Wireless World Circard

Series 1: Basic active filters

This set of cards—the best selling one of all—deals mainly with second-order active filters i.e. those with active devices for gain having a second-order transfer function. The most popular nowadays are probably the Sallen & Key types using unity-gain amplifiers (cards 4 and 5). Though widely used, they suffer from the disadvantage that two resistors need to be made variable and ganged for variable frequency applications. The positive feedback Wien-bridge notch filter too needs to have two components made variable to alter its centre frequency. Three cards deal with filters that are tunable by just one potentiometer (cards 3, 8, 10), and card 9 includes a bibliographical reference to a not so well-known twin-T circuit having single-component variation of the notch frequency.

One tunable filter technique that uses digital integrated circuits is given in the *n*-path filter card. Here resonant frequency is varied by changing commutating frequency, leaving bandwidth constant.

Many of the circuits given use integrated-circuits, mostly the 741 op-amp. (Card 6 is an exception, using the TAA960 i.c. with four inverters.) But for many low-frequency, low-Q applications, the op-amps can be replaced by a single transistor or Darlington pair. In the Sallen & Key types, the discrete device would need to be used as an emitter follower.

Basic Active Filters

Like the doting parent reproached for a birthday present of an encyclopaedia with the response "It tells me more about things than I want to know", any writer on this topic has to exercise care.

The filters to be described belong to that broad class of circuits using active devices together with passive components, some of which are reactive, to produce a frequency-dependent transfer function. The possibilities range from single *CR* time-constants in the feedback path of an amplifier to complex multi-gyrator circuits. The former have applications in audio frequency work, while the latter are still the subject of continuing research. It is the in-between varieties that we shall consider, in particular second-order filters of various kinds.

These have numerous applications used separately but may in addition be combined to produce more complex filter functions. By second-order is meant that the transfer function is represented by an equation containing ω^2 as well as ω. This is because of the presence of at least two reactive components — capacitors or inductors. The latter are inconvenient and expensive below the radio-frequency range, and active-filters dispense with them by synthesizing the desired function using capacitors.

The common filter functions are

Low-pass: signals below a given cut-off frequency are passed at, or amplified to, a fixed level, falling by 3 dB at the cut-off frequency with a slope that reaches 12 dB per octave for a second-order filter — Fig. 1.

High-pass: the pass-band lies above the cut-off frequency, with 3 dB attenuation at the cut-off frequency, and with attenuation increasing at 12 dB/octave below this frequency — Fig. 2.

Band-pass: signals at both low and high frequencies are progressively attenuated with a maximum response at the centre frequency; the sharpness of the response is measured by the Q of the circuit with $Q = f_0/2\Delta f$, where f_0 is the centre frequency and $2\Delta f$ is the bandwidth for an attenuation of 3 dB relative to that at the centre frequency — Fig. 3.

Band-stop or notch: low and high frequencies are passed with a maximum attenuation (ideally zero transmission) at the centre or notch frequency — Fig. 4.

All-pass: the magnitude of the transfer-function is constant but with a frequency-dependent phase-shift — Fig. 5.

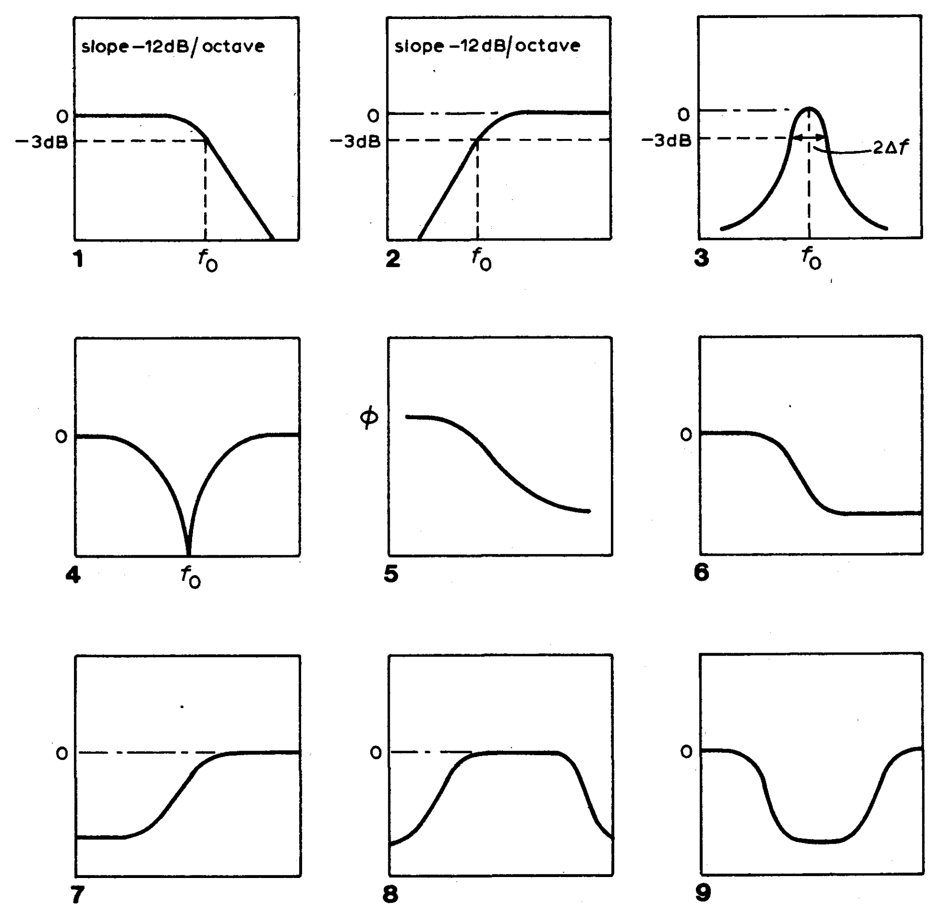
The low- and high-pass filters are sometimes modified so that the output falls to a low but constant value at some other frequency — Figs. 6 and 7. The band-pass and band-stop filters may require extended frequency ranges within which the pass and stop functions are maintained. One possibility is the use of complex high-order filters. Alternatively a combination of one or more second-hand filters tuned to different frequencies can be used. A perfect null response is possible only at precise specified frequencies — Figs. 8 and 9.

In determining which active circuit configuration most suits a particular filter function, the required Q is a major factor. High Q is often a requirement for band-pass and notch filters. It can be attained by two fundamentally different methods.

The first, requiring a single amplifier in the simplest cases, arranges for Q to depend on the difference between two nearly equal terms in the transfer function. The circuits using this technique are sometimes recognizable as a bridge circuit with the bridge almost at balance, combined negative and positive feedback with the latter almost equalling the former, or a negative-impedance converter in which an impedance in one part of a circuit appears elsewhere with a negative sign.

