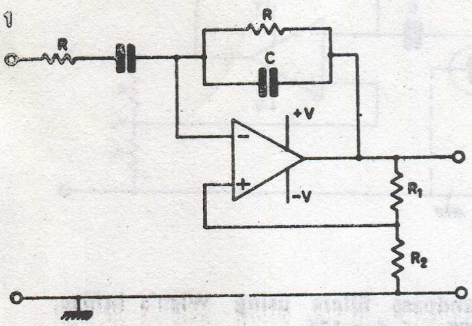
Wireless World Circard

Series 1: Basic active filters–1

Wien-bridge bandpass filter

1



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

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

Voltage gain, $\frac{V_{out}}{V_{in}} = 2Q$

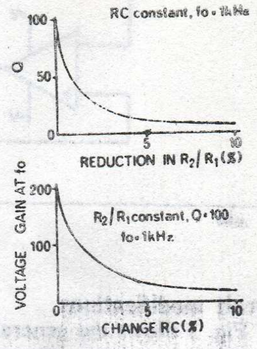
To find Q measure voltage gain and divide by 2.

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.

Typical performance at 1kHz

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



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.

Wireless World Circard

Series 1: Basic active filters–2

Wien-bridge all-pass network

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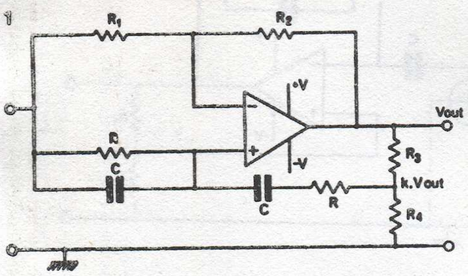


180° phase shift occurs at f_0, where

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

$$\lambda = R_1/R_2$$

For unity voltage gain

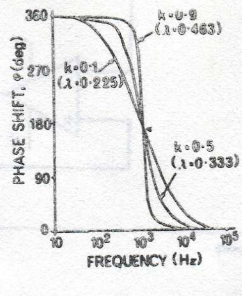
$$k = \frac{5\lambda - 1}{\lambda + 1}$$

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.

Typical performance at 1kHz

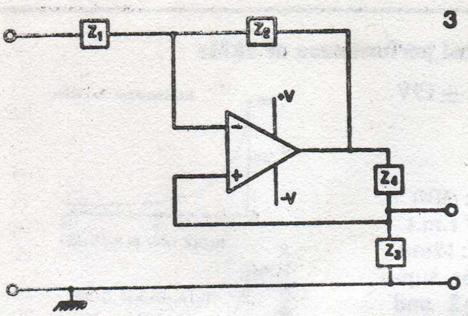
IC: 741; supplies ± 15V
$C = 1$nF$\pm 1\%$
$R = 150$k$\Omega \pm 5\%$
$(R_3+R_4) = 1$kΩ
$(R_1+R_2) = 10$kΩ
Source resistance: 60Ω
With unity voltage gain V_{out}(max.) = 26V pk–pk.
No significant change in performance for supplies in the range ± 3 to ± 18V.



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.

3



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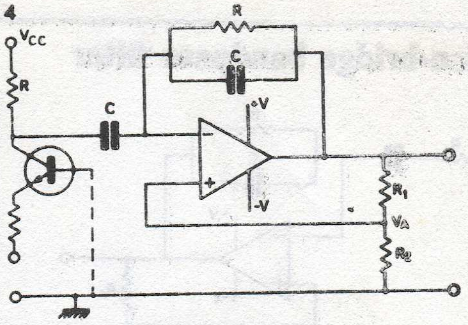


Circuit modifications

- 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.

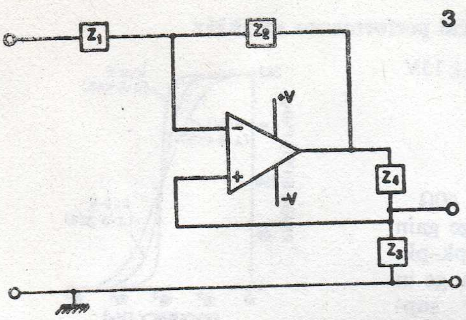
Further reading

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.

