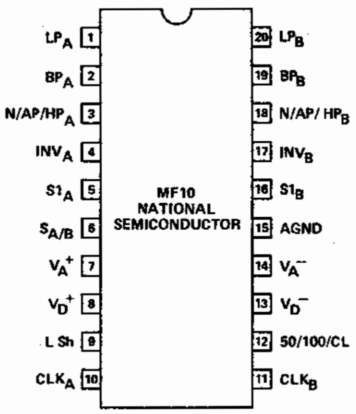
The following section is a review of some of the currently available devices. Next month, we'll look at a few examples of practical filter circuits using switched capacitor ICs.

MF10 — National Semiconductor

The MF10 is a dual independent active filter building block. Virtually any classic filter structure can be fabricated with this device.



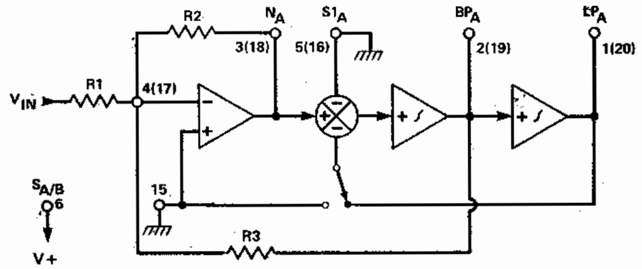
DUAL-IN-LINE PACKAGE



TOP VIEW

SUPPLY VOLTAGE ± 5V

LP, BP, N/AP/HP	low-pass, band-pass, notch, all-pass and high-pass outputs respectively. All can sink 1 mA and source 3 mA; N/AP/HP can sink 1.5 mA
V_A+, V_D+	positive analogue and digital supply rails. These are linked internally, and so must have the same supply voltage applied to them, normally +5 V
V_A-, V_D-	negative supply rails, also internally connected; normally −5 V
AGND	analogue ground, which should be at 0 V, ie mid way between positive and negative supply rails
L Sh	level shift for clock inputs. For TTL input, tie to 0 V; for CMOS operated from 10 V, tie to negative supply rail
CLK A or B	clock inputs for each filter unit
50/100/CL	defines relationship between clock frequency and filter centre frequency; tie to positive supply for 50:1 ratio, tie to 0 V for 100:1
$S_{A/B}$	activates internal switches, see section on use; note that there is only one $S_{A/B}$ for both filters



MODE 1

MODE 1: Notch 1, Band-pass, Low-pass outputs: $f_{notch} = f_o$

f_o = centre frequency of the complex pole pair

$= \frac{f_{CLK}}{100}$ or $\frac{f_{CLK}}{50}$

f_{notch} = centre frequency of the imaginary zero pair = f_o

H_{OLP} = Low-pass gain (as $f \to 0$) = $\frac{R2}{R1}$

H_{OBP} = Band-pass gain (at $f = f_o$) = $-\frac{R3}{R1}$

H_{ON} = Notch output gain as $\left\{ \begin{matrix} f \to 0 \\ f \to f_{CLK}/2 \end{matrix} \right\} = -\frac{R2}{R1}$

$Q = \frac{f_o}{BW} = \frac{R3}{R2}$

= quality factor of the complex pole pair.

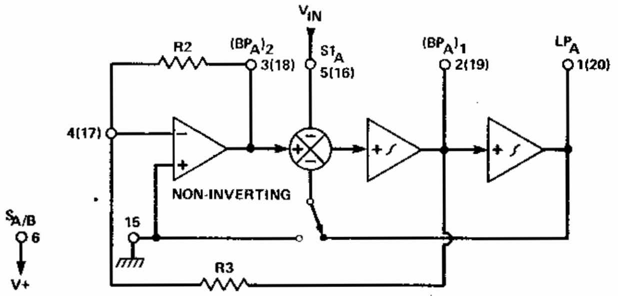
BW = the −3dB bandwidth of the band-pass output.

Circuit dynamics:

$H_{OLP} = \frac{H_{OBP}}{Q}$ or $H_{OBP} = H_{OLP} \times Q = H_{ON} \times Q$.

$H_{OLP(peak)} = Q \times H_{OLP}$ (for high Q's)

The above expressions are important. They determine the swing at each output funtion of the desired Q of the 2nd order funtion.



MODE 1a

MODE 1a: Non-inverting BP, LP.

$f_o = \frac{f_{CLK}}{100}$ or $\frac{f_{CLK}}{50}$

$Q = \frac{R3}{R2}$

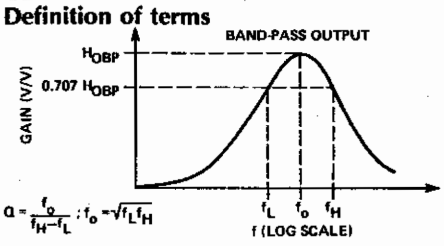
$H_{OLP} = 1$; $H_{OLP(peak)} \cong Q \times H_{OLP}$ (for high Q's)

$H_{OBP_1} = -\frac{R3}{R2}$

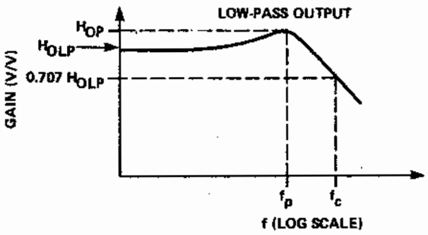
$H_{OBP_2} = 1$ (non-inverting)

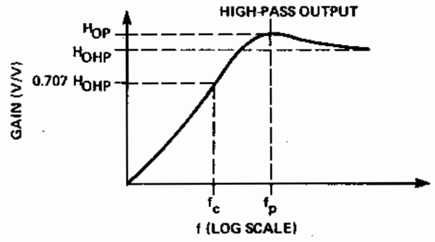
Circuit dynamics: $H_{OBP_1} = Q$

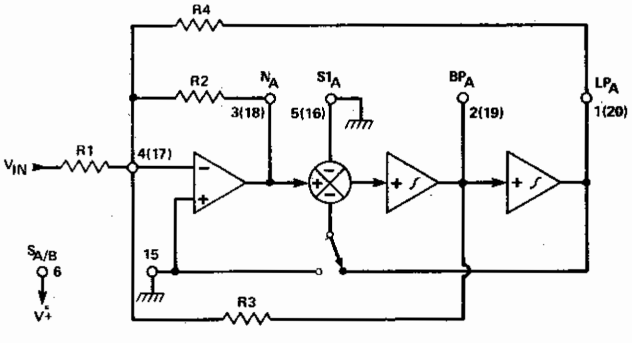
Definition of terms



$Q = \frac{f_o}{f_H - f_L}$; $f_o = \sqrt{f_L f_H}$







MODE 2

MODE 2: Notch 2, Band-pass, Low-pass: $f_{notch} < f_o$

f_o = centre frequency

$$= \frac{f_{CLK}}{100}\sqrt{\frac{R2}{R4}+1} \text{ or } \frac{f_{CLK}}{50}\sqrt{\frac{R2}{R4}+1}$$

$$f_{notch} = \frac{f_{CLK}}{100} \text{ or } \frac{f_{CLK}}{50}$$

Q = quality factor of the complex pole pair

$$= \frac{\sqrt{R2/R4+1}}{R2/R3}$$

H_{OLP} = Low-pass output gain (as $f \to 0$)

$$= -\frac{R2/R1}{R2/R4+1}$$

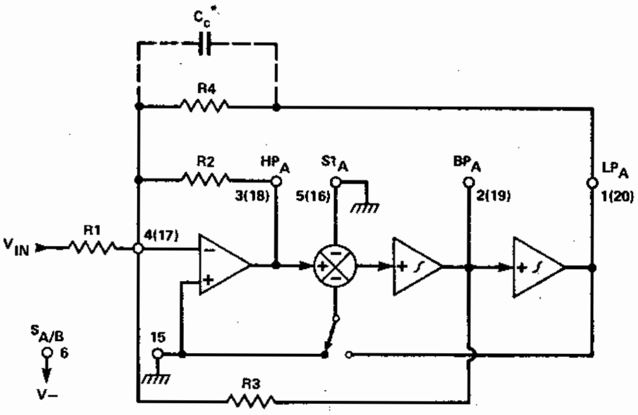
H_{OBP} = Band-pass output gain (at $f = f_o$) = $-R3/R1$

H_{ON_1} = Notch ouput gain (as $f \to 0$)

$$= -\frac{R2/R1}{R2/R4+1}$$

H_{ON_2} = Notch output gain $\left(\text{as } f \to \frac{f_{CLK}}{2}\right) = -R2/R1$

Filter dynamics: $H_{OBP} = Q\sqrt{H_{OLP}H_{ON_2}} = Q\sqrt{H_{ON_1}H_{ON_2}}$



MODE 3

* In Mode 3, the feedback loop is closed around the input summing amplifier; the finite GBW product of this op amp causes a slight Q enhancement. If this is a problem, connect a small capacitor (10pF-100pF) across R4 to provide some lead.

MODE 3: High-pass, Band-pass, Low-pass outputs

$$f_o = \frac{f_{CLK}}{100} \times \sqrt{\frac{R2}{R4}} \text{ or } \frac{f_{CLK}}{50} \times \sqrt{\frac{R2}{R4}}$$

Q = quality factor of the complex pole pair

$$= \sqrt{\frac{R2}{R4}} \times \frac{R3}{R2}$$

H_{OHP} = High-pass gain $\left(\text{as } f \to \frac{f_{CLK}}{2}\right) = -\frac{R2}{R1}$

H_{OBP} = Band-pass gain (at $f = f_o$) = $-\frac{R3}{R1}$

H_{OLP} = Low-pass gain (as $f \to 0$) = $-\frac{R4}{R1}$

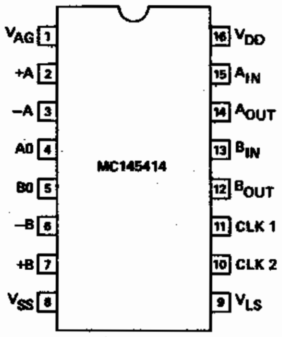
Circuit dynamics: $\frac{R2}{R4} = \frac{H_{OHP}}{H_{OLP}}$; $H_{OBP} = \sqrt{H_{OHP} \times H_{OLP}} \times Q$

$H_{OLP(peak)} = Q \times H_{OLP}$ (for high Q's)

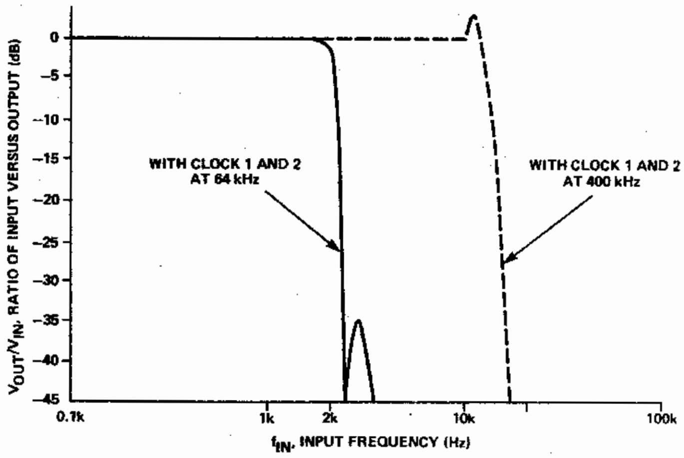
$H_{OHP(peak)} = Q \times H_{OHP}$ (for high Q's)

MC145414 — Motorola

This is a dual low-pass filter which can operate with a break frequency between 1.25 and 10 kHz. It also contains two completely uncommitted op amps in the same package. Filters are fifth-order elliptic, and have a clock-to-break-frequency ratio of approximately 36:1 (clock frequency must be between 50 and 400 kHz). Filter A has 18 dB gain in the pass band, while filter B has unity gain. The clock input voltage may be selected to be 5 V or 12 V, or the whole IC may be powered down by the application of suitable supply voltages.



V_{AG}	analogue ground; all analogue signals are referred to this level, and it should normally be around mid-way between positive and negative supply rails. If taken to within 1 V of positive rail, IC will power down	V_{LS}	logic shift voltage; for TTL clock input, take to mid-way between supply rails; for CMOS operating on to 12 V, take to negative supply rail. To power down IC, take to positive supply rail
+A, −A, AO	non-inverting, inverting inputs, and output of op-amp A	CLOCK 1, 2	clock inputs — should always be tied together
+B, −B, BO	as above for op-amp B	A_{IN}, B_{IN}	inputs to filters A, B
V_{DD}, V_{SS}	positive and negative supply rails. V_{SS} is also ground for digital inputs	A_{OUT}, B_{OUT}	outputs from filters A, B



R5620 — Reticon (not illustrated)

This device is a switched capacitor universal active filter, with digital setting of both the Q-factor and the filter centre frequency. It is a second-order filter capable of high-pass, low-pass, band-pass, notch and all-pass. The filter frequency is determined by the clock frequency and a five-bit binary input that moves the frequency over a two-octave spread in 32 logarithmically spaced intervals. This enables direct digital control of the centre frequency. The Q-factor is similarly controlled with a five-bit code. The Q range is from 0.57 to 150, and the break frequency range is 0.5 Hz to 25 kHz.

R5604, R5605, R5607 — Reticon

These are octave filters. All contain six-pole Chebyshev filters; however, they contain different numbers of filters with different pass band widths. The R5604 contains three ⅓ octave ANSI Class III filters that together cover an entire octave; the R5605 contains two ½ octave filters that together cover an octave, and the R5607 contains one full-octave ANSI Class II filter. The centre frequency of all the filters is controlled by the single clock input — see the response curves for details. The dynamic range is better than 80 dB and distortion is less than 0.1%. The filters can handle input signals greater than 10 V peak-to-peak and have an insertion loss of less than 0.2 dB in the pass band.