Any given circuit may be recognizable under more than one of the above headings which might be considered as different manifestations of the same underlying process. While high Q is possible, it becomes over sensitive to component and environmental changes, as it depends on the near-cancellation of comparable terms. The fractional change in Q caused by a change in some component is broadly proportional to the nominal Q aimed for, e.g. a circuit with $Q = 20$ as the design value might have a practical value of $Q = 25$ for a 1 or 2% change in a particular resistor.

The corresponding practical values for nominal Q values of 10 and so might be 11 and 75 respectively. The centre frequency is also dependent on component values but active circuits are comparable to passive networks in this respect. The exception is at high frequencies when the amplifier limitations may be all too apparent. This is shown by the increasing departures at the observed



In these graphs illustrating low-pass, high-pass, band-pass, band-stop and all-pass active filters, transfer function magnitude is plotted as the ordinate with frequency as the abscissa, except in 5 where the ordinate is phase angle.

centre-frequency from the nominal value, coinciding often with a fall in the achievable value of Q.

The second high-Q approach uses multiple active devices. One type, well-known in analogue computing, uses two integrators in a closed loop such that the open-loop gain of the system at the centre frequency, and hence Q, is controlled by a single resistor. The same pattern of passive components reappears in filters based on the gyrator, a device which has an impedance at one port related to that presented to a second port in such a way that a capacitance is "gyrated" into an inductance, a parallel tuned circuit into a series tuned circuit, etc. Gyrators may be constructed from various configurations of amplifiers, with the usual limitations on frequency response, for example.

In neither approach is the Q excessively sensitive to component variations. Broadly the sensitivity is comparable to that obtaining with passive circuits, i.e. a 1% change in any component will not normally change the Q by more than 1% even at high values of Q. With multiple amplifier filters, there is more than one output available, and some circuits have band-pass, low-pass and high-pass functions available simultaneously. The falling cost of operational amplifiers makes such solutions attractive, particularly for larger systems where the flexibility is a considerable advantage. A problem can be the phase-shift in multiple amplifiers, which can result in oscillation when high Q is attempted at frequencies near the upper frequency limit of the amplifier.

For simple low-pass and high-pass applications the single-amplifier methods are suitable, and as an example scratch and rumble filters may use twin-gang potentiometers for widely variable cut-off frequencies. Using special i.c. amplifiers with very high input resistance the cut-off frequency can be readily extended below 1 Hz if required, while the upper limit approaches 1 MHz.

A completely different approach which is certain to become dominant in many fields is that of the n-path filter. There will presumably be as many variants as there are at present with the analogue approach, but the basic technique appearing in some recently published circuits is as follows.

In place of a single filter a number, say n, of identical units are switched into the circuit in succession. These may be, for example, low-pass filters, passive or active, or simply the capacitors of a set of CR circuits using a common R. Switching frequency is set to n times the fundamental frequency to which the filter is to be tuned.

The filter has responses at harmonics of the fundamental frequency and may have a very high Q at each, but simple low-cost filters of the type described earlier can be used to attenuate these responses.

A vital point is that the filter centre-frequency is controlled by the clock-rate driving the switches, and therefore readily variable over a wide range, as is bandwidth.

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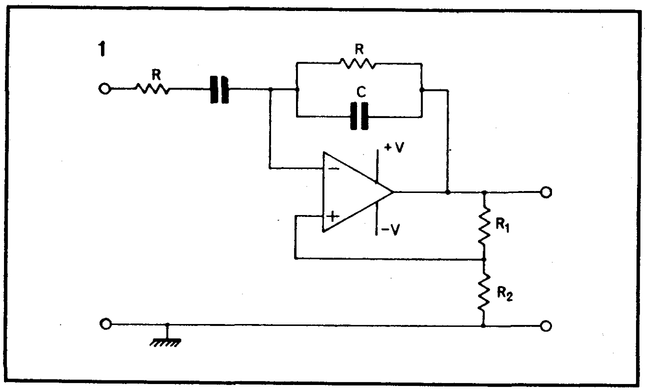
Wien-bridge bandpass filter

1


Typical performance at 1kHz

$$f_0 = \frac{1}{2\pi CR}\ \text{Hz}$$

$$Q = \frac{1}{2-(R_2/R_1)}$$

Voltage gain, $\dfrac{V_{out}}{V_{in}} \approx -3Q$

To find Q measure voltage gain and divide by -3

IC : 741; supplies ± 15V
$C = 1$nF$\pm 1\%$
$R = 150$kΩ $\pm 5\%$
$(R_2 + R_1) = 10$kΩ
Source resistance: 60Ω
V_{out} (max.) $= 9{\cdot}2$V r.m.s.
Max. load current: 12mA
Q is constant for supplies between ± 3V and ± 18V if clipping is avoided.
$Z_{in} \propto 1/Q$; $Z_{out} \propto Q$.

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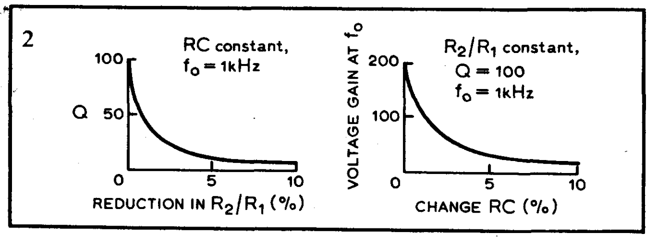
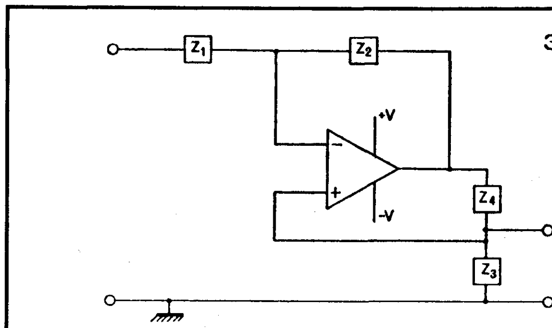


Circuit description

The Wien network–Fig. 1–gives a maximum response at $f_0 = 1/2\pi RC$ Hz. Positive feedback through R_1, R_2 sharpens the response, e.g. with R_1 and R_2 replaced by a potentiometer, continuous control of circuit Q is obtained without change in centre frequency. Changing both Rs or both Cs, maintaining equality, varies f_0 without changing Q. Sensitivity to component changes is proportional to Q for large Q. At high frequencies, amplifier phase shift causes Q and f_0 to depart from nominal values. Input impedance falls as Q increases. Satisfactory for moderate Q values (5–25) at frequencies in the audio range. High Q is obtainable if only short-term stability is required.