3



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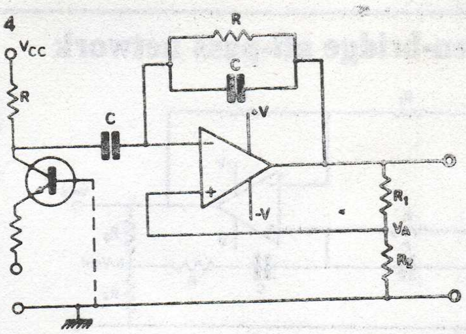

Circuit modifications

- 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.

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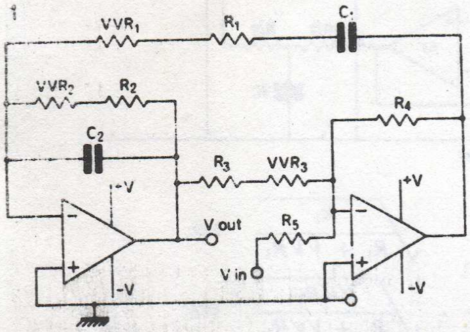
Wireless World Circard

Series 1: Basic active filters–3

Voltage-controlled filters



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

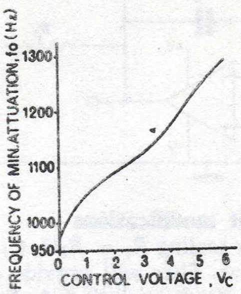
where $R = (R_1 + VVR_1) = (R_2 + VVR_2)$
ICs 741; supplies $\pm$ 6V
VVRs: $1/4 \times$ CD4016AE
$C_1 = C_2 = C = 0.1\mu \pm 2\%$
$R_1 = R_2 = 1\text{k}\Omega \pm 5\%$
$R_3 = 1.2\text{k}\Omega \pm 5\%$
$R_4 = 3.3\text{k}\Omega \pm 5\%$
$R_5 = 10\text{k}\Omega \pm 5\%$

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. VVR₁ and VVR₂ control f_0 and VVR₃ controls Q by varying the ratio $R_4/(R_3 + VVR_3)$.

Typical performance at 1kHz

$V_{DD} - V_{SS} = +6$V
$V_C = 0$ to $+6$V
VVRs variable within the range approximately 680 to 280Ω.
$V_{C3} = 4.32$V to produce $Q \approx 10$ when $V_{C1} = V_{C2} = 0$V.



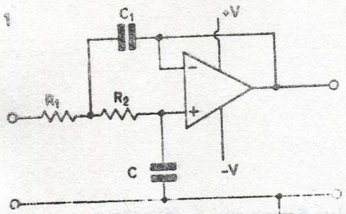
Component changes

- Supplies may be varied within the range ± 3 to ± 18V and are conveniently chosen to be compatible with supplies required for the VVR elements.
- 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 $R_3 \to 0$.

Wireless World Circard

Series 1: Basic active filters–4

Low-pass Sallen & Key filter



$f_0 = \dfrac{1}{2\pi\sqrt{C_1C_2R_1R_2}}$ Hz

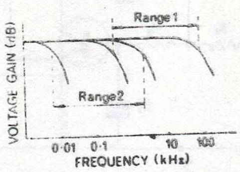
MC 1741 supply: ± 15V
Drive: 1V r.m.s.
$R_1, R_2 (=R)$: $22\text{k}\Omega \pm 5\%$
$C_1 (=2C_2)$: $940\text{pF} \pm 1\%$
$f_0 = 10$kHz
For C_2/C_1 constant, $f_0 \propto 1/R$

Circuit description

An easy and effective second-order filter. Any unity gain amplifier and standard operational amplifiers with 100% negative feedback are ideal at frequencies in the audio band. I.C. voltage followers allow frequency range to be extended well beyond 100kHz, their high input resistance also allowing large R values for low cut-off frequencies. Ratio of R_1/R_2 and C_1/C_2 must be kept constant for fixed filter characteristics. For audio purposes, tolerances are not critical and the resistors may be replaced by a twin-gang potentiometer for a wide range variable-cut-off filter. Voltage gain is unity within the passband.