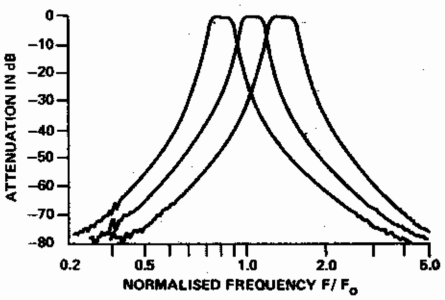
PIN	R5604 FUNCTION	R5605 FUNCTION	R5606 FUNCTION
1	V–	V–	V–
2	N/C	N/C	N/C
3	IN 2	IN 1	N/C
4	IN 1	N/C	N/C
5	V+	V+	V+
6	N/C	N/C	N/C
7	TRIG IN	TRIG IN	TRIG IN
8	N/C	N/C	IN 1
9	IN 3	IN 2	V–
10	V–	V–	N/C
11	N/C	N/C	N/C
12	V+	V+	V+
13	COM	COM	COM
14	OUT 3	OUT 2	N/C
15	OUT 2	N/C	N/C
16	OUT 1	OUT 1	OUT 1

(a) R5604

R5604 3-1/3 OCTAVE BAND-PASS FILTERS

f_{SAMPLE} = 55 kHz

F_o = 1 kHz

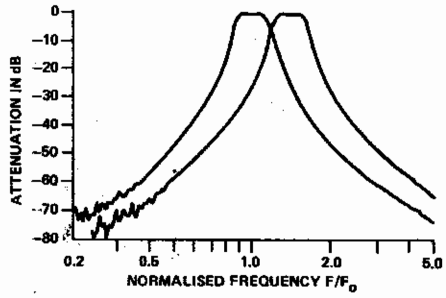


(b) R5605

R5605 2-½ OCTAVE BAND-PASS FILTERS

f_{SAMPLE} = 55 kHz

F_o = 1 kHz

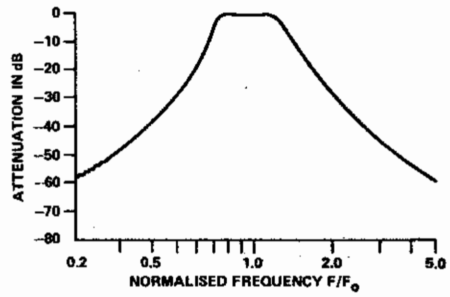


(c) R5606

R5606 1 FULL OCTAVE BAND-PASS FILTERS

f_{SAMPLE} = 55 kHz

F_o = 1 kHz



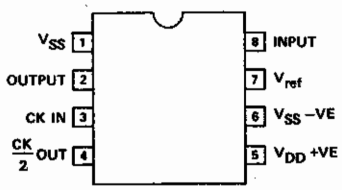
R5609, R5611, R5612 — Reticon

The R5609 is a seven-pole, six-zero elliptic low-pass filter with over 75 dB out-of-band rejection and less than 0.2 dB of pass band ripple.

The R5611 is a five-pole Chebyshev high-pass filter with 30 dB per octave rolloff and less than 0.6 dB of pass band ripple.

The R5612 is a four-pole notch filter with over 50 dB of rejection at the notch frequency.

The corner/center frequencies of these switched capacitor filters is tuneable by the input trigger frequency over a wide frequency range from 0.1 Hz up to 25 kHz. The dynamic range is better than 75 dB and distortion is less than 0.3%. Signal handling capability is over 12 V peak-to-peak, and typical insertion loss is 0 dB. Supply voltages may be ±4 V to ±11 V maximum.



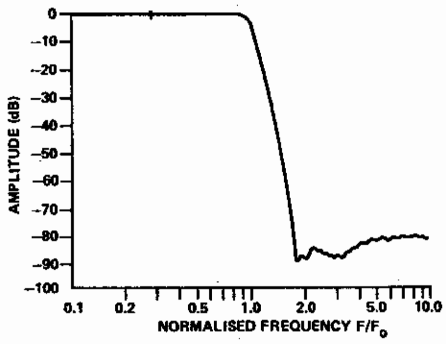
PIN OUT
(ALL 3 DEVICES)

SUPPLY VOLTAGE ± 4V TO ± 11V MAX.

(a) R5609

FREQUENCY RESPONSE OF R5609 LOW-PASS FILTER

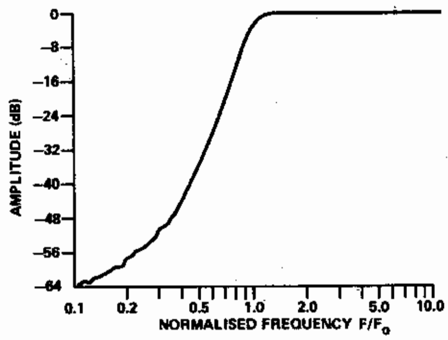
NOTE: f_c/F_o = 100 (typical)



(b) R5611

FREQUENCY RESPONSE OF R5611 HIGH-PASS FILTER

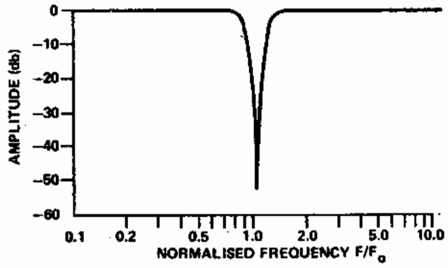
NOTE: f_c/F_o = 500 (typical)



(c) R5612

FREQUENCY RESPONSE OF R5612 NOTCH FILTER

NOTE: f_c/F_o = 930 (typical)



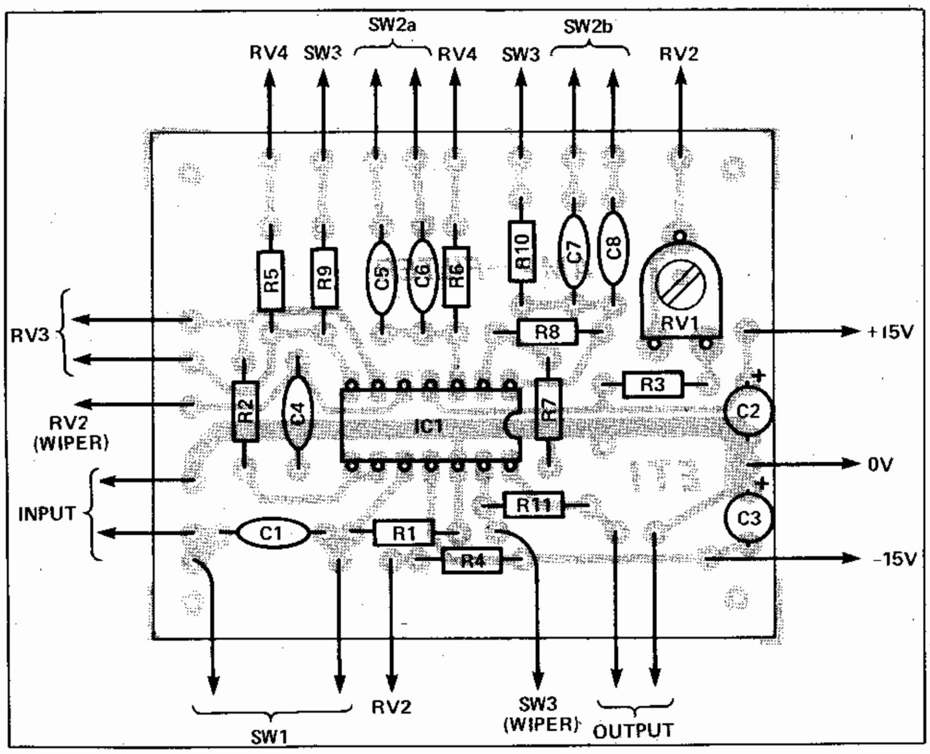
DAC/ADC Filter-Amplifier

PROJECT

ETI supplies the missing link for all your analogue to computer interface circuits. Design by C.D. Oddy.

ANYONE WHO HAS ever seriously experimented with analogue interfaces to home computers must be all too aware of the problems involved — if you're not too careful you wind up with op-amps everywhere in a variety of hastily concocted amplifier, buffer, and filter circuits. And some of them won't even be hastily concocted; some of them will have taken up a considerable amount of your precious time in the designing. What you need is a handy little unit that will perform any or all of the functins of amplifier, buffer, and filter at the mere twiddle of a control or two. What you need is the ETI filter-amplifier.

The filter-amplifier consists of two active blocks together with input and output buffers and switchable AC or DC coupling. The two active blocks are an amplifier with a variable gain of 0 to 100 and a plus or minus five volt variable offset followed by a low pass filter whose cutoff frequency may be varied over the range 16 Hz to 30 kHz and which may be switched out of circuit when not required. It can be used as an amplifier to match low level signals to the input of an ADC, as a buffer to correct mismatching of signal sources, as a filter to smooth the stepped output of a DAC, and for a multitude of other similar purposes.



Component overlay for the PCB.

Construction

Assembly is simplified by the use of a single quad op-amp for the four stages, and the only point to check on the PCB itself is that this IC and the two electrolytic capacitors (C2, C3) are inserted the right way around. Take care with the wiring of the potentiometers and switches, especially SW2 which has two separate elements whose wiring

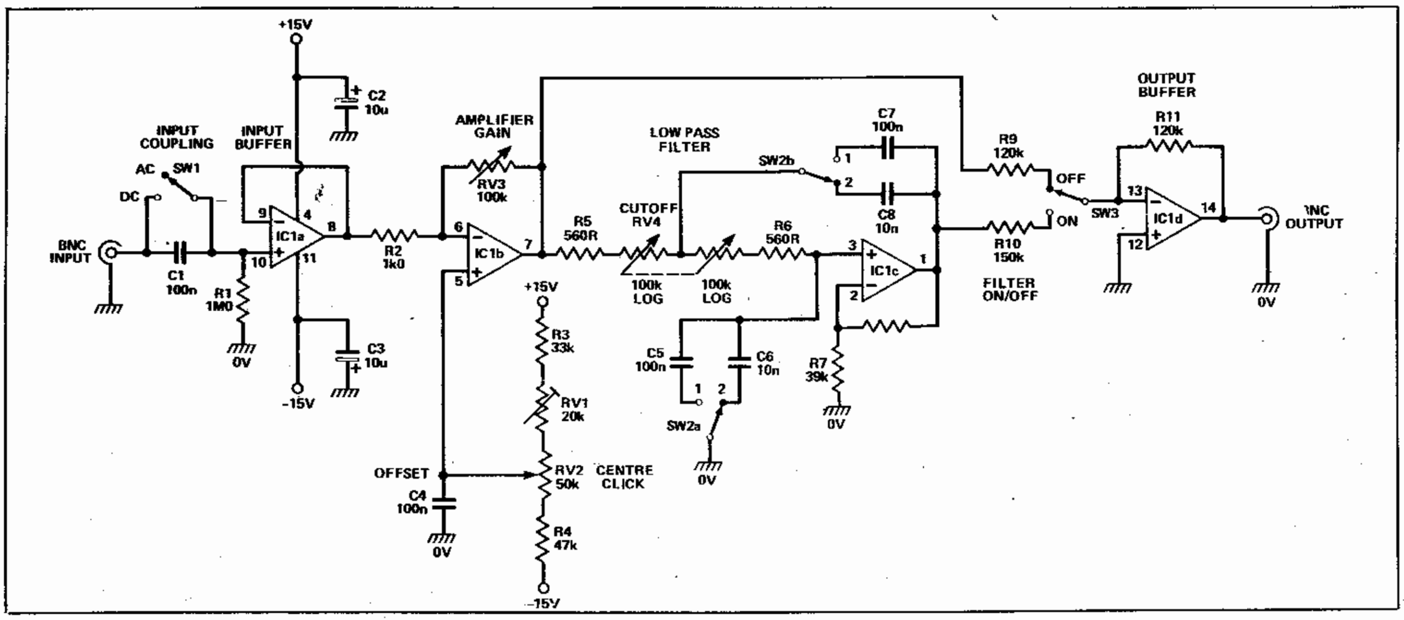


Fig. 1 Circuit diagram of the filter-amplifier.

must be in agreement if the filter is to work correctly. Note also the wiring around RV4 and the connection between this potentiometer and SW2, as shown in **Fig. 2.**

We have not shown either a case or a power supply for this project because it is intended more as a building block than as a complete, self-contained unit. For the really serious experimenter there is obviously a lot to be gained from having several of these filter-amplifiers built into a single unit and equipped with a common power supply, whereas for others, one would probably be sufficient and could be powered from whatever you're using to supply the ICs you're experimenting with.

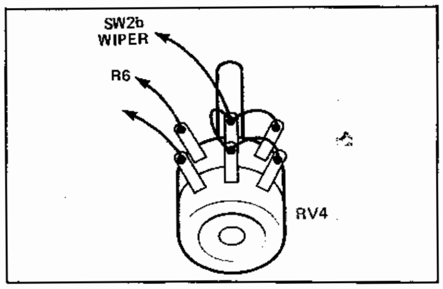
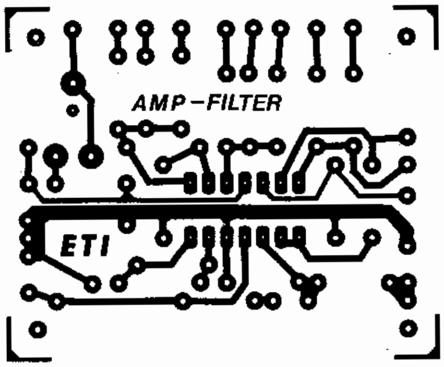


Fig. 2 (below) Connections to RV4



PCB for the DAC/ADC filter amplifier.

HOW IT WORKS

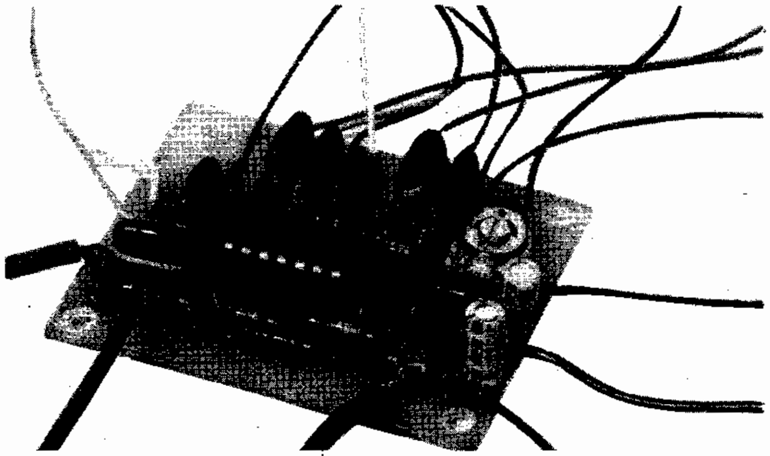
The input signal is passed via either C1 or SW1 to IC1a, the first stage of the TL074 quad BIFET op-amp, which acts as a buffer. The signal is then amplified by IC1b, the gain being set by R2 and RV3. RV1 and RV2 set the offset voltage, RV1 being preset such that the offset is zero when RV2 is in its centre click-stop position.