Component changes

- Using $\pm 1\%$ capacitors and $\pm 5\%$ resistors, f_0 will typically be within $\pm 5\%$ of theoretical value up to approximately 15kHz and within $\pm 10\%$ up to about 22kHz.
- Large R-values for low f_0 produces an output d.c. level up to about 1V which can be reduced with offset null adjustment.
- Source resistance should be $\ll R$ for predictable f_0 and $< R/15Q$ for predictable Q within -5%.
- Circuit will oscillate when $R_2/R_1 \geqslant 2$.
- 741 op-amp may be replaced by a 748 or 301 using a 30-pF compensation capacitor.
- Reducing compensation capacitor of 748 or 301 to 3.3pF typically makes f_0 predictable to within $\pm 5\%$ of theoretical value up to about 22kHz.



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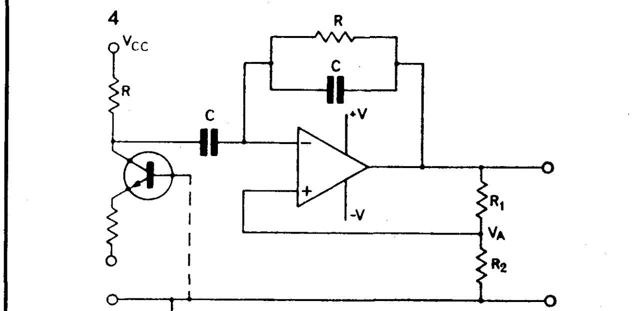


Circuit modifications

- Fig. 3 shows the general form of the circuit, one version of which has been discussed. Other configurations are possible by interchanging Z_1 and Z_4 and/or Z_2 and Z_3. Note that the output may not then be taken from the op-amp output.
- Since $Z_{in} \propto 1/Q$ for high Q a buffer amplifier may be added at the input. Buffer at the output is only required for alternative versions (except at very high Q).
- R at the input of the circuit may be taken to ground and from a current source e.g. from collector of a common-base stage–Fig. 4–or a cascode stage.

Further reading

1. Williams, P., Bandpass filters using Wien's bridge, *Electronics Letters*, 1970, vol. 6, p.186.
2. Kesner, D. An op. amp. RC bandpass filter, Motorola application note AN-452.

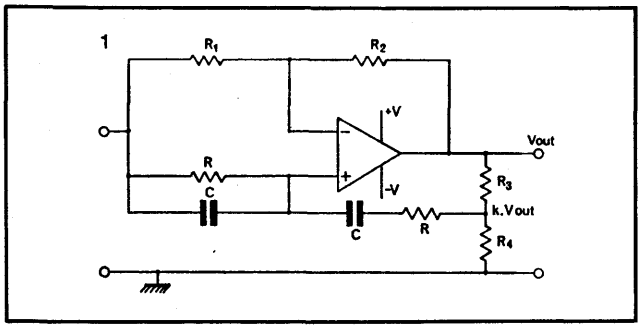
Cross references

Series 1, cards 3, 6, 7, 8, 11 & 12.

Wireless World Circard

Series 1: Basic active filters 2

Wien-bridge all-pass network



Typical performance at 1kHz

180° phase shift occurs at f_0, where

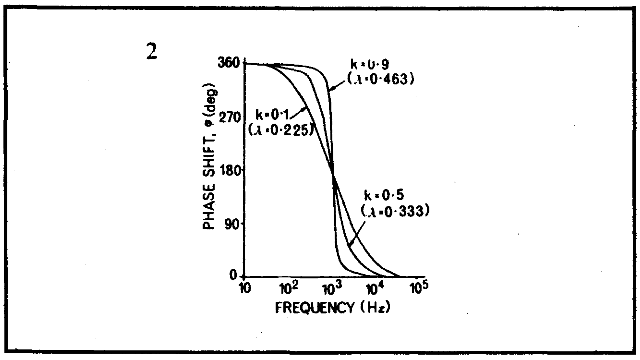
$$f_0 = \frac{1}{2\pi RC}\ \text{Hz}$$

$\lambda = R_1/R_2$

For unity voltage gain

$$k = \frac{5\lambda-1}{\lambda+1}$$

IC : 741; supplies ± 15V
$C = 1$nF $\pm 1\%$
$R = 150$kΩ $\pm 5\%$
$(R_3+R_4) = 1$kΩ
$(R_1+R_2) = 10$kΩ
Source resistance: 60Ω
With unity voltage gain V_{out}(max.) = 26V pk–pk.
No significant change in performance for supplies in the range ± 3 to ± 18V.

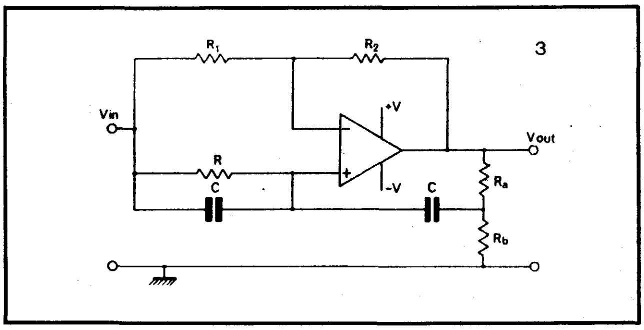


Circuit description

When λ has the correct value for a given value of k the network has unity voltage gain but the phase difference (φ) between output and input is frequency dependent. At $f_0 = 1/2\pi RC$ Hz, $\varphi = 180°$. Increasing k and adjusting λ to the appropriate value increases the magnitude of $d\varphi/df$ in the region of f_0. Suitable for use at audio frequencies.

Component changes

- 741 i.c. may be replaced with a 748 or a 301 using a 30-pF compensation capacitor without change in performance e.g. $d\varphi/df$ characteristic predictable up to $f_{0(max)}$ of about 10kHz, with unity voltage gain.
- Reducing compensation capacitor to 3.3pF extends range of predictable performance up to $f_{0(max)}$ of about 20kHz.
- With $R = 150$kΩ and $(R_1 + R_2) = 10$kΩ, $(R_3 + R_4)$ may be increased to approximately 22kΩ without significant change in performance.



Circuit modifications

- As in the bandpass filter (card 1) the arms of the bridge may be interchanged to produce alternative configurations having differing $d\varphi/df$ effects.
- For fixed $d\varphi/df$ and f_0, the arrangement in Fig. 3 may be used where R_a and R_b in parallel equals R and their ratio determines the value of K.
- For variable $d\varphi/df$ characteristic, without having to incorporate output potential divider resistance in CR calculations, insert a voltage follower between the potential divider and the series network.
- A buffer amplifier may be used at the input, though the input impedance is not likely to be as low as in the high-Q bandpass filter–card 1.