Typical performance

At high f_0 ($\to$100kHz), distortion minimized with R_1, $R_2 = 6.8\text{k}\Omega$ (min.) and $C_1 = 2C_2 = 300$pF. For designed values of $f_0 = 100$kHz, actual may be between 50 to 90kHz using 741 op-amp (see circuit modifications).



Component changes

f_0 maintained for supply variations from ± 6 to ± 18V (this minimum is dependent on signal drive). Use LM310 for wide range filter using twin-gang potentiometer. See graph for typical performance. Range 1: 5 to 1200Hz, $C_1 = 2C_2 = 0.1\mu$F, $R_1 = R_2 = 2.2$ to 500 kΩ. Range 2: 400Hz to 94kHz, $C_1 = 2C_2 = 940$ pF, $R_1 = R_2 = 2.2$ to 500kΩ.

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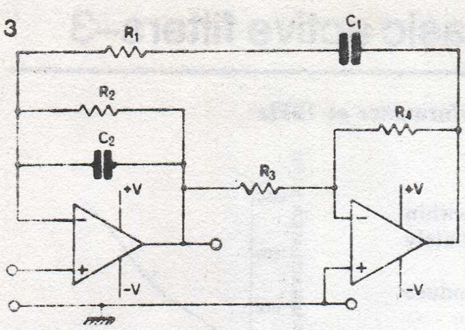

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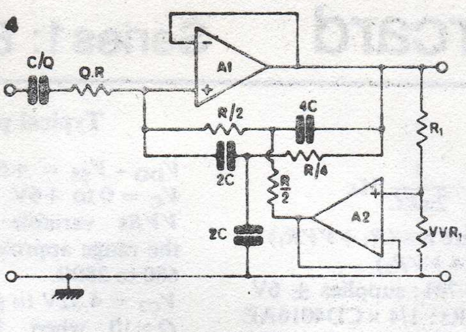

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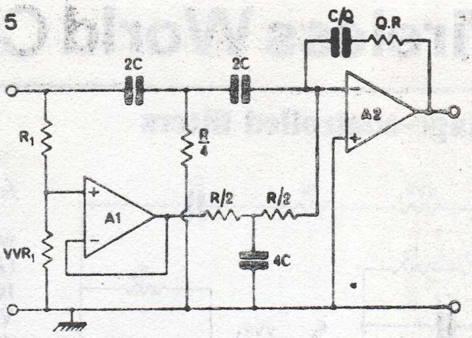

Circuit modifications

● 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.

● 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.

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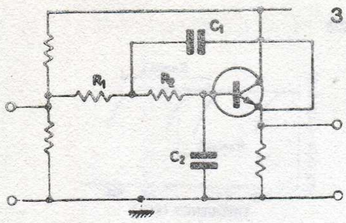


Circuit modifications

The op-amp can be replaced by any other unity gain amplifier including emitter follower (Fig. 3), Darlington pair, i.c. voltage follower, source follower (j.f.e.t. or i.g.f.e.t.) etc.

Restrictions on performance:

●Emitter follower—input resistance is finite and thus loads network; output resistance finite and is loaded by network. Accurate transfer function difficult, but often o.k. for audio amplifiers.

●Darlington pair—one stage better, but worse than emitter follower in respect of d.c. offset. In both cases a bias network is necessary. R-C coupling at input and output is possible i.e. low-pass does not extend to d.c.

●Voltage follower such as LM310 is best—for large signal swings at high frequency, into capacitive loads, increased bias from 1kΩ between booster terminal and negative supply minimizes distortion.

Further reading

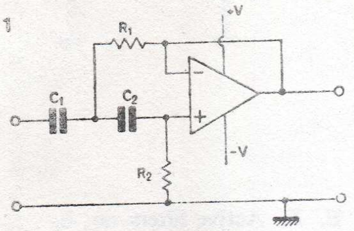
Girling, F. E. J., and Good, E. F. Active filters no. 6, *Wireless World*, vol. 76, January 1970, pp.27–31.
LM310 voltage follower. National Semiconductor Corporation, 1970.
Sallen, R. P. and Key, E. L., Practical method of designing RC active filters, *IRE Trans.* vol. CT-2, 1955. pp.74–85.