IC1c and its associated circuitry form a low pass filter of the second order Bessel type which is well suited to the smoothing of stepped signals, such as those obtained from DAC outputs. The filter's cut-off frequency may be adjusted using RV4 and the range switch SW2, the latter selecting either 16Hz to 3 kHz (position 1) or 160 Hz to 30 kHz (position 2). The gain of the stage is set by R7 and R8.

SW3 selects either the amplifier or amplifier plus filter outputs via R9 and R10, the slight difference in their values being necessary in order to allow for the gain (approx. 2.3 db) of the filter. The signal then passes through IC1d, the final buffer stage, to the output.

PARTS LIST

Resistors (all ¼W, 5%)		RV4	100k dual gang log
R1	1M0	**Capacitors**	
R2	1k0	C1, 4, 5, 7	100n polyester
R3	33k	C2, 3	10u 35V tantalum
R4	47k	C6, 8	10n polyester
R5, 6	560R		
R7	39k	**Semiconductor**	
R8	10k	IC1	TL074
R9, 11	120k	**Miscellaneous**	
R10	150k	SW1	SPST
RV1	20k horizontal trimpot	SW2	DPDT
RV2	50k linear with centre click-stop	SW3	SPDT
RV3	100k linear	PCB; case, input and output sockets, knobs, etc. as desired.	



ETI

NORMAL FILTER type musical effects units (phase and waa-waa units for example) use a low frequency oscillator to provide automatic sweeping of the filter frequency. Another approach is to use the envelope of the input signal or the control voltage from an envelope shaper (as in a synthesizer) to provide the sweeping. A third method is to simply use a foot pedal to control the effect. This novel effects unit uses yet another approach, and this is to have a computer to control the filter frequency. Although only a relatively crude method of control is possible, this enables a useful range of effects to be obtained, and the versatility is increased by having two filters. Notch, bandpass, and lowpass filtering are available, with each filter having a Q or resonance control. This enables simple phasing plus two forms of waa-waa effect to be produced.

While perhaps not being ideal as a guitar effects unit, this project is useful to anyone interested in sound synthesis, particularly in stereo set-ups. For stereo operation the main way of using the unit is to feed one signal to both filter inputs, and then take a pseudo stereo output from the two filter outputs. This would be pointless if the filters were to be swept up and down in frequency in unison, but some interesting effects can be obtained if the filters are operated in antiphase, with the frequency of one being swept downwards while the other is swept upwards. The effective sweep waveform can be altered under software control, and can be triangular, sawtooth (of either type), squarewave, or even a random modulation of the filters can be used if desired. The two filters do not need to be set for the same type of filtering, and interesting results can be produced using bandpass filtering in one channel and notch filtering in the other. Each filter has a separate input, and the unit can be fed from a stereo source if desired. Obviously there are numerous ways in which the unit can be used, and it is ideal for the more adventurous electronic music enthusiast.

The unit is primarily intended for use with the Commodore 64 computer, but it will work in a more basic manner with the VIC-20 or any computer which has a user port provided by port B of a 6526 or 6522 device. With machine code software it should be possible to drive it from any computer that can provide one or two digital outputs, but this goes beyond the scope of this article.

Computer Controlled Filter

There are various ways of controlling the filter frequency of a musical effect. This one uses a Commodore 64.

By R.A. Penfold

Operating Principle

One approach to a computer controlled filter would be to have an ordinary voltage controlled filter driven from the computer via a digital to analogue converter. A more simple alternative, and the one adopted for this design, is to use a switched capacitor filter with the computer providing the clock signal. Switched capacitor filters may be unfamiliar to many readers, and while not being a particularly new concept, have received relatively little use.

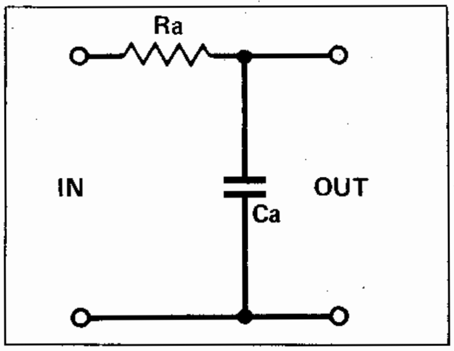


Fig. 1a A simple low pass filter circuit.

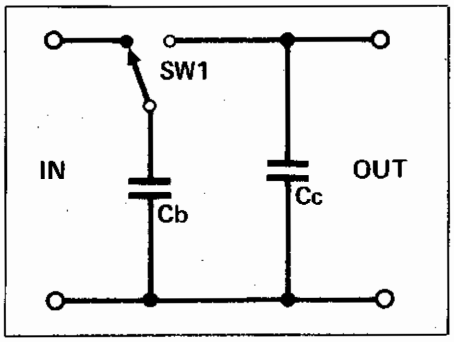


Fig. 1b The circuit for a switched capacitor filter.

An ordinary single stage lowpass filter just consists of a resistor and a capacitor connected as shown in **Fig. 1(a)**. Ra and Ca form a potential divider circuit, but the degree of attenuation through the circuit is not fixed, and depends on the impedance of Ca. This decreases as the input frequency is increased, giving losses that rise as the input frequency is increased. The losses are minimal at first, but once the 6dB point has been reached (the output potential is half the input voltage) the attenuation rate is 6dB per octave (doubling the input frequency halves the circuit gain).

A switched capacitor filter operates in a similar fashion, but the resistor is replaced with a low value capacitor and an electronic (SPDT) switch, as in **Fig. 1(b)**. Cb and SW1 are the capacitor and the switch, while Cc is the normal filter capacitor. A clock oscillator controls the rate at which the electronic switch alternates between the position where it takes a charge from the input, and is discharged into Cc. A change in the potential at the input of the circuit is obviously transferred to the output of the circuit by the charge/discharge action of Cb and SW1, but changes in potential are not instantly transferred. The time taken for the output of the circuit to respond to changes at the input depends on the relative size of Cb and Cc, and on the rate at which SW1 is clocked. It is this last point which is of importance, since raising the clock frequency provides a faster transfer, and boots the cutoff frequency of the filter. In other words, the filter's cutoff frequency is proportional to the clock frequency.

In a practical switched capacitor filter there are usually two filters wired in series, and some other circuits are included so that the basic lowpass filter action can be modified to give any other form of filtering. The switched capacitor device used in this project is the National Semiconductor MF10CN which can operate in any filter mode, has adjustable Q, and two 12dB per octave filters. The filter frequency can be either 1/50th or 1/100th of the clock frequency but it must be the same for both filters.

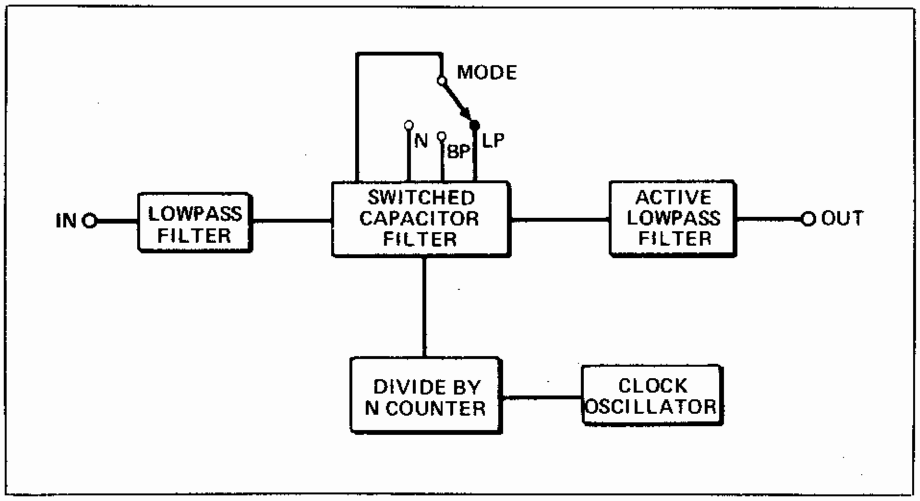


Fig. 2 The block diagram for one channel of the filter; the other channel is identical.

Circuit Operation

The block diagram of Figure 2 shows the make-up of one channel of the unit (the other channel is identical). A passive lowpass filter is used ahead of the switched capacitor filter, and this is needed to prevent strong signals close to the clock frequency from entering the switched capacitor filter and producing heterodyne 'whistles'. The switched-capacitor filter operates in a mode which gives notch, bandpass, and lowpass filtering, and a switch is used to select the desired mode. Clock glitches only appear at the output of the filter at a relatively low level, but an active filter here gives further attenuation and a clean output signal. This stage also provides output buffering.

The clock signal is provided by the system clock of the computer, and is at a frequency of about 1MHz, but the precise frequency depends on the particular computer used (980kHz in the case of the Commodore 64). It is connected to the filter via a divide by N counter, but this is also part of the computer and not included in this project. The divider can be pro-

Continued on page 42

Continued from page 27

grammed to divide by any even integer from 2 to 131070, and a vast range of some 65535 output frequencies is therefore available. However, most of these are unusable in this case as they are within the audio range, and would produce noticeable breakthrough at the output. The available range enables some useful effects to be generated though.

The full circuit diagram of the unit appears in **Fig. 3**. If we take the channel based on IC1a, R1 and C2 are the input filter, while IC2 operates as a standard third order 18dB per octave lowpass filter at the output. SW1a selects the required type of filtering, and on the prototype this is ganged with SW1b in the other channel. However, separate switches should be used here if you wish to use the two channels in different filter modes. RV1 is the 'Q' control. The filters are connected to operate with a 100 to 1 clock/cutoff frequency ratio, which gives an effective sweep range of about 150Hz to 5kHz.

The circuit requires dual balanced 5 volt supplies. The +5 volt supply is available rom the Commodore 64's user port, but the −5 volt supply has to be derived from one of the 9 volt AC outputs using a simple rectifier, smoothing, and regulator circuit based on IC4.

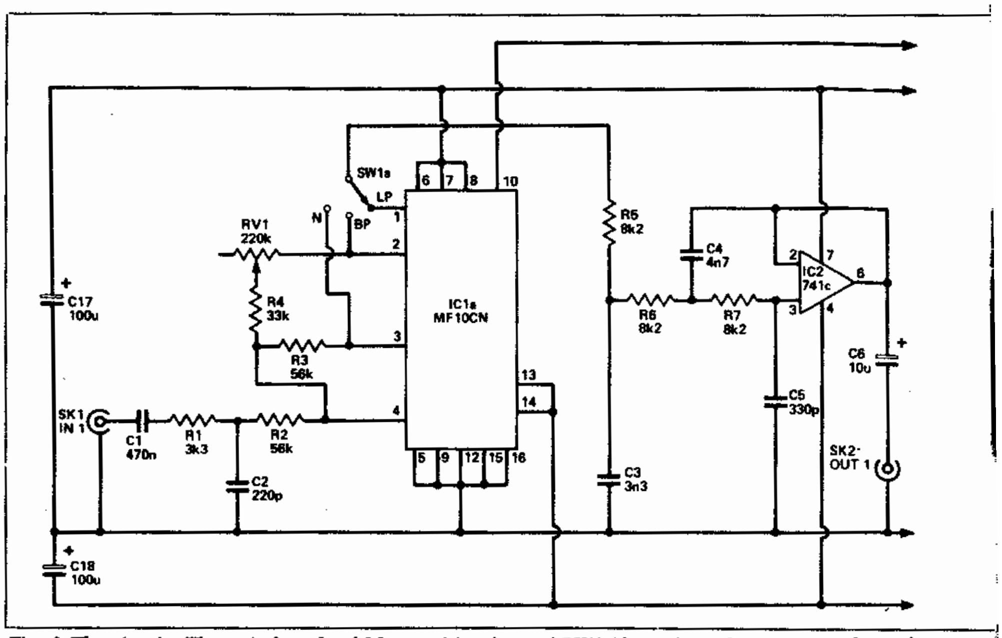


Fig. 3 The circuit. The switches should be used in place of SW2 if you intend to use two channels in different filter modes.

Construction

Details of the printed circuit board and wiring are shown in **Fig. 4**. There is nothing unusual about construction of the board, but note that IC1 is a MOS device and should be mounted in a socket. The other MOS handling precautions should also be observed. The connections to the computer are taken via a 5 way DIN socket fitted on the rear panel of the case. From here the connections are carried to the computer by way of a 5 way ribbon cable up to about a metre long and fitted with a 5 way DIN plug to connect to SK5, and a 2 by 12 way edge connector at the end which connects to the user port of the computer. **Fig. 5** gives wiring details for the edge connector.

The case for the prototype is a plastic Verocase having metal front and rear panels, with approximate outside dimensions of 180 by 120 by 39 millimetres. Any case of roughly the same size should take all the parts without any difficulty though. The controls and input/output sockets are mounted on the front panel. I used a stereo jack socket at the inputs, and another at the outputs, but any type of audio connector that is convenient for your particular setup can be used.