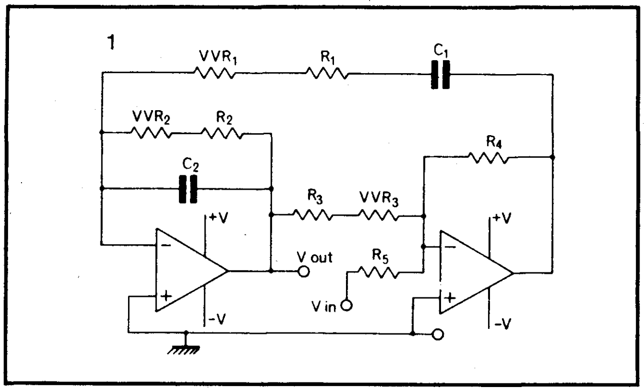
Further reading

1. Williams, P. Allpass networks using Wien's bridge *Electronics Letters*, vol. 6, 1970 p.183.
2. As 1: Alternative allpass networks using Wien's bridge, p.188.
3. Mitra, S. K., Analysis and synthesis of linear active networks, Wiley, 1969 pp.479–81.

Cross references

Series 1, card 1.

Voltage-controlled filters



Typical performance at 1kHz

$f_0 = \dfrac{1}{2\pi RC}$ Hz

where $R = (R_1 + VVR_1) = (R_2 + VVR_2)$

ICs 741; supplies $\pm$ 6V

VVRs: 1/4 × CD4016AE

$C_1 = C_2 = C = 0.1\mu \pm 2\%$

$R_1 = R_2 = 1\text{k}\Omega \pm 5\%$

$R_3 = 1.2\text{k}\Omega \pm 5\%$

$R_4 = 3.3\text{k}\Omega \pm 5\%$

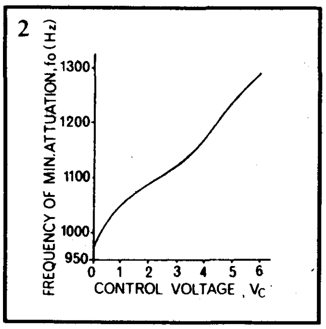
$R_5 = 10\text{k}\Omega \pm 5\%$

$V_{DD} = +3\text{V}$, $V_{SS} = -3\text{V}$

$V_C = 0$ to $+6\text{V}$

*VVR*s variable within the range approximately 680 to 280Ω.

$V_{C3} = 4.32\text{V}$ to produce $Q \approx 10$ when $V_{C1} = V_{C2} = 0\text{V}$.



Circuit description

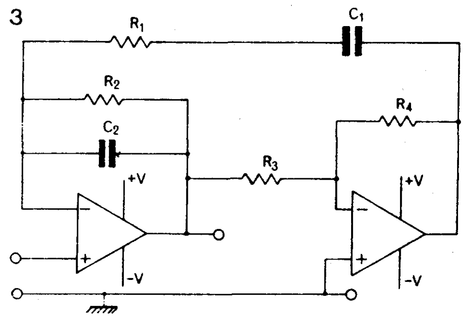
This bandpass circuit is related to that of card 1. Separation of the frequency-sensitive and resistive networks allows each section to have one terminal as a virtual earth. This simplifies the injection of signals (defined input resistance e.g. through R_5) and allows devices such as m.o.s. transistors to be used as frequency and/or Q control elements, with bias voltages referred to ground. The control elements used were complementary m.o.s. transmission gates with resistances controlled by the applied bias voltages. VVR_1 and VVR_2 control f_0 and VVR_3 controls Q by varying the ratio $R_4/(R_3 + VVR_3)$.

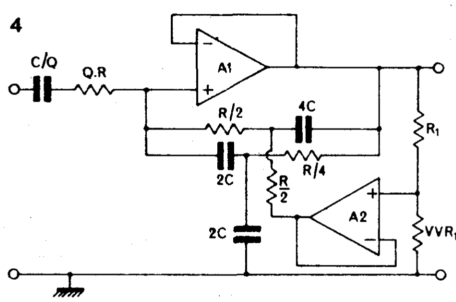
Component changes

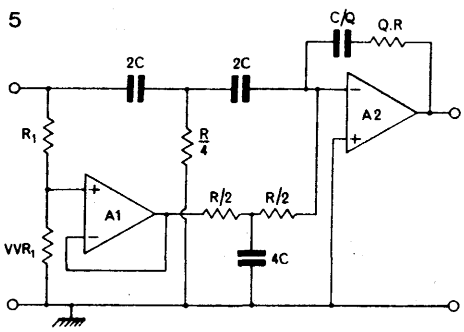
- Supplies may be varied within the range ± 3 to ± 18V and are conveniently chosen to be compatible with supplies required for the VVR elements. (N.B. CD4016AE max $V_{DD} - V_{SS} = 15\text{V}$)
- The greater the ratio VVR_n/R_n, $n = 1$ or 2, the wider the range of f_0 variation for given values of C within the audio band.
- For a given value of R_4, Q variation is greatest when $R_3 \rightarrow 0$.

Circuit modifications

- By scaling $R_1 = R_2/2$, $C_1 = 2C_2$ and setting $R_3 \rightarrow R_4$ the bandpass filter will provide outputs from the op-amps that are in antiphase and of almost equal amplitude.
- Any device with a fairly linear resistance, variable by an external parameter may replace the c.m.o.s. gates in this and other filters, (easiest where *R*s have a common point at or near ground potential, permitting this point to be connected to sources of f.e.ts, for example). Possibilities include CdS photocells, temperature-sensitive resistances like copper, platinum, and thermistors where signal swings do not change resistance, i.e. relatively high-power types.
- As in all virtual-earth amplifiers, the signal may be injected in voltage form into an otherwise-grounded, non-inverting terminal, as in Fig. 3, providing a high input resistance may alter filter characteristic.
- Figs. 4 and 5 show the voltage control principle applied to a parallel-T bandpass and notch filter respectively. In each circuit f_0 is controlled by the VVR.







In Fig. 4: $f_0 = \dfrac{1}{2\pi RC} \cdot \sqrt{\dfrac{R_1}{R_1 + VVR_1}}$ Hz.

In Fig. 5: $f_0 = \dfrac{1}{2\pi RC} \cdot \sqrt{\dfrac{VVR_1}{R_1 + VVR_1}}$ Hz.

Further reading

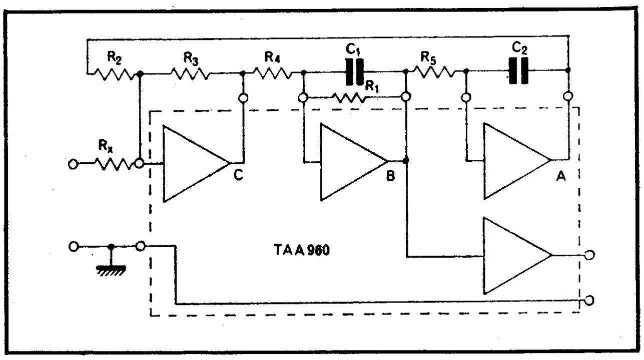
1. Good, E. F., Selective amplifiers with parallel-T feedback. *Electronic Engineering*, 1963 vol. 35, p.330.
2. Douce, J. L. and Edwards, K. H., Simple null filter with variable notch frequency, *Electronic Engineering*, 1964 vol. 36, pp.478/9.
3. Penn, T. C., Simple tuneable RC null networks, *Electronic Engineering*, 1964, vol. 36, p.849.
4. RCA Databook SSD-203, 1972, p.67.