Cross references
Series 1, cards 6 & 7.

Wireless World Circard

Series 1: Basic active filters–5

High-pass Sallen & Key filter



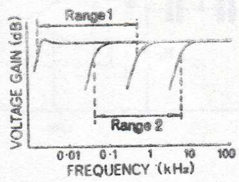
Cut-off frequency

$$f_0 = \frac{1}{2\pi\sqrt{C_1C_2R_1R_2}}$$

Typical values:
MC 1741 supply ±15V
Drive voltage 2V pk–pk.

Typical performance

$R_1 = R_2 = 100k\Omega \pm 5\%$
$C_1 = 2C_2 = 0.02\mu F \pm 10\%$
$f_0 = 110Hz$
f_0 maintained for supply voltage changes from ±6 to ±18V.



Circuit description

High-pass filter is obtained from low-pass unit by interchanging capacitors and resistors. Gain is unity at high frequencies (reactance of capacitors tending to zero) falling at 12dB/octave below the cut-off frequency f_0. The response near the cut-off frequency can be modified by changing C_1/C_2 or R_1/R_2 ratios. Cut-off frequency depends on the CR product values. Amplifier should have unity voltage gain, high input and low output impedance, and wide bandwidth e.g. op-amps connected as voltage followers or specially designed i.c. voltage followers like the LM310.

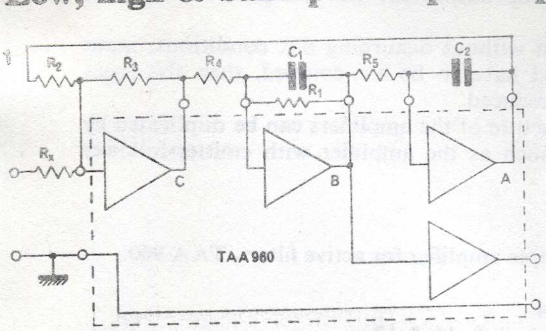
Component changes

$R_1 = R_2$: 3.9kΩ to 1MΩ
$C_1 = 2C_2$: 0.02 to 0.2μF
Amplifiers N5741V, LM310.
Wide-range filter: use LM310 with twin-gang potentiometer, connecting booster terminal via 1kΩ resistor to –ve supply. See graph for typical performance. Range 1:2.5 to 600Hz, $R_1 = R_2 = 2.2k\Omega$ to $500k\Omega$, $C_1 = C_2 = 0.2\mu F$. Range 2: 40Hz to 8.8kHz, $R_1 = R_2 = 2.2k\Omega$ to $500 k\Omega$, $C_1 = C_2 = 0.02\mu F$.

Wireless World Circard

Series 1: Basic active filters–6

Low, high & band-pass triple amplifier

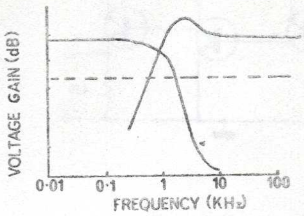


$$f_0 = \frac{1}{2\pi\sqrt{C_1C_2R_4R_5}}$$

Drive: 400mV pk–pk.
Supply: +6V.
R_2, R_3, R_4, R_5 = 10kΩ ±5%

Typical performance

$C_1, C_2 = 10nF \pm 1\%$
$R_X = 470k\Omega \pm 5\%$
$R_1 = \infty$
$f_0 = 1{,}583Hz \quad Q = 49.5$



Circuit description

Three inverting amplifiers form the filter. A fourth amplifier buffers the output. There are two integrators, one damped by resistor R₁, 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 R₁ absent, the Q of the band-pass filter is approximately 50. As R₁ 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 $R_1 = 6.8k\Omega$,

$Q = R_1/R_4 = 0.68$

Drive signal 2V pk-pk.
Output approx. 50mV pk–pk.