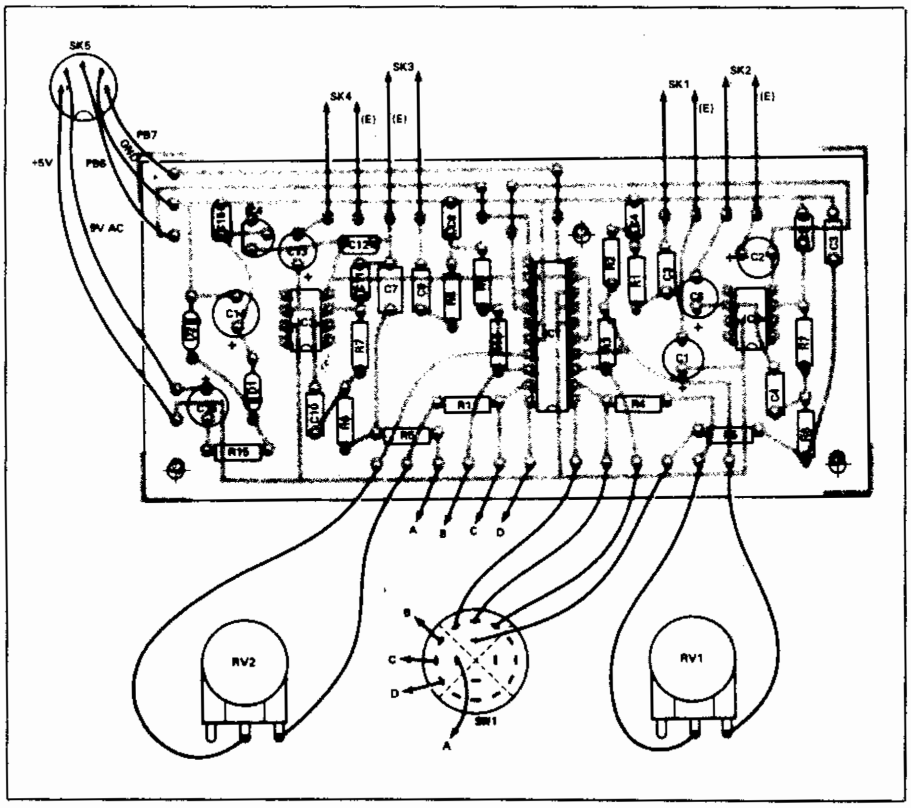
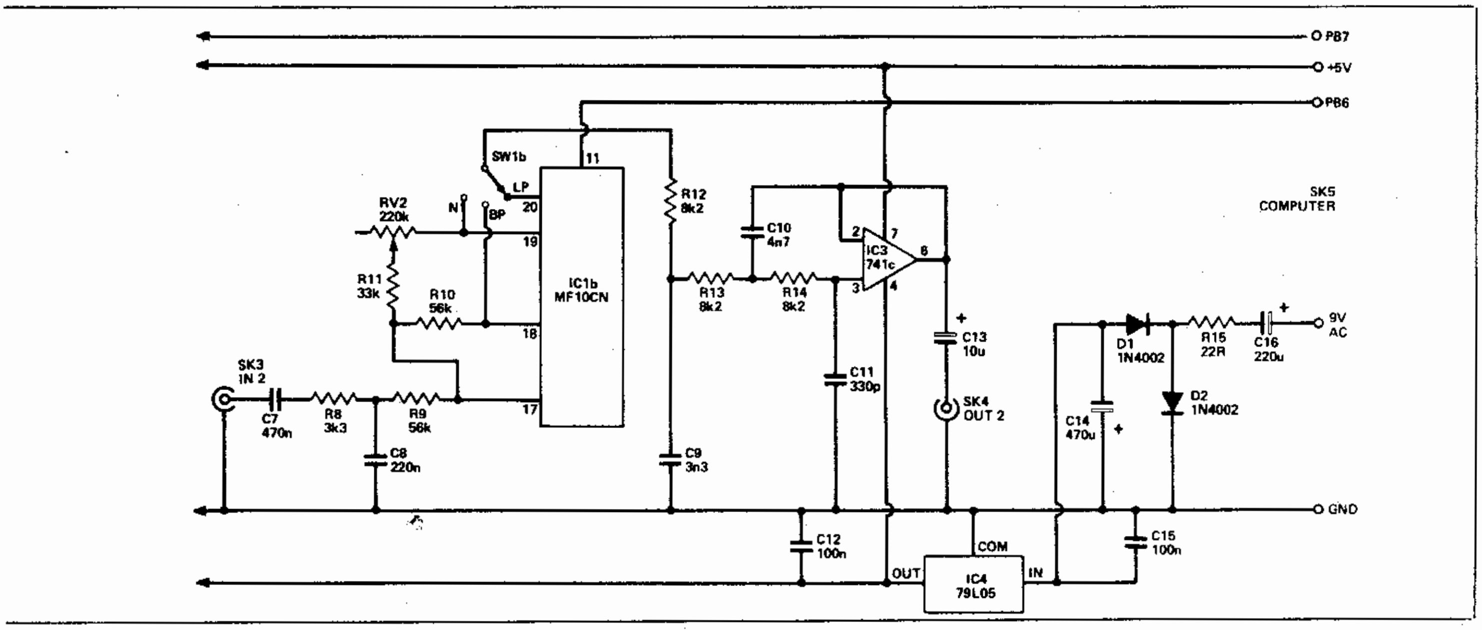


Fig. 4 The component layout: straightforward, but use an IC soket for IC1.

In Use

It is advisable to plug the unit into the computer prior to switching on the latter. The filters have a medium input impedance, low output impedance, and nominally unity voltage gain, and they should therefore fit into most set-ups with no problems. Signals of up to about 2 volts RMS can be handled.

In order to set the timer/counters of the computer's interface chip to the correct operating mode a value of 23 must be written to the two control registers. These are at addresses 56590 and 56591. The high bytes of the two timer/counters are at addresses 56581 and 56583, and in this application these are set to and simply left at zero. The filter frequencies are then controlled by the values written to the low bytes at addresses 56580 and 56582. These control the filters fed from PB6 and PB7 respectively. Values from 1 to about 25 can be used, and although this gives rather crude control of the filter frequency, it nevertheless produces some good effects. This simple program demonstrates the basic way in which the filter is used.



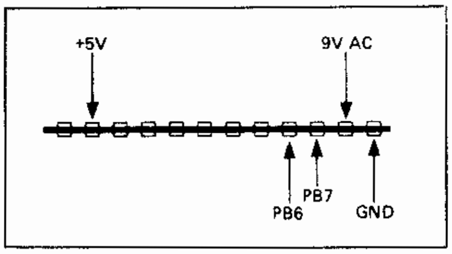


Fig. 5 Wiring details for the edge connector.

Parts List

Resistors (All 1/4W 5%)
R1,8 .... 3k3
R2,3,9,10 .... 56k
R4,11 .... 33k
R5,6,7,12,13,14 .... 8k2
R15 .... 22R
RV1,2 .... 220k linear pot.

Capacitors
C1,7 .... 470nF carbonate
C2,8 .... 220pF ceramic plate
C3,9 .... 3n3 carbonate
C4,10 .... 4n7 carbonate
C5,11 .... 330pF ceramic plate
C6,13 .... 10uF 25V radial elect.
C12,15 .... 100nF ceramic
C14 .... 470uF 16V radial elect.
C15 .... 220uF 16V radial elect.
C17,18 .... 100uF 10V radial elect.

Semiconductors
IC1 .... MF10CN
IC2,3 .... 741C
IC4 .... 79L05 5V 100mA regulator
D1,2 1N4002

Miscellaneous
SW1 .... 4 pole 3 way rotary
SK1-4 .... standard or stereo jacks
Sk5 .... 5 way DIN socket (180 degree)

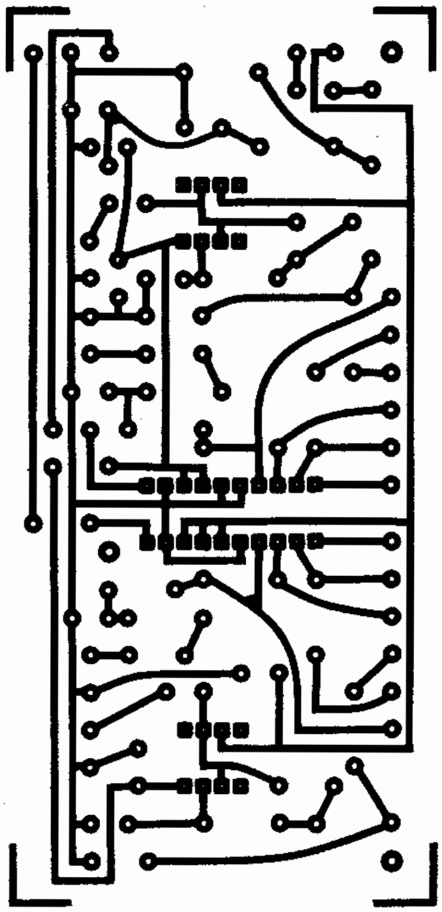
Case about 180 x 120 x 39mm; three control knobs; five-way cable; 5-pin (180 degree) DIN plug, and 2 x 12 way 0.1 inch pitch female edge connector; 20 pin DIL IC socket; wire, etc. *The MF10 switched capacitor device is available from Hobbilt Electronics, 7454 Langelier, St. Leonard P.Q., H1S 3B7, (514) 259-5581.*

```
10 POKE 56590,23
20 POKE 56591,23
30 POKE 56581,0
40 POKE 56583,0
50 R=25
60 FOR L = 1 TO 25
80 R=R-1
90 NEXT
100 FOR L = 25 TO STEP-1
110 POKE 56580,L:POKE 56582,R
120 R=R+1
130 NEXT
140 GO TO 60
```

This program moves the filter frequencies up and down in antiphase, using what is effectively a triangular modulation waveform. You do not really need to be a software genius in order to work out simple routines to vary the filter frequencies in various ways, and timing loops or dummy instructions can be added in order to reduce the modulation frequency. If you are familiar with 6510 machine code programming then it is possible to obtain faster operation if desired.

With the VIC-20 computer both filters must be driven from PB7 of the user port. A value of 192 is written to the control register at address 37147 in order to set the timers to the correct mode, 0 is written to the high byte of the timer/-counter at 37141, and the filter frequency is controlled by writing a value in the range 1 to about 25 or so to the low byte at 37140.

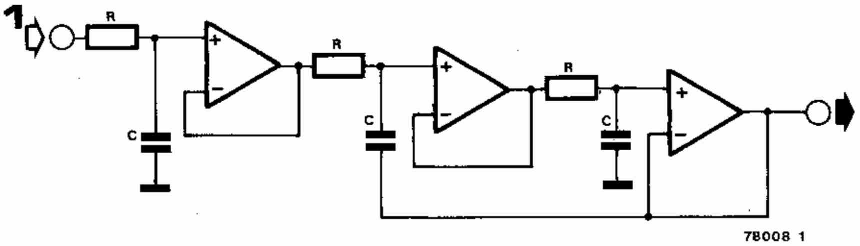
As pointed out earlier, any computer with one or two digital outputs could probably provide suitable clock signals with the aid of a machine code program. ■



27 18 dB per octave high/lowpass filter

Calculating the correct RC values for high- and lowpass filters can be something of a chore and is often regarded by amateur constructors as a subject that is best left well alone. This is especially true the more complicated the filter and the steeper the slope of the filter becomes. That being the case, the following circuit for a third order (i.e. with a slope of 18 dB per octave) high/lowpass Butterworth filter, together with the accompanying nomogram which supplies the correct RC values for any given turnover point, should prove extremely useful. The circuit shown is for a lowpass filter, however by changing over the position of the resistors and capacitors a highpass filter is obtained. The beauty of the circuit lies in the fact that all the resistors and capacitors have the same respective value. Either operational amplifiers or emitter followers may be used as voltage followers.

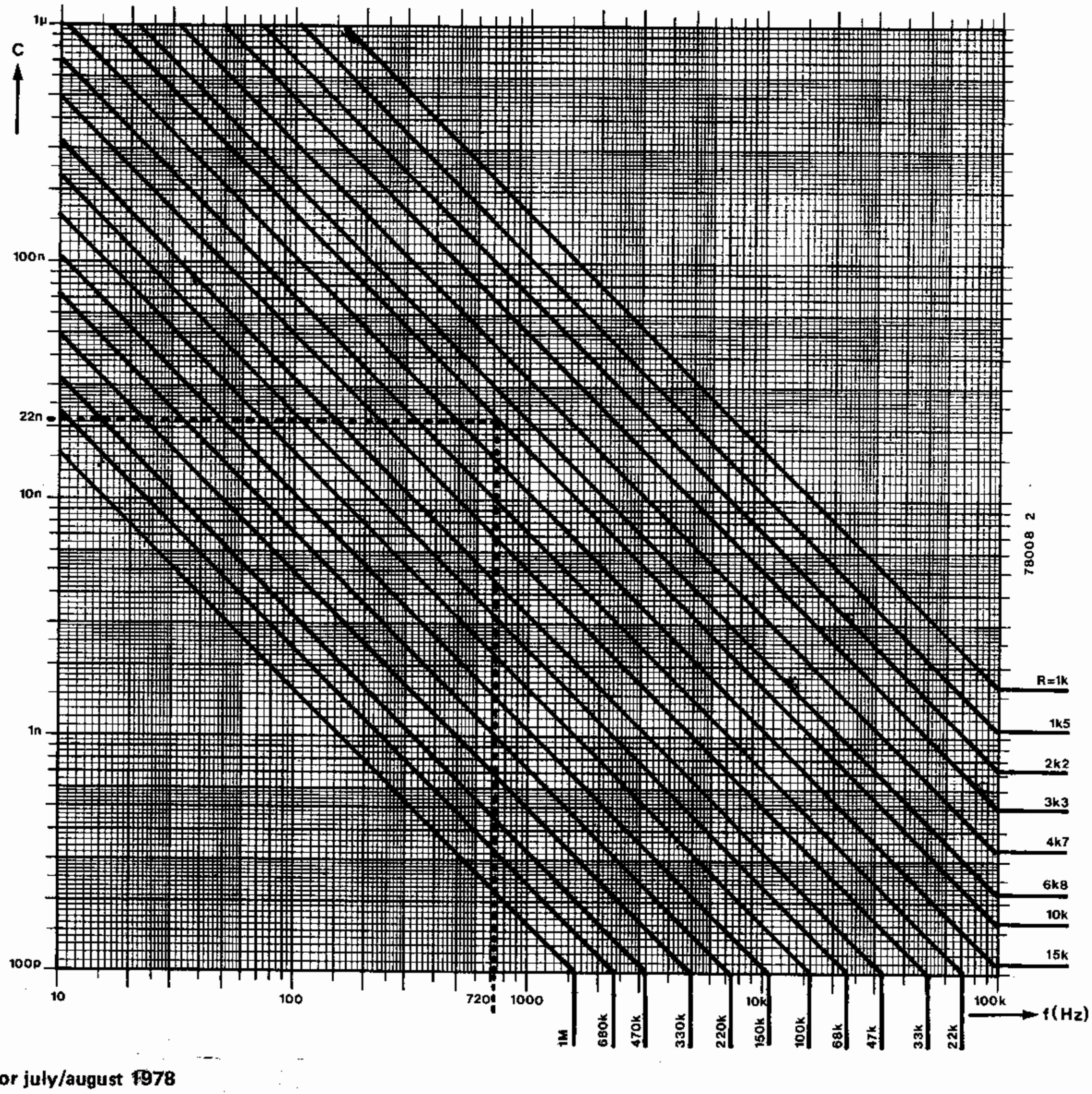
1



Normally, to find the turnover frequency of a filter (i.e. the point at which the output voltage of the filter is 3 dB down on the passband response) one uses the equation f_0 (the turnover point) $= 1/(2\pi RC)$. However one can forget about having to go through these calculations by using the accompanying nomogram. The turnover points are displayed along the horizontal axis, whilst the corresponding values for C are shown along the vertical axis. Furthermore, a number of resistance values are also indicated by the diagonal lines running across the nomogram.