Cross references

Series 1, cards 1, 6–10.

Wireless World Circard

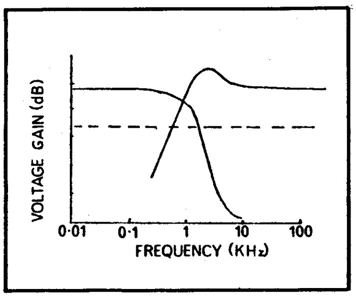
Low, high & band-pass triple amplifier



Typical performance

$$f_0 = \frac{1}{2\pi\sqrt{C_1C_2R_4R_5}}$$

Drive: 400mV pk–pk.
Supply: +6V.
R_2, R_3, R_4, R_5 = 10kΩ ±5%
C_1, C_2 = 10nF ±1%
R_X=470kΩ ±5%
$R_1 = \infty$
f_0=1,583Hz Q=49.5



Circuit description

Three inverting amplifiers form the filter. A fourth amplifier buffers the output. There are two integrators, one damped by resistor R_1, and a unity-gain inverting amplifier. A different filter function is obtained at the output of each amplifier; low-pass at A, bandpass at B and modified high-pass at C. Where all four amplifiers are included in a single i.c. as above, a restricted performance is offered by each amplifier individually, e.g. lower voltage gain, and the output resistance of a value that requires buffering if load resistances are not to change the filter characteristic. With R_1 absent, the Q of the band-pass filter is approximately 50. As R_1 is decreased, the response at A and C approach that usually required for second order, low-pass and high-pass responses respectively.
Single-supply operation and direct coupling place a lower limit on R_x of approximately 10kΩ, with $R_x = 50R_2$ for correct biasing. In addition, all outputs have a d.c. content.

Component changes

Supply voltage + 6V: f_0 = 1583Hz, gain = 1.15, Q = 49.5.
Supply voltage + 5V: f_0 = 1562Hz, gain = 0.95, Q = 40.
For best Q values, capacitors should be matched within 1%
For low-pass and modified high-pass filters R_1 = 6.8kΩ,
$Q = R_1/R_4 = 0.68$
Drive signal 2V pk-pk.
Output approx. 50mV pk–pk.

Circuit modifications

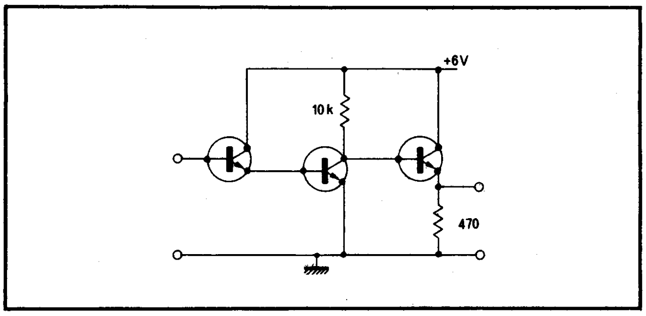
- As the three amplifiers are identical, the buffered amplifier may be used as the output for the low-pass or modified high-pass filter.
- For higher gain without disturbing d.c. conditions, input and output would have to be a.c.-coupled, then the input resistor could be reduced.
- The simple structure of the amplifiers can be duplicated by discrete circuits such as the amplifier with emitter-follower output shown.

Further reading

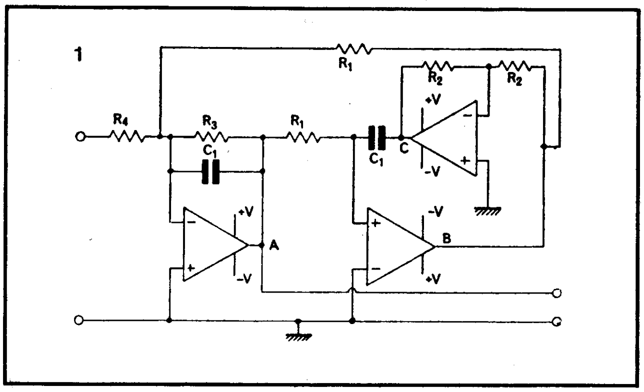
Mullard Ltd.: Triple amplifier for active filters, TAA 960.

Cross references

Series 1, cards 1, 3, 7, 8, 11 & 12.



Op-amp triple with phase compensation



Typical data

$R_1 = 3.3\text{k}\Omega \pm 5\%$
$C_1 = 0.1\mu\text{F} \pm 10\%$
$R_3 = 330\text{k}\Omega \pm 5\%$
$f_0 = \frac{1}{2\pi R_1 C_1} = 482\text{Hz}$

$Q_0 = R_3/R_1 = 100$
$f_0(\text{obs}) = 470\text{Hz}$
$Q(\text{obs}) = 105$
$V_{out}/V_{in} \approx Q$ at $f = f_0$

Circuit description

The same passive components as in card 6 may be used with separate op-amps having much higher gain and more suitable input and output characteristics. With the same amplifier configuration high Q is possible with excellent stability. At high frequencies the observed Q rises because of amplifier phase shifts. The alternative configuration shown provides some degree of cancellation of phase shifts allowing high-Q designs up to 100kHz with low-cost amplifiers. Precautions against over driving under these conditions are necessary as slew-rate limiting changes the amplifier response leading to the possibility of sustained oscillation.

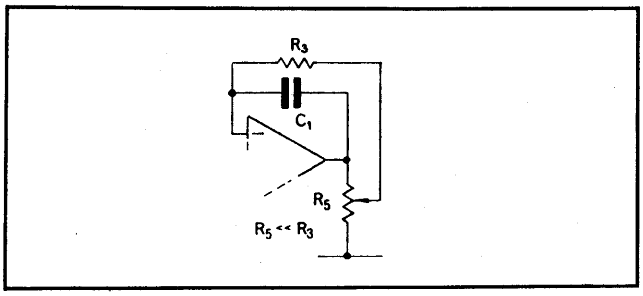
Component changes

R_1 = 470Ω to 1MΩ
R_3 = 470Ω to ∞
C_1 = 330pF to 10μF (non-polarized)
Amplifiers: N5741V, SN72741P, MC1741, etc.
Supply ±5 to ±18V (±3V in some cases).

Alternative amplifiers: any op-amp compensated for integrator operation.
f_0 = < 1Hz to > 100kHz
Q < 1 to > 100

Circuit modifications

● Variable damping at high Q inconvenient because of large value for R_3. Replace by fixed resistor fed from tapping across amplifier as shown e.g. R_5 = 10kΩ pot., R_3 = 100kΩ.
● Summing outputs at three amplifiers with conventional virtual earth summer gives more general transfer function.
● Low-pass output available at C (normally used with Q = 0·7 for simple low-pass filters as in audio applications—higher Q gives peak in response just below cut-off frequency).
● Modified high-pass output at B—comments as above.
● Where configuration as in card 6 is used phase lead may be introduced by a small capacitor across input resistor of invertor-alternative method of neutralizing phase-shift.