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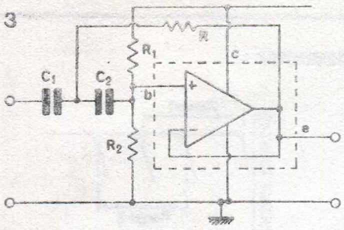
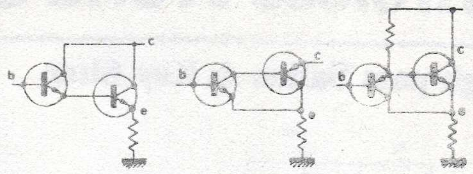




Circuit modifications

- Single-ended supply operation possible for both op-amps and transistors, replacing shunt resistor by potential divider with $R_1R_2/(R_1 + R_2) = R$ (or $2R$ or $R/2$ as required by filter characteristic). Fig. 3.
- Any of the multi-transistor equivalents of an emitter follower may be used, e.g. Figs. on right can be substituted for the op-amp within the dotted box in Fig. 3.
- Third-order filters may be constructed simply by adding a first-order passive R-C section at the output. The ratios of R_5 and C_5 in the original have to be re-adjusted if optimum performance is to be achieved near the cut-off point.

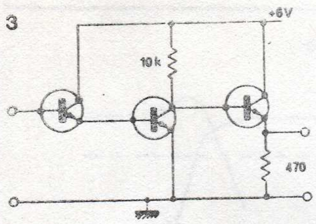
Further reading

1. Girling, F. E. J. and Good, E. F. Active filters no. 6, *Wireless World*, January 1970, pp.27–31.
2. Sallen, R. P. and Key, E. L., Practical method of designing RC active filters, *IRE Trans.* vol. CT-2, 1955 pp.74–85.
3. LM310 voltage follower. National Semiconductor Corporation, 1970.

Cross references

Series 1, cards 6 & 7.

3



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.

Wireless World Circard

Series 1: Basic active filters–7

Op-amp triple with phase compensation

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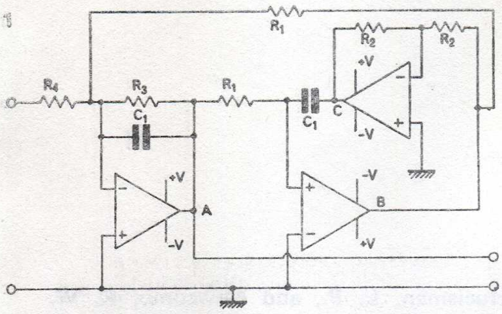


Typical data

$R_1 = 3.3\text{k}\Omega \pm 5\%$
$C_1 = 0.1\mu\text{F} \pm 10\%$
$R_3 = 330\text{k}\Omega \pm 5\%$

$$f_0 = \frac{1}{2\pi R_1 C_1} = 482\text{Hz}$$

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

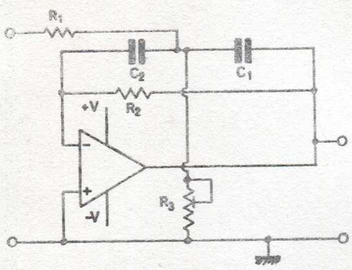
R_1 = 470Ω to 1MΩ
R_3 = 470Ω to ∞
C_1 = 330pF to 10μF (non-polarized)
Amplifiers: N5741V, SN72741P, MC1741, etc.
Supply ±5 to ±18V (±3V in some cases).
Alternative amplifiers: any op-amp compensated for integrator operation.

f_0 = < 1Hz to > 100kHz
Q < 1 to > 100

Wireless World Circard

Series 1: Basic active filters–8

Multi-feedback filter



For $C_1 = C_2 = C$

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

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

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

Circuit description

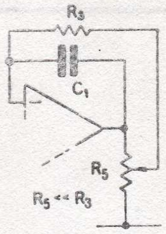
The inverting amplifier has multipath feedback which has a minimum at a single frequency. The overall response is then band-pass peaking at that frequency, which may be changed by varying C_1, C_2 together, the circuit