To use the nomogram one first draws an imaginary vertical line through the desired turnover point. An imaginary horizontal line is then drawn through the point at which the vertical line intersects with the desired resistance. The intersection of that horizontal line with the x-axis gives the correct value of C for the chosen turnover frequency. In the example shown (in dotted lines), a turnover frequency of 720 Hz is obtained with R = 10 k and C = 22 n.

2



Mains rejection tracking filter

Using a tracking "n-path" filter with wide dynamic range

by K. F. Knott, B.Eng., Ph.D., M.I.E.E. and L. Unsworth, B.Sc.
University of Salford

The filter described greatly reduces interference at mains frequency and harmonics on wideband signals without seriously affecting these signals. It has the ability to track changes in the mains frequency, enabling very sharp rejection characteristics to be obtained. Useful rejection is maintained up to the 5th harmonic. The filter is based on the well-known principles of the commutating *CR* network but several improvements have been made to extend the dynamic range of this network without sacrificing signal bandwidth. For example, at mains fundamental a rejection greater than 40dB is maintained down to signal levels of 50mV r.m.s., the signal bandwidth being 100kHz. Consider the situation in which *N* identical capacitors are switched into a *C-R* network in sequence at a rate of Nf_o Hz (Fig. 1).

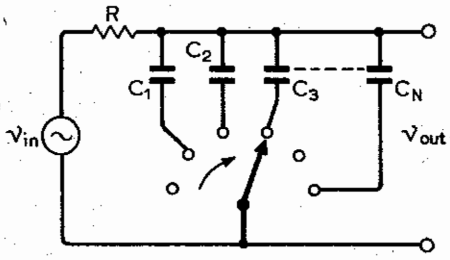


Fig. 1

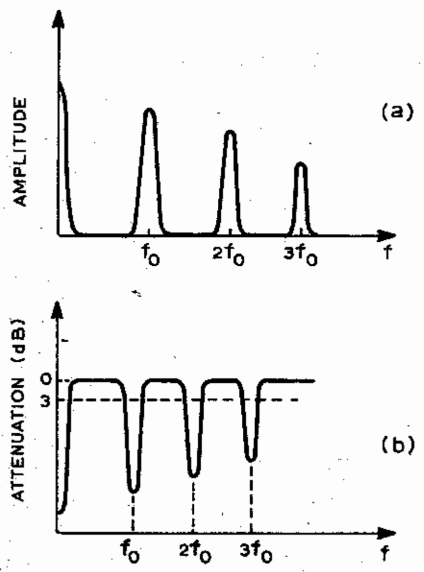


Fig. 2

The transfer characteristic of the network has the form indicated by Fig. 2(a), i.e. the network acts as a comb filter, the centre frequencies of which are set by the commutating frequency of the switch.[1] Alternatively, if the output is taken across the resistor the transfer characteristic of Fig. 2(b) is obtained.

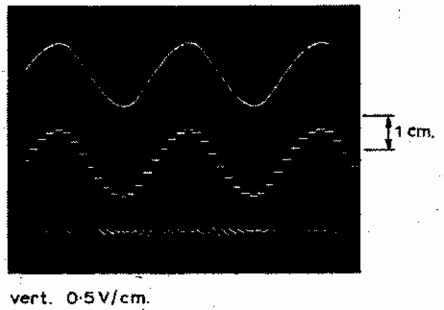
If the commutating frequency, Nf_o, is controlled to follow variations in f_o, the filter has the ability to track varying-frequency input signals therefore enabling the use of sharp notches while maintaining high attenuation. This is in contrast to fixed-frequency notch filters such as the bridged-T network. Although the mathematical treatment of commutating filters is well established it is useful to describe their operation in a non-mathematical way for the purpose of discussing problems which arise in the design of an instrument.

Principle of operation

Suppose the input signal v_{in} in Fig. 1 is sinusoidal at a frequency *f* Hz. If *f* is equal to nf_o, where *n* is an integer, the input signal will be in synchronism with the switch and each individual capacitor will be switched in at the same instant in each cycle of the input waveform. Each capacitor will charge up to the corresponding instantaneous value of the input waveform. This is analagous to sampling the input waveform with *N/n* samples per cycle. Obviously the upper limit on *n* is *N*/2.

The voltage waveform across *C* will not be sinusoidal but will resemble a "staircase" replica of the sinusoidal input voltage. The voltage across *R* will be the difference between the sine-wave and the staircase waveform. Consequently the action of the filter necessarily introduces high-frequency switching noise. An illustration of this noise is shown in the photograph of Fig. 3, which was taken for the case with f_o = 50Hz, $n = 1, N = 16$.

Consider now the action of the filter if *f* is a non-integral value of f_o. The input is no longer in synchronism with the switch and each individual capacitor will be switched in at varying points in successive cycles of the input waveform. The voltage across each capacitor will therefore be averaged to zero and the voltage across *R* will be equal to the input voltage. At input signal frequencies very much lower than f_o the switch may be considered to be rotating so rapidly that all *N* capacitors appear to be connected simultaneously. The circuit can then be thought of as a simple network with a time constant of *NCR* i.e. the voltage across *R* is down by 3dB at a frequency $1/2\pi NCR$ Hz. At input frequencies much higher than f_o the switch may be considered stationary and the network thought of as a simple network with a time constant of *CR*. This usually means that the voltage across *C* is very much smaller than the input voltage at frequencies greater than $Nf_o/2$ even though the commutation is no longer effective. Hence the voltage across *R* will be almost equal to the input voltage. The switching has the effect of reflecting the loss-pass response about f_o, $2f_o$, etc, thereby generating the comb-filter response of Fig. 2(a). The bandwidth is 2/*N* times the bandwidth of the original low-pass sections, i.e. $(2/N)(1/2\pi CR) = 1/\pi NCR$.



vert. 0·5V/cm.
horiz. 5m sec./cm.

Fig. 3

Design considerations

The desirable characteristics of a tracking mains interference rejection filter may be summarized as follows.

1. Minimum degradation of the signal which is to be transmitted through the filter.
2. Wide dynamic range and signal bandwidth.
3. High rejection of the fundamental and lower harmonics of the mains frequencies bearing in mind that interference signals are liable to fluctuate in amplitude.
4. Ability to track changes and rates of change of the nominal mains frequency.

As point 4 is subsidiary to the operation of the filter it is considered briefly before proceeding to a more detailed discussion of points 1, 2 & 3.

Tracking requirements

Statutory limits of the mains frequency in this country are 49.5Hz and 50.5Hz, although the likelihood of these limits being reached is low under normal circumstances. The rate of change of mains frequency is governed by the inertia of the generating plant and it is extremely unlikely that a rate of change of 0.1Hz/min. would be exceeded. The tracking requirements are modest therefore and the circuit described later has an adequate performance.

Rejection, signal bandwidth and dynamic range

A convenient way in which to discuss the performance of the filter is to consider the various properties of the basic circuit and then discuss how these properties may be improved. The basic filter, omitting the tracking loop, is shown in Fig. 4.

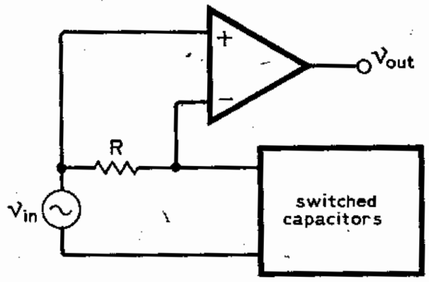


Fig. 4

Considering firstly the rejection characteristics of this circuit, as illustrated in Fig. 2(b), the sharpness of rejection is proportional to NCR. In theory one can obtain a very high Q-factor by choosing an appropriately large value of NCR. But an interference signal is likely to have a fluctuating amplitude. Suppose, for the sake of argument, that a 50-Hz interference signal was fluctuating sinusoidally in amplitudes with a period of ten seconds. Obviously this may be considered as a double-sideband signal with a carrier at 50Hz and sidebands at 50 ± 0.1Hz. If the Q of the filter at 50Hz were greater than 50/0.2 the sidebands would not be greatly affected. Although the analysis of sinusoidally modulated mains interference is a fictitious case it serves to illustrate that one must not have too high a Q-factor if fluctuating interference signals are to be rejected. Also, the step response of the filter is determined by its Q such that a slow response would result if a very high value of Q were used.

Theoretical magnitudes of rejection obtained at the synchronous frequencies can be found fairly easily by numerical analysis for specific values of N. The procedure is explained in the following paragraph.

Consider a sinusoidal input signal of frequency nf_o Hz. In the steady-state condition the voltage across each capacitor will reach the value of the input sine-wave averaged over the period for which the capacitor is connected. The voltage across each capacitor may be assumed constant provided that the CR time constant is large compared with the time spent on each capacitor and also if there is negligible discharge of the capacitors during the time between consecutive connections, i.e. $1/f_o$ sec. The waveform across the capacitors will thus be as illustrated in Fig. 5.

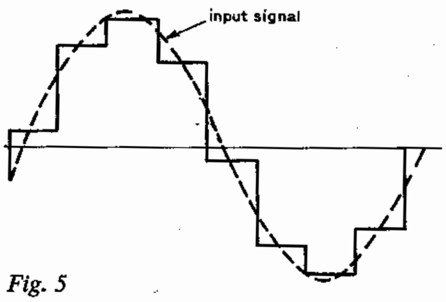


Fig. 5

The Fourier analysis of this type of waveform appearing across the capacitors may be found numerically by the "jump" technique.[2] As an example, suppose N were equal to 16. The analysis yields the result that for input signals of frequency f_o, $2f_o$ and $3f_o$, the fundamental components of the waveforms across the capacitors are respectively 0.97, 0.95 and 0.905 times the input. This would lead to rejections of 30.4, 26 and 20.4dB respectively if these fundamental components alone were subtracted from the input signal. However, these figures may be improved by weighting one of the inputs of the subtractor. In this way infinite rejection can be achieved at one of the synchronous frequencies, i.e. f_o, $2f_o$ or $3f_o$, etc. For example, if the circuit were trimmed to effectively increase the 0.97 figure to 1.00, the theoretical rejections at f_o, $2f_o$ and $3f_o$ would be ∞, 33 and 23dB respectively.

Considering, secondly, the dynamic range of the circuit, it was mentioned previously that the commutating action of the filter introduced high-frequency switching noise. Being more specific, if a 50-Hz signal were present at the input, switching noise would be introduced at $50N+50$, $50N$, $100N$, $150N$. . . etc, Hz. Furthermore, amplitudes of the switching noise components are at fixed levels below the 50-Hz signal. In general, the switching-noise component amplitudes decrease as N increases. As there is obviously a practical limit to the value of N the output of the basic filter will contain components of switching noise which will limit the dynamic range of the filter.

The simplest way in which to improve the dynamic range is to add a low-pass filter to the output as shown in Fig. 6, this of course reducing the signal bandwidth. To exploit the rejection properties of the commutating filter this low-pass filter should have negligible attenuation up to say $(N/2)50$Hz and high attenuation at $N50$Hz. The inevitable choice would be an active R-C filter.

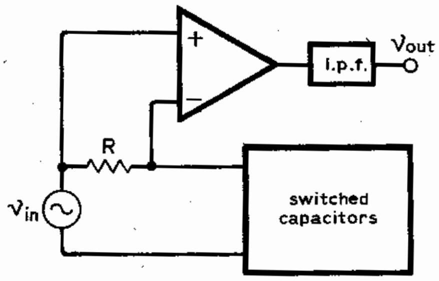


Fig. 6

Good dynamic range and signal bandwidth can be achieved if a low-pass filter is inserted in the position shown in Fig. 7.

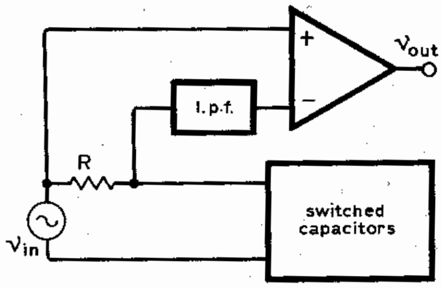


Fig. 7

The low-pass filter must again have a very sharp cut-off but unfortunately this cannot be achieved without introducing phase-shift in the pass-band. As a result the rejection decreases since the interference signals present at the differential amplifier inputs will no longer be exactly in phase.

This disadvantage may be overcome by inserting an all-pass filter in the signal path, having exactly the same phase response as the low-pass filter so that the interference signals present at the inputs of the differential amplifier are now always in phase, resulting in the final block diagram of Fig. 8.

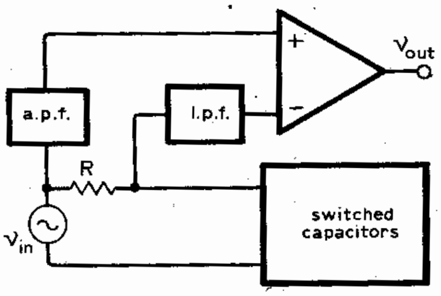


Fig. 8

Unfortunately the wanted signal now undergoes the phase-shift of the all-pass filter. This may or may not be important depending on the application.

To summarize, the filters based on the block diagrams of Figs 6, 7 & 8 have the following properties:

Fig. 6—high rejection, low signal bandwidth, good dynamic range

Fig. 7—high signal bandwidth, good dynamic range, moderate rejection

Fig. 8—high signal bandwidth, high rejection, good dynamic range but unsuitable for applications which require little phase-shift through the filter.

All of these characteristics may be obtained from the constituent parts of Fig. 8 by a suitable switching arrangement, though not simultaneously.