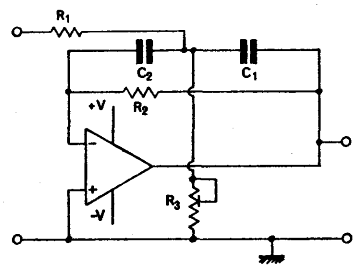
Further reading

Kerwin, W. J., Huelsman, L. P., and Newcomb, R. W. State-variable synthesis for insensitive integrated circuit transfer functions, *IEEE J. Solid-State Circuits*, vol. SC-2, 1967 pp.87–92.
Vogel, P. W. Method for phase correction in active RC circuits using two integrators, *Electronic Letters* 1971, pp.273-5.
Girling, F. E. J. and Good, E. F., Active filters-7 The two-integrator loop, *Wireless World*, vol. 77, 1970 pp.76–80.

Cross references

Series 1, cards 1, 3, 6, 8, 11 & 12.

Multi-feedback filter



Typical performance

For $C_1 = C_2 = C$

$$f_0 = \frac{1}{2\pi C}\sqrt{\frac{R_1 + R_3}{R_1 R_2 R_3}}$$

$$Q = \tfrac{1}{2}\sqrt{\frac{R_2 (R_1 + R_3)}{R_1 R_3}}$$

$A_0 = R_2/2R_1$
$C = 0.01\mu F \pm 10\%$
$R_1 = R_3 = 1.5k\Omega \pm 5\%$
$R_2 = 470k\Omega \pm 5\%$
Amplifier: 741
f_0(obs) = 845Hz
Q(obs) = 11.3

Circuit description
The inverting amplifier has multipath feedback which has a minimum at a single frequency. The overall response is then band-pass peaking at that frequency, which may be changed by varying C_1, C_2 together, the circuit Q remaining constant. Changing R_3 also varies the centre frequency keeping the bandwidth constant.

Component changes
R_2: 1.5kΩ to 1MΩ
R_1, R_3: 1kΩ to 100kΩ
C_1, C_2: 150pF to 10μF
Q = 1 to 50 (At high Q values, sensitivity to component variations is excessive)
f_0 < 1Hz to > 100kHz

Circuit modifications

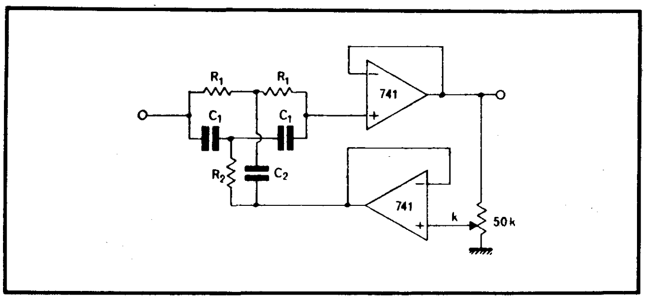
- Varying R_3 changes the centre frequency, leaving bandwidth and centre-frequency gain unchanged.
- Feeding from a current source, R_1 can be omitted, readjusting R_3 to give required characteristic. Alternatively if preceding stage has specified output resistance it can be incorporated into R_1.
- For low-gain low-Q applications the amplifier may be replaced by a single transistor in the common-emitter mode. R_2 provides base-current. Typical collector load resistances in range 1 to 10kΩ for passive network given. Alternatives include Darlington-pair amplifiers.
- Alternative passive networks for low and high-pass characteristics given in the reference.

Further reading
Graeme, J. G., Tobey, G. E., Huelsman, L. P., Operational amplifiers: design and applications, McGraw-Hill, pp.287–95.

Cross references
Series 1, cards 1, 3, 6, 7, 8, 11 & 12.

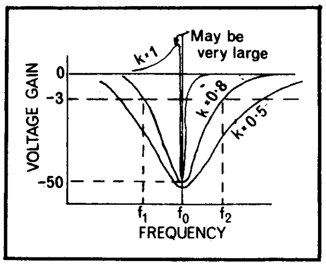
Wireless World Circard

Adjustable-Q twin-T notch filter



Typical performance

$f_0 = \dfrac{\sqrt{n}}{2\pi R_1 C_1}$
where $n = 2C_1/C_2 = R_1/2R_2$ (=1 usually)
$f_0 = 1000$Hz with
$C_1 = 10$nF,
$C_2 = 20$nF (trimmed)
$R_1 = 16.2$kΩ
$R_2 = 8.1$kΩ (trimmed)
$f_1 = 680$Hz $f_2 = 1650$Hz
$f_0 = 50$Hz with
$C_1 = 10$nF, $C_2 = 20$nF (trimmed)
$R_1 = 324$kΩ
$R_2 = 162$kΩ (trimmed)
$f_1 = 33$Hz $f_2 = 76$Hz

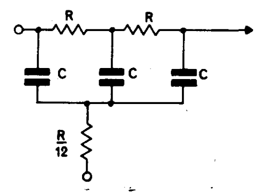


Circuit description

The passive network has a perfect null at a defined frequency if the components are accurately matched. The rate of approach to that notch is sharpened by positive feedback, the buffer amplifier having a negligible output resistance and hence theoretically not disturbing the depth of the notch. Variation in notch frequency requires simultaneous variation e.g. of all three capacitors or of R_2 and C_2 and the circuit is most suitable for fixed frequency operation. In addition too much positive feedback ($k \to 1$) may give unsatisfactory results.

Component changes

With low values of C_1 gain at high frequencies may not be 0dB. A 12% deviation from the nominal 0dB was observed with $C_1 = 270$pF, $R_1 = 600$kΩ, $n = 1$, $f_0 = 1$kHz. Very large resistors (> 10MΩ) should be avoided.



$$f_0 = \frac{\sqrt{3}}{2\pi RC}$$

Circuit modifications

- Any circuit with a notch e.g. above may be used in place of the twin-T network e.g. ref. 1. Trimming of $R/12$ may be necessary if k is varied over a wide range.
- Notch filters having a narrow notch width can be unsatisfactory with signals whose frequency stability is not good. This can be overcome by cascading two notch filters, having slightly different notch frequencies e.g. two filters tuned to 48 and 52Hz respectively will effectively remove mains pick-up[2]. It may be possible to simply cascade the passive networks with due attention to loading of the first by the second, and still only use two op-amps.
- Single component variation of the notch frequency is possible with a twin-T network (card 3 and ref. 3) and is also possible with other networks (card 10). Buffer amplifier may be omitted if potentiometer value is low.