Choice of N and CR

Good rejection and tolerable levels of switching noise without overdue circuit complexity can be achieved with $N=16$. If a bandwidth of 1Hz at 50Hz is specified, i.e. $Q=50$, the filter will have a negligible effect on a wideband signal. Also, with a half-bandwidth of 0.5Hz reasonable rejection will still result at frequencies between 49.8 and 50.2Hz, i.e. the filter would reject a 50-Hz interference signal even if its amplitude were fluctuating over periods as short as 5s and further with a Q of 50, the time constant of the filter is 0.3s so that a rapid response to step changes in interference level is achieved.

Complete layout

The complete block diagram of a practical mains rejection filter is shown in Fig. 9. A switching arrangement has been adopted to make maximum use of the characteristics of the commutating network.

In position 1 (cf. Fig. 8) there is high signal bandwidth, high mains rejection, good dynamic range but considerable phase-shift between input and output. Position 2 again yields high signal bandwidth and good dynamic range but moderate mains rejection (cf. Fig. 7). However, the phase-shift is now constant over the audio range of frequencies. This is accomplished simply by shorting out the all-pass filter. The effect of the phase shift of the low-pass filter is to reduce the rejection of mains frequencies. However, the 50-Hz rejection is improved by introducing a simple lead network (C_1, R_1)

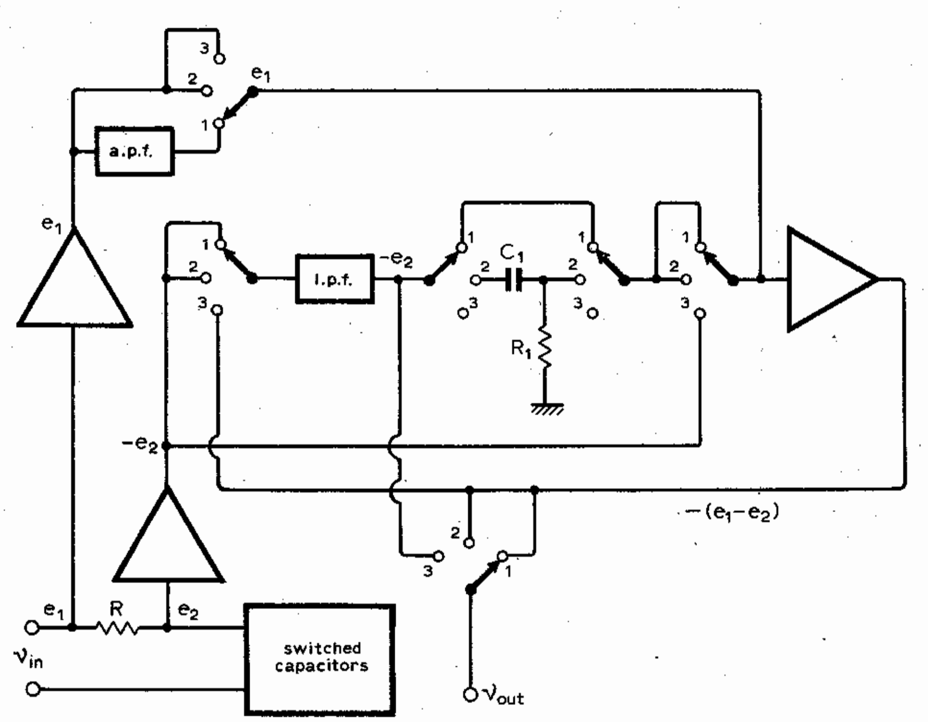


Fig. 9

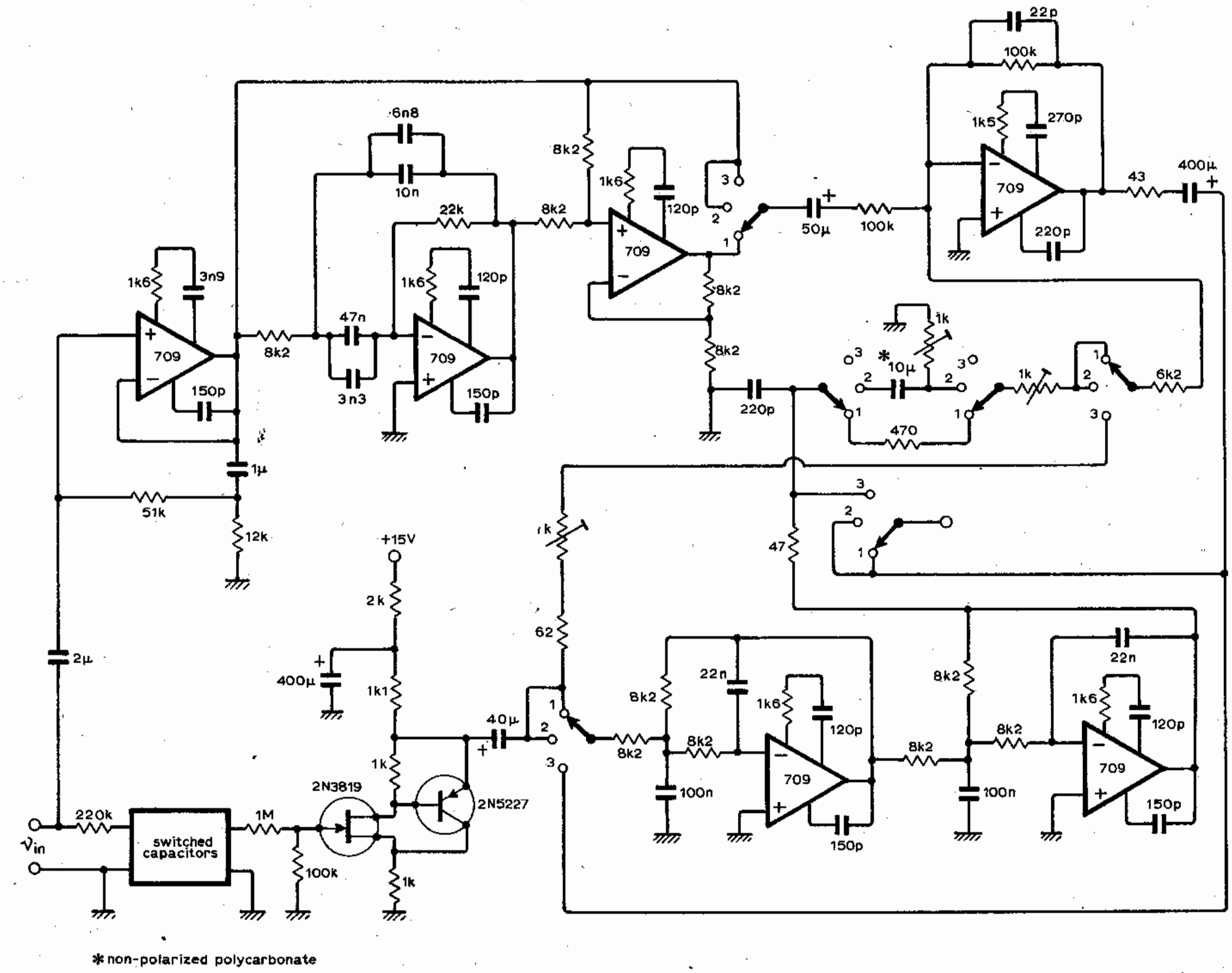


Fig. 10

chosen so that at 50Hz, though not at higher harmonics, the interference signals are exactly in phase at the inputs of the differential amplifier.

Position 3 gives high mains rejection, good dynamic range but low signal bandwidth, determined by the low-pass filter (cf. Fig. 6). This position was found to be desirable in certain applications where high frequency signals cause problems.

The low-pass and all-pass filters are both non-inverting and need to be preceded by buffers. Because an adder is far easier to align than a subtractor with its four variables we made the buffer preceding the all-pass filter a follower and the other an inverter, thus enabling an adder to be used to derive the required difference between the interference signals.

The circuit diagram corresponding to the block diagram of Fig. 9 is shown in Fig. 10.

Commutation

The 16 capacitors must be commutated electronically at 16 × mains frequency. Any one of a number of methods may be used to this end and the technique chosen is to drive two 8-way multiplexers alternately, both consisting of eight m.o.s.f.e.ts, each of which is switched on in turn with consecutive input clock pulses. The multiplexers are connected thus

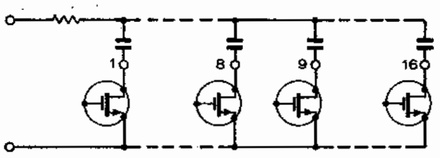


Fig. 11

The f.e.ts 1 to 16 are therefore arranged to switch on in turn. An 800-Hz clock (described later) drives a four-stage binary counter, the output of which is a 50-Hz square wave.

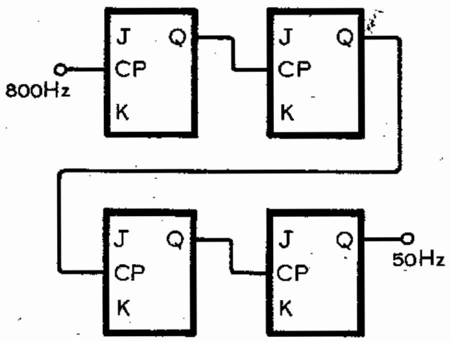


Fig. 12

In Fig. 12 all J and K inputs are permanently high. The 800-Hz clock is used to drive the two multiplexers. Consider just one multiplexer. Each f.e.t. is energized in turn as consecutive clock pulses appear at the input but after eight pulses, the clock waveform must be diverted to the second multiplexer which then switches capacitors 9 to 16 and then back to the first multiplexer, etc.

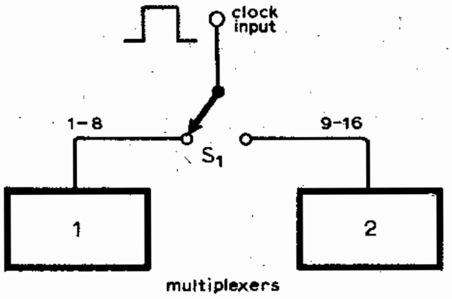


Fig. 13

Referring to Fig. 13, switch S_1 must toggle every eighth clock pulse. Now the output of the counter of Fig. 12 toggles every eighth clock pulse and so switch S_1 may be simulated as follows

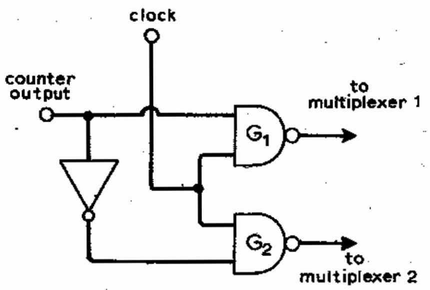


Fig. 14

When the counter output is high, gate G_1 is enabled and its output will then consist of the 800-Hz clock waveform. Meanwhile G_2 is closed. After eight clock pulses the counter output assumes a low state and gate G_2 is now enabled while G_1 closes.

Tracking oscillator

A multivibrator with a pulse repetition rate of N × mains frequency will provide the clock waveform. If the mains frequency changes slightly, then so must the multivibrator repetition rate to maintain synchronism.

Consider the following circuit

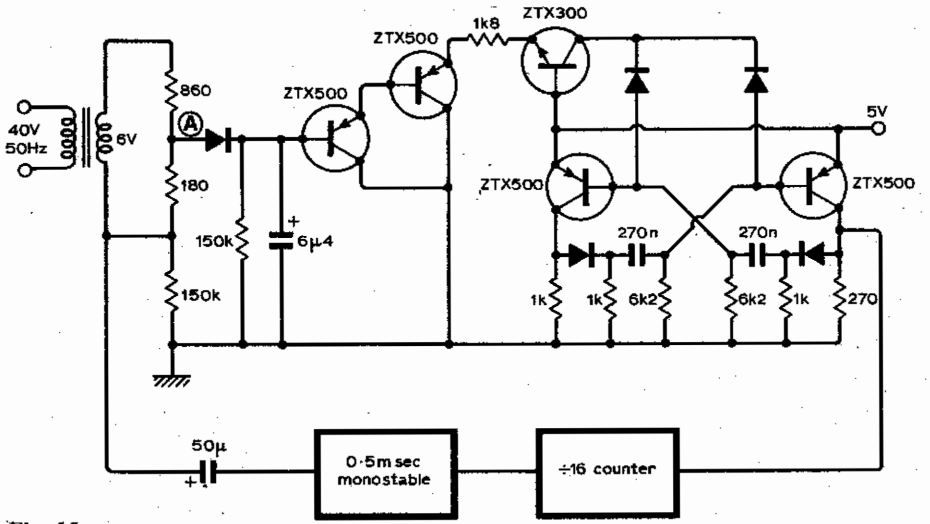


Fig. 15

The waveform at point A will be a 50-Hz sinewave with a pulse superposed on it:

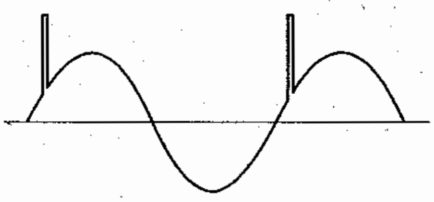


Fig. 16

When the multivibrator is synchronized to the mains frequency, the 0.5ms pulse will sit on the sinewave at some particular point. If the mains frequency now changes slightly, the pulse will climb up or slide down the sinewave and if the peak value of the waveform of Fig. 16 is detected, the resulting voltage can be used to vary the multivibrator rate to maintain synchronism with the mains.

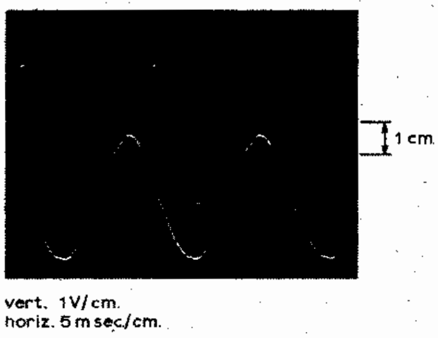


Fig. 17

Fig. 17 shows a photograph of the waveform at point A. The monostable of Fig. 15 is based on that given in reference 4.

A graph of p.r.r. versus mains frequency is shown below

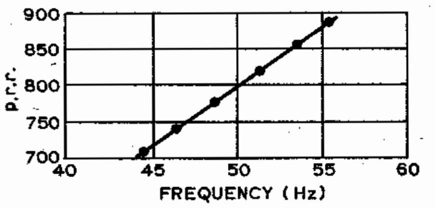


Fig. 18

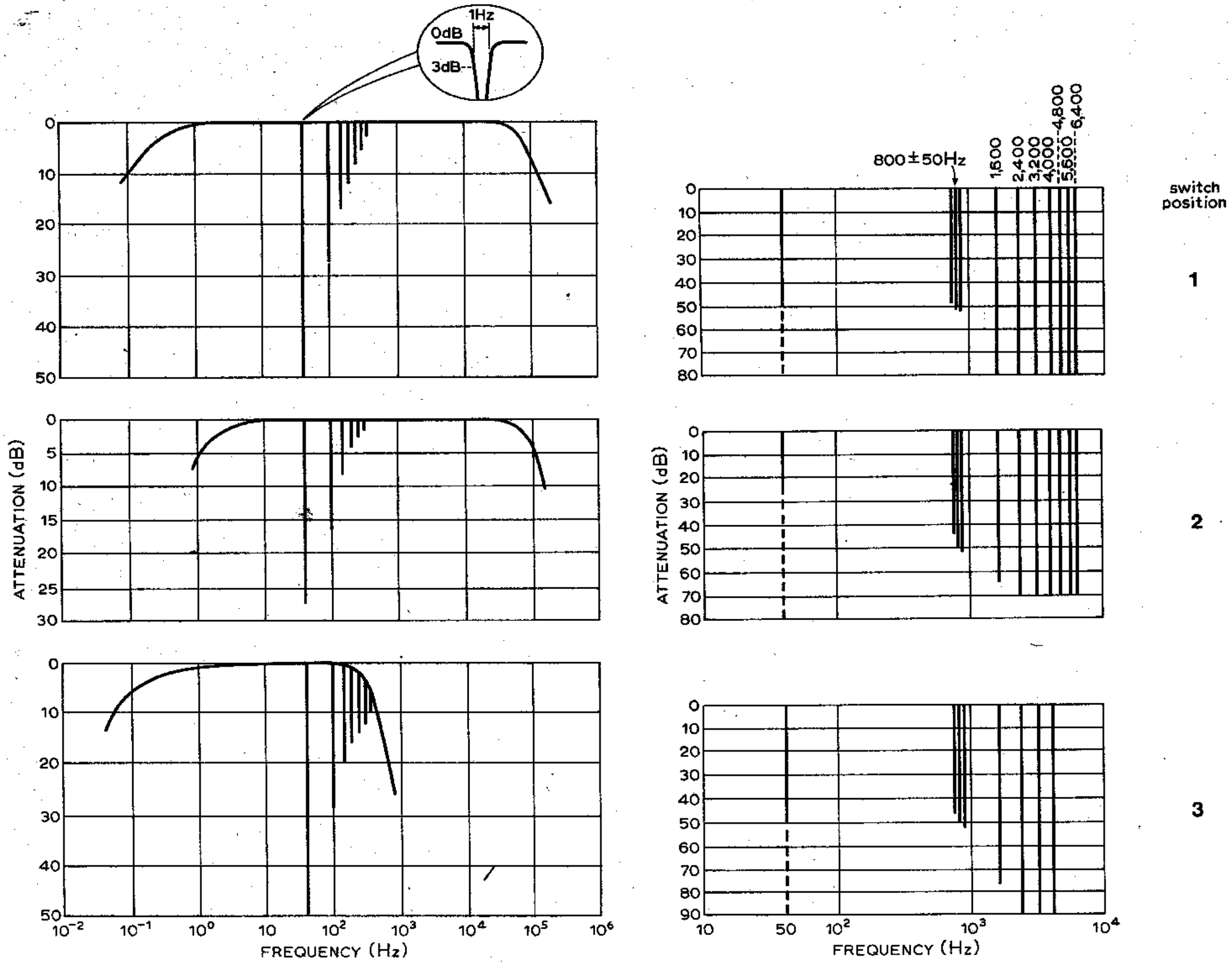


Fig. 19

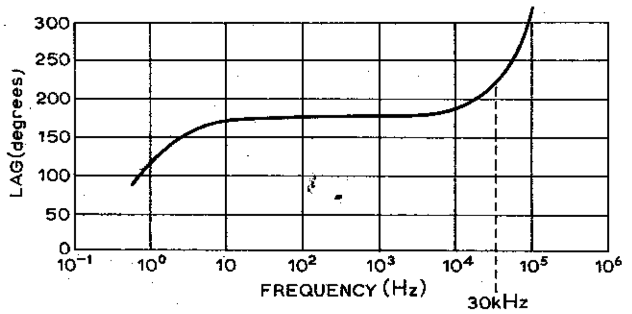


Fig. 20

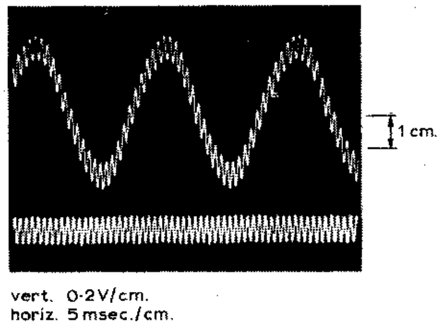


Fig. 21

Performance

In position 1 (see Fig. 19, top left), 50dB of rejection at 50Hz was maintained down to 100mV and up to 2V rms and 40dB of rejection down to 50mV. A bandwidth of 100kHz was maintained up to levels at which the slew rate of the operational amplifiers employed (709s) imposed restrictions.

The graphs on the right-hand side of Fig. 19 illustrate the relative amplitudes at the output terminals of an unwanted 50-Hz signal and its associated switching components, the input 50-Hz signal level being 0dB.

In position 2, 27dB of rejection was achieved at 50Hz, again from 100mV to 2V r.m.s. Phase response is shown flat from 2Hz to 30kHz in Fig. 20.

In position 3, 50dB of attenuation was measured between 100mV and 2V r.m.s.

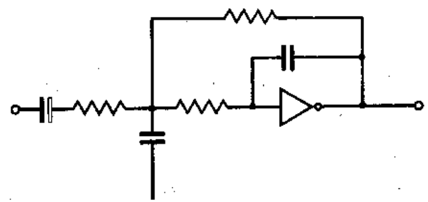
The 3-dB bandwidth of all the notches of the left-hand graphs was approximately 1Hz.

Fig. 21 illustrates the effectiveness of the filter where the top trace shows a 1-kHz sinewave swamped by 50Hz and the lower trace displays the 1-kHz signal after being processed by the filter.

References

1. Broeker, W. Commutating Techniques, Motorola application note AN534.
2. Kreyszig, E. Advanced Engineering Mathematics, Wiley 1964.
3. Unsworth, L. Using junction f.e.ts, *Wireless World*, vol. 78 1972 p.222 (article covers pp. 219–22).
4. Cole, H. A. TTL trigger circuits, *Wireless World*, vol. 78, 1972, pp.31/2.

FILTERS USING CMOS



High pass and low pass filters may be readily constructed using CMOS inverters (CD4007 CD4069 74C04) since these have only a single complementary pair, and hence lower power dissipation and less likelihood of instability. A form of Sallen and Key:

Standard equations are used to determine component values. It is recommended that passband gain be restricted to unity.

DESIGNING & USING ACTIVE FILTERS PART 1

A SHORT SERIES BY TIM ORR WHICH WILL ENABLE THE HOME CONSTRUCTOR TO UTILISE CIRCUITS OF HIGH COMPLEXITY AS EASILY AS PLUGGING IN A RESISTOR!

THERE IS NO DOUBT that active filters are very useful devices. Also, there is no shortage of literature on the subject. This would seem to suggest that designing active filters is a fairly straightforward business. Well, it is and it isn't. It is if you read *this* article. It isn't if you read the aforementioned literature. Most of the books on this subject have filled our heads with terms such as poles and zeros, Laplace transforms, transfer functions, etc, which haven't actually helped *us* to 'design anything!

Some basic theory

It is advisable quickly to run through some basic terms and expressions. Firstly, consider a simple low pass filter, Fig. 1a. The frequency response (Fig. 1b) is nearly flat until the break point, denoted by fb. After this point the response rolls off at 6dB/octave, that is signals above this frequency are increasingly attenuated. The break point is defined as being the frequency where the resistance equals the capacitive reactance. When this occurs, the output is attenuated to 0.707(−3dB) of the input. Although the resistance equals the capacitive reactance, the output is *not* half of the input. (This is because it is the vector sum of the two and hence equals 0.707 of the input!)

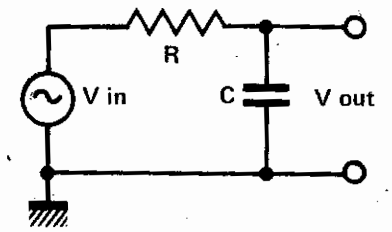


Fig. 1a. Simple low pass filter.

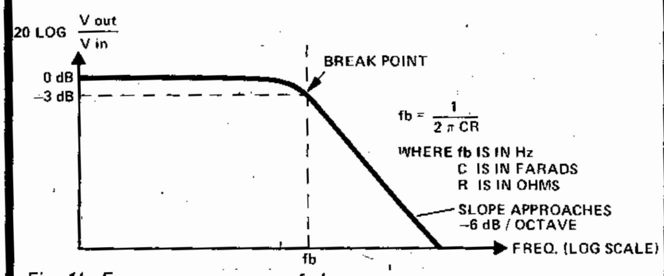


Fig. 1b. Frequency response of above.

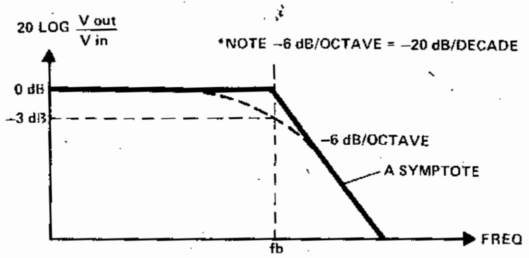


Fig. 1c. Approximation to response.

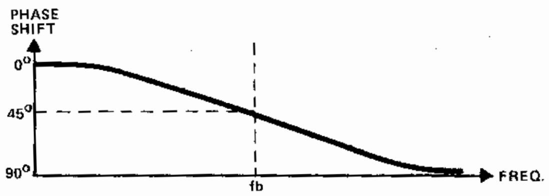


Fig. 1d. Phase shift v Frequency plot.

As the frequency response is a rather complex curve it is very useful to use a straight line approximation to it. These lines are called asymtotes (Fig. 1c). Note that the frequency response graph uses the convention of logarithmic scales, octaves or decades along the frequency axis, and dBs along the vertical axis representing output voltage divided by input voltage.

Phase shift with respect to frequency is also often plotted as in Fig. 1d. These two (the phase and frequency response plots) are known as Bode diagrams and are generally considered the most useful way of representing a filter's performance.

You will note that for the lowpass filter of Fig. 1a, the phase shift starts at 0°, is 45° at fb and then approaches 90° as the frequency approaches infinity. This is not an *active* filter, it is composed entirely of *passive* components which means that its output cannot be effectively loaded without changing its performance.

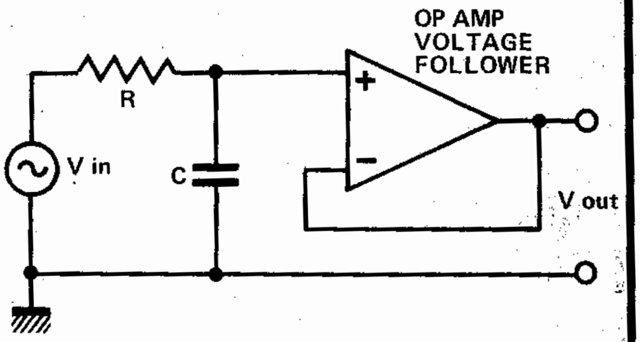


Fig. 1e. Active filter to perform the same task as the passive circuit of Fig. 1a.

Fig. 1e shows the same filter but in its active form, the op amp being used as a voltage follower serving only to isolate the filter's output. This type of filter is known as a First Order filter — a measure of the roll off slope.

When a more rapid slope is required, a higher order filter structure (one with more reactive elements) must be used. This is dealt with later.

ACTIVE FILTERS

Summary of low pass filter of Fig. 1.

Filter type	Low pass
Filter order	First order
Roll off slope	−6dB/octave or −20dB/decade (the same)
Breakpoint fb	fb=1/2πCR Hz
Phase shift at fb	45°

TABLE 1.

Passing highs

Next, let us consider the simple high pass filter of Fig. 2a. It is the complement of the low pass filter, the elements having been interchanged. Therefore it is not

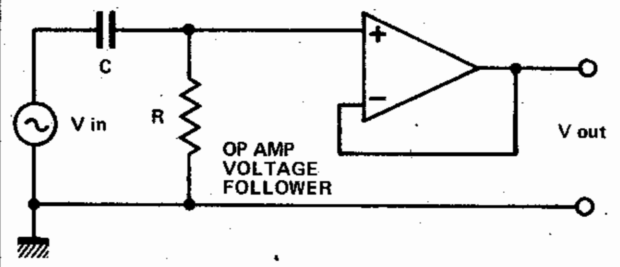


Fig. 2a. Simple high-pass active filter.

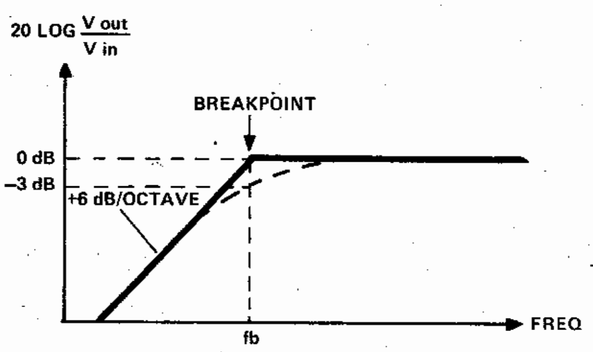


Fig. 2b. Frequency response of the high-pass filter.