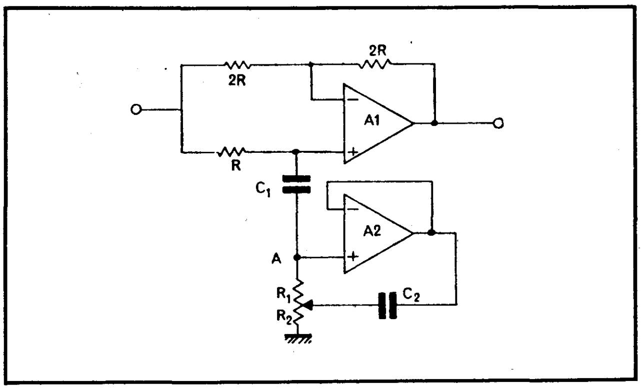
Further reading

1. Graf. Electronic Design Data Book, Van Nostrand Reinhold.
2. N. B. Rowe, Designing a low frequency active notch filter, *Electronic Engineering*, April 1972.
3. Douce and Edwards, *Electronic Engineering*, July 1964.
4. National Semiconductor Linear Brief LB-5.

Cross references

Series 1, cards 3 & 10.

Easily-tuned notch filter



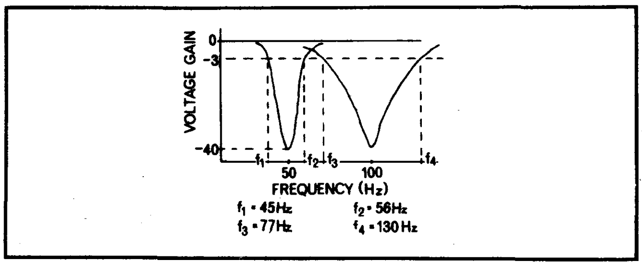
Typical performance

Q of resonant branch is

$$\frac{1}{R_1+R_2}\cdot\sqrt{\frac{C_2R_1R_2}{C_1}}$$

$f_0=1/2\pi\sqrt{C_1C_2R_1R_2}$
$f_0=50$Hz with
R_1, R_2 : 50kΩ, C_2 : 1μF
C_1 : 4nF R : 100kΩ

Results shown are for these values, with some trimming of one of the $2R$ resistors for good notch depth. 741 op-amps were used. Considerable trimming is necessary at higher frequencies.



Circuit description

The impedance between A and ground is equivalent to an inductor, $C_2R_1R_2$, in series with a resistor, R_1+R_2. At a specific frequency C_1 and the equivalent inductance series resonate, leaving the equivalent resistance to bring the bridge into balance provided $(R_1 + R_2) = R$ i.e. an overall notch characteristic. Variation in the equivalent inductance or in C_1 changes the notch frequency with no theoretical change in notch depth. In practice component imperfections, circuit strays etc. may require significant departures from bridge nominal resistances to reach balance, particularly at frequencies above 100Hz. If the circuit Q is large, A2 will saturate at the notch frequency unless V_{in} is kept low. At frequencies of the order of 1kHz or more the notch depth and Q are sensitive to resistance in series with C_2. A low-loss capacitor and low-contact-resistance potentiometer should then be used (avoid potentiometers with non-metallic contacts).

Component variations

R: 50kΩ to 1MΩ
C_2: 10μF to 0.1μF
C_1: 40 nF to 100pF.

Circuit modification

R_1 and R_2 may be kept as fixed resistors if a variable capacitor for C_1 is available.

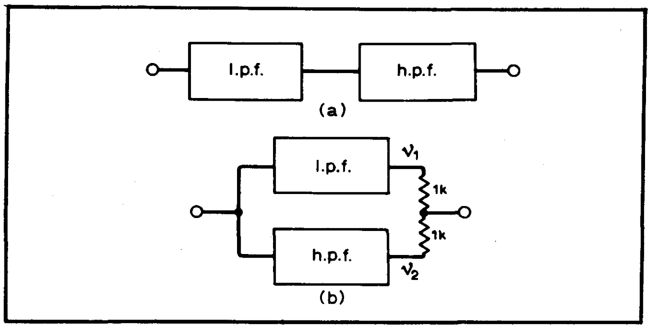
Further reading

National Semiconductor LM307 Data Sheet.

Cross references

Series 1, card 9.

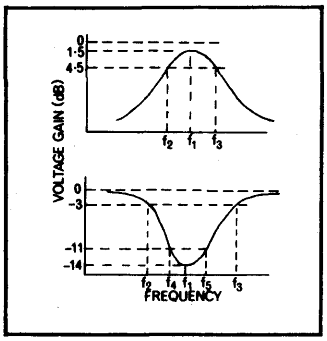
Compound filters



(a) provides the band-pass characteristic shown. Low and high-pass filters used are described in cards 4 and 5 with cut-off frequencies 10kHz and 2.35kHz respectively. $f_1 = 5$kHz, $f_2 = 1.9$kHz and $f_3 = 13$kHz.

(b) provides the band-stop characteristic shown. In this case the low and high-pass filters had cut-off frequencies 0.76kHz and 12kHz respectively.

$v_0 = (v_1 + v_2)/2$,
$f_1 = 3$kHz, $f_2 = 0.76$kHz,
$f_3 = 12$kHz,
$f_4 = 2.1$kHz and
$f_5 = 4.5$kHz.



Circuit description

The basic filter functions described may be combined in a variety of ways to produce alternative functions or to improve on existing ones. The two examples shown illustrate ways of producing simple band-pass and band-stop filters from low-pass and high-pass sections. Cascading l.p. and h.p. filters with the l.p. filter cut-off frequency above that of the h.p. filter gives a pass-band lying between these two frequencies, and for second-order filters, an attenuation approaching 12dB/octave outside this range. Multi-order l.p. and h.p. filters may be used to increase this attenuation.

Conversely if the filter responses are added with the l.p. filter cut-off frequency being the lower the response is minimum between the cut-off frequencies with the depth of the attenuation dependent on the separation. Simple resistor summing, or a virtual earth summing amplifier may be used.

Circuit modifications

One can cascade second-order low-pass and high-pass sections to achieve steeper sides to the stop and passbands. This only produces even-order filters with poor roll-off characteristics. Improved higher-order filters can easily be used (see ref.).

One can also cascade the bandstop characteristic with a notch filter to achieve precise nulling of a fixed frequency with reasonable attenuation on either side of the notch. Several notch filters tuned to slightly different frequencies may be used to obtain a bandstop characteristic (card 9). Several band-pass filters (card 1) can be similarly cascaded to give a good bandpass characteristic.