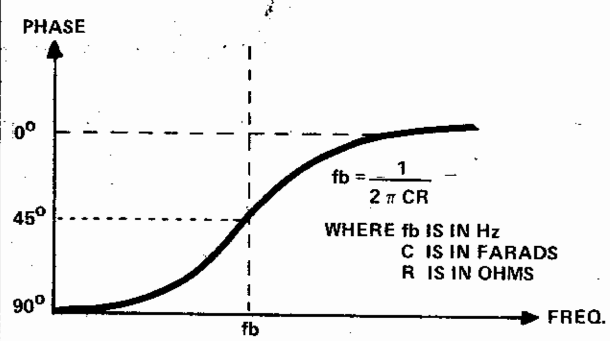


Fig. 2c. Phase v frequency plot of the same filter.

difficult to accept the complementary phase and frequency response curves of Fig. 2b. Note that the break point is the same and so is the roll off slope.

Summary of the high pass filter of Fig. 2.

Filter type	High pass
Filter order	First order
Roll off slope	+6dB/octave or +20dB/decade
Break point fb	fb=1/2πCR Hz
Phase shift at fb	45°

TABLE 2

Passing bands

The next type to be considered is a simple band pass filter shown in Fig. 3a. Although it uses an inductor it is only to illustrate the bandpass theory. Later on in this series, inductors will be replaced by their active equivalents.

The frequency response (Fig. 3b) shows that this circuit is symmetrical, having roll off slopes of 6dB/octave on either side of its RESONANT peak. This

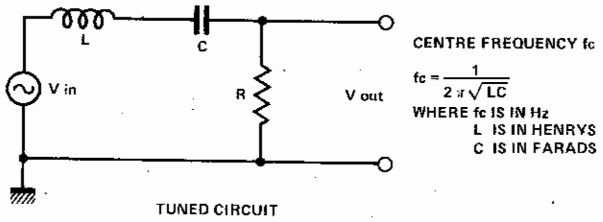


Fig. 3a. Simple band-pass filter.

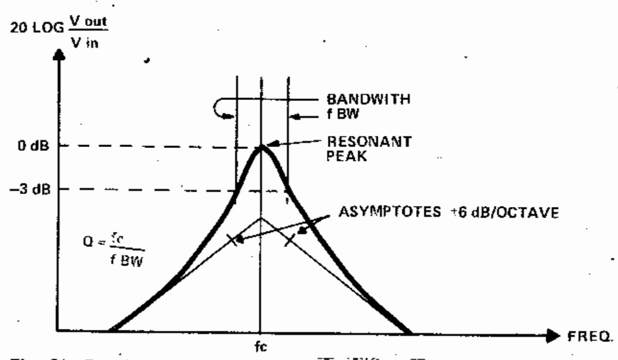


Fig. 3b. Band-pass frequency response.

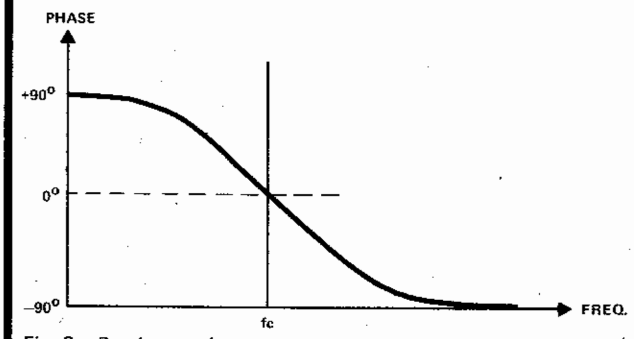


Fig. 3c. Band-pass phase response.

filter is known as a second order filter, because it has two reactive sections, the L and the C. The C is responsible for the +6dB/octave portion of the slope, the L for the −6dB/octave portion. But where these two slopes should meet, the response of the filter peaks and the slopes become much larger (Resonance). The sharpness of this peak is described as the Quality of the filter, the Q factor. Resonance occurs at a frequency known as the Centre frequency denoted by fc.

The bandpass filter is so called because it only passes signals within a certain bandwidth, which is defined as being the frequency range contained between the two points that are 3dB below the resonant peak. There is a fixed relationship between Centre frequency (fc), bandwidth (fbw) and Q factor, given by Q=fc/fbw.

The centre frequency is given by $fc \simeq 1/2\pi\sqrt{LC}$ Hz. This is only approximate, as it assumes that the value of R is relatively low. As R decreases, the Q factor increases. Thus R has the effect of damping the resonances, and so as it approaches zero ohms, Q approaches infinity.

The phase shift is shown in Fig. 3c. As this filter is a second order structure, then the total phase movement will be twice that for a first order structure, i.e. 180°. Fig. 3d shows the phase and frequency responses for different values of Q. Note that a high Q has a very rapid rate of change of phase, a low Q has only a slow rate of change.

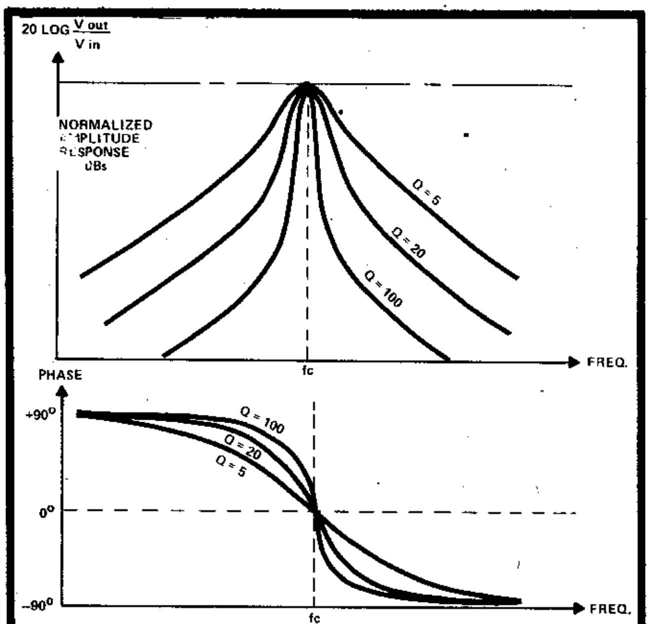


Fig. 3d. Effect of varying Q on the frequency and phase characteristics.

Time please

Bandpass filters also have a time response, as opposed to their frequency response. When an impulse is applied to a bandpass filter it rings (Fig. 3e). The filter oscillates at the centre frequency, fc, the amplitude of the oscillations decaying exponentially in time. The ringing time, T_r, is the time taken for the oscillations to decay to 37% of their initial value. Ringing time is related to the Q and fc by the following equation:

$$T_r = Q/2\pi fc$$

When a high Q filter has been constructed, it may prove difficult to measure its Q factor accurately due to the narrowness of its bandwidth. However, if the filter is made to ring, a reasonably accurate measurement of the Q can be obtained by measuring T_r and fc.

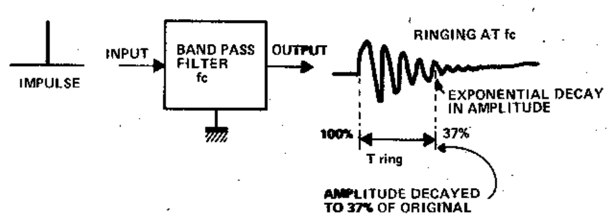


Fig. 3e. Ringing in a band-pass filter.

Filter type	Band pass
Filter order	Second order
Roll off slopes	+ and −6dB/octave greater near to resonance
Centre frequency fc	$fc \sim 1/2\pi\sqrt{LC}$
Phase shift at fc	0°
Q factor	fc/fbw where flow is the 3dB bandwidth
3dB bandwidth fbw	fc/Q
Ringing time, Tr	$Q/2\pi fc$

TABLE 3. Summary of band-pass filter.

Failed band

Another common filter structure is the band reject or notch filter. There are many ways of realising this filter, one of which is shown in Fig. 4. The input signal is subtracted from the bandpass output. By adjusting Ra with respect to R, complete cancellation can be obtained at fc.

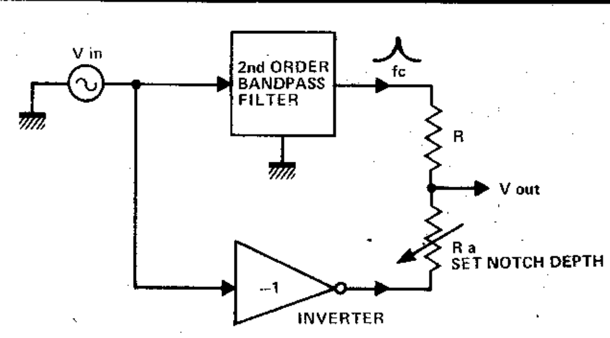


Fig. 4. Notch filter using Op-Amps.

Thus the centre frequency of the bandpass filter is the centre frequency of the notch, whose depth can be varied by altering Ra.

Very deep notches are possible, 50dB is easily obtained. As the Q of the bandpass filter is increased, so is the Q of the notch filter. However, Ra will have to be reset for each value of Q.

Filter Order

Consider the ideal low pass filter shown in Fig. 5a. Its response is flat right up until the break frequency fb. Frequencies above fb are attenuated to nothing! You won't be surprised to learn that filters like this don't exist. However, it is a common requirement to produce filters with very steep roll off slopes and this is achieved by designing filters with lots of sections, to increase the filter order. Each reactive element in the filter increases the filter order by one, therefore a low pass active filter with three capacitors is known as a third order filter and will have an ultimate roll off of three times 6dB/octave, which is 18dB/octave.

However, designing a third order lowpass filter is not just a simple case of sticking three first order RC circuits in a line. What you get when you do this is a very soggy curve indeed! The filter should be flat in the pass band, then it should turn over and rapidly assume its ultimate roll off slope. Examples of this type of Maximally flat filter are shown in Figs 5b and c. The effect of order number upon a bandpass filter is shown in Fig. 5d.

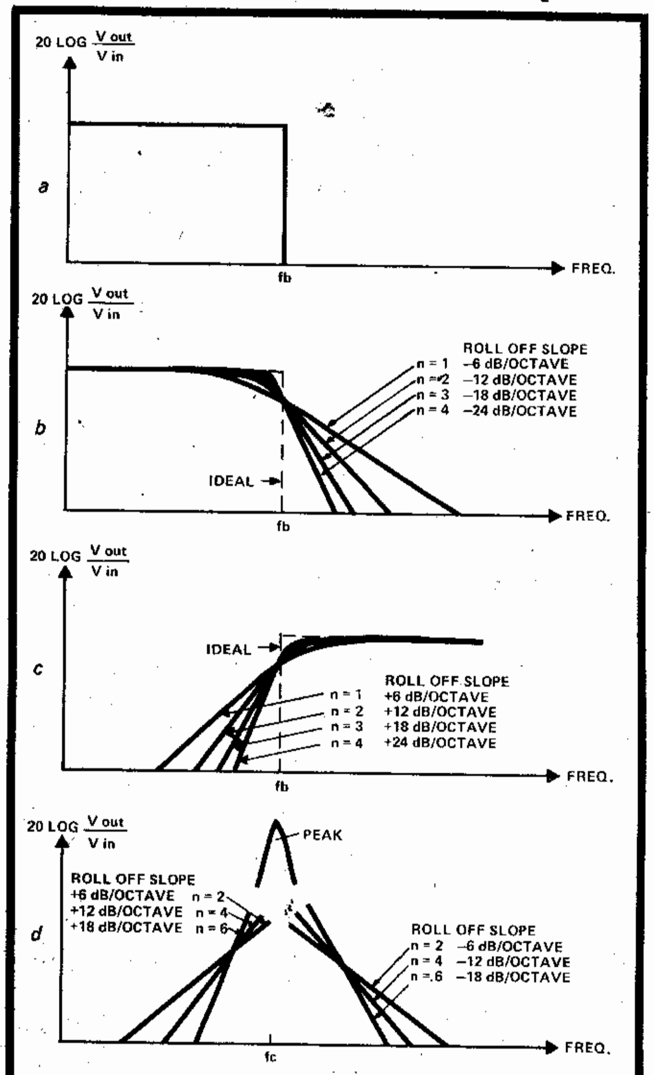


Fig. 5a. Ideal low-pass response;
b, Examples of maximally flat;
c, Filter responses;
d, Effect upon band-pass filter of increasing order number.

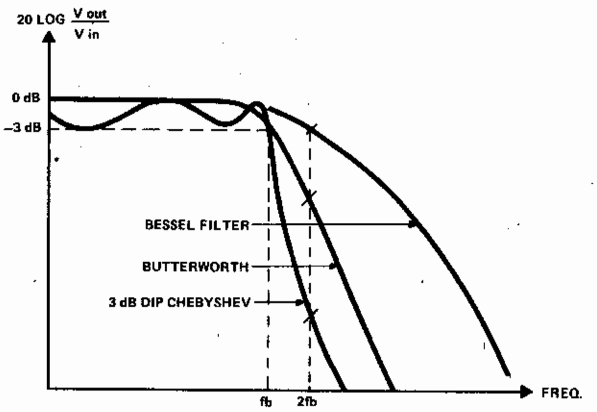
Later on in this series the circuit diagrams and design charts are given for various filter types and order numbers. It would seem that to get a filter to approach its ideal response, all that is needed is to increase the order number. This is in fact true, but there are certain tolerance problems. (When 8th order filters are designed, component tolerances of about 1% are required!)

Which filter shape?

The type of filter that is chosen to do a particular job will depend on what parameters are thought to be important. There are three basic characteristics to be considered (lowpass and highpass filters only).
1. Good transient response.
2. Maximum flatness of the filter within its passband.
3. Maximum rolloff slope outside the passband.

The type of filter used should be chosen to fit the job that they are being designed for. The filters have been categorised into three basic types for the purpose of simplicity.

Filter number 1 is known as a Bessel filter. Its phase changes almost linearly with frequency. It is useful for systems where a good transient response is required, such as joining the dots up on the output of a digital to analogue converter. It has a very poor initial roll off slope.
Filter number 2 is known as a Butterworth filter. It has the flattest pass band possible. Its other two parameters are a compromise. That is it has a reasonable overshoot and a fairly fast initial roll off.
Filter number 3 is known as a Chebyshev filter. It has some ripple in its pass band, although this is small, and a very fast initial roll off, and a poor transient response.



ATTENUATION AT FIRST OCTAVE (2 fb)

*3 dB CHEBYSHEV	17	28	39	51	62	75
BUTTERWORTH	12	18	24	30	36	42
η/FILTER ORDER	2	3	4	5	6	7

*NOTE THE IMPROVED ATTENUATION

Fig. 6. Response of all three types of filter discussed, with table showing variation in attenuation between them.

Next Month: Full design charts and circuits for three types of Active Filter.