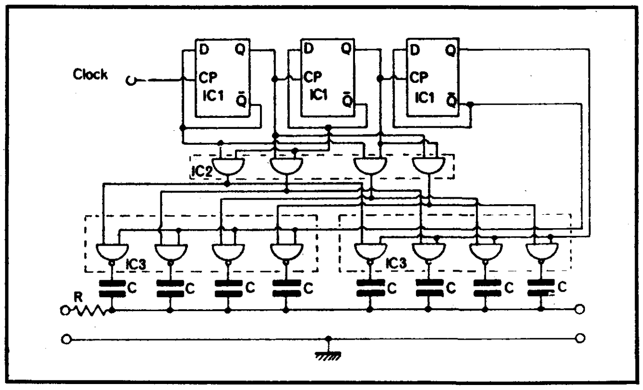
Component changes

Any low-pass filter or high-pass filter may be used in the above configurations to provide the band-stop and band-pass characteristics.

Further reading

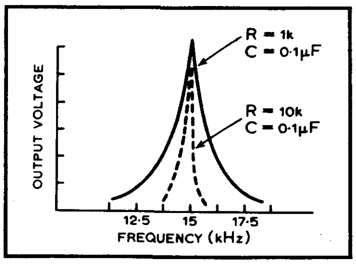
Graeme. J. G., Tobey, G. E. & Huelsman. L. P. Operational amplifiers. McGraw-Hill p.320.

N-path filter



Typical performance

IC_1 $\frac{1}{2}$ SN7474 D-type flip flop
IC_2 2 off SN7400 connected as two-input AND gates
IC_3 SN7401 quadruple two-input NAND gates with open collector output.
R: 1kΩ $\pm 1\%$
C: 0.1μF$\pm 10\%$
f_0 = 14.9kHz Q = 30
Drive signal 600mV pk–pk.
Resonant frequency f_0 = clock frequency/N
Bandwidth: $2/NRC$ Hz,
N = number of low-pass filter sections = 8.
Peak on graph: 460mV pk–pk.



Circuit description

The input signal is effectively switched between eight low-pass filter sections at a clock rate, eight times the required filter centre frequency. The output waveform comprises discrete levels approximating the input waveform. This stepped format may be removed by a low-Q bandpass filter.
The resonant frequency can be varied simply by changing the clock frequency thus giving a tunable bandpass filter with constant bandwidth.

Component changes

Increasing R for same C values, increases Q e.g. R = 10kΩ $\pm 10\%$, Q = 200.
IC_1 $\frac{1}{2}$MC7479; IC_2 MC3001; IC_3 MC7403.

Circuit modifications

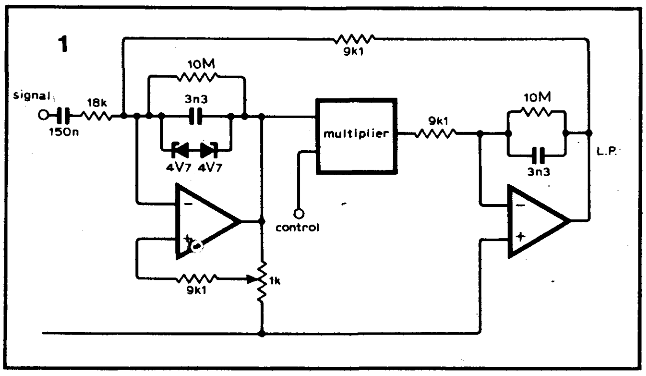
Due to the sampling action of this filter, responses are obtained at multiples of the clock frequency. If a 4-path filter with a fundamental resonant frequency f_0 is cascaded with a 3-path filter with a fundamental resonant frequency $2f_0$, a single bandpass characteristic is obtained, centred on $2f_0$ (ref. 2).

Further reading

1. Broeker, B. Commutating filter techniques *Design Electronics*, 1971, pp.52–5.
2. Roberts, J. N-path filter techniques *Electron*, June 1972, pp.18–9.
3. Bruton, L. T. and Pederson, R. T. Time-multiplexed active filters, *IEEE Journal of Solid-State Circuits*, vol. SC-7, June 1972, pp.259–65.

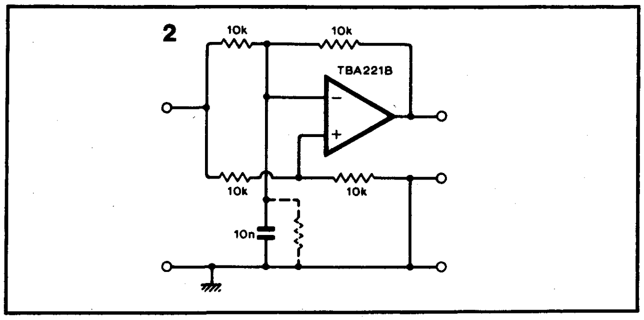
Basic active filters

1. Voltage control of active filters is achieved using an analogue multiplier (see Audio circuits, set 5, for details). A loop consisting of a bandpass filter and an integrator has the signal coupled between them by the multiplier acting as a voltage controlled amplifier. The reference article also shows a square-law generator for shaping the frequency/control voltage characteristics. In this configuration the 1-kΩ potentiometer sets the circuit Q, and the direct control voltage the centre frequency. A low-pass output is also available. The zener diodes prevent overdrive from latching the circuit into a permanently switched state. The article also indicates how the circuit can be modified into an RC oscillator with stabilized amplitude, and a notch filter.
Orr, T. and Thomas, D. W. Electronic sound synthesizer, part 2. *Wireless World*, vol. 79, 1973, pp.429–34.



2. The limited amplitude-frequency response of operational amplifiers restricts the upper frequency at which defined filter characteristics can be obtained; it may result in instability/oscillation in some filters. As many op-amps, such as that shown here (and also 741, 307 etc), have a frequency response defined by a single dominant lag internal to the amplifier, then some second-order filters are possible using just one external capacitor. The reference article shows that the resulting transfer function is that of a band-pass filter with centre-frequency $\approx$ 31kHz, centre-frequency gain $\approx$ 115 and $Q \approx 11.5$. Adding a resistor across the capacitor controls Q and gain without varying the centre frequency. The filter characteristics are supply and temperature dependent – the former can be turned to advantage in trimming centre frequency if variation in the other parameters is acceptable.

Radhakrishna Rao, K and Srinivasan, S. "Bandpass filter using the operational amplifier pole", *IEEE Journal of Solid-State Circuits*, June 1973, pp.245–6.



3. The design of higher-order filters can be based on cascaded first/second-order sections. Alternative approaches, economical of active and passive elements, use interconnected networks of impedance convertors (n.i.cs, gyrators etc) and passive networks. All the parameter values interact but the approach is very efficient and capable of very low sensitivity to component variation. The example shown uses blocks within dotted lines that act as impedance converters to simulate lossy indicators, and results in a third-order low-pass elliptic filter with 1.0dB ripple in the passband, 30dB loss in the stopband and with a passband edge frequency of 3.4kHz. Other networks are shown in the article generating higher-order filters.
Wise, D. R. Active simulation of floating lossy inductances, *Proc. IEEE*, vol. 121, 1974 pp.85–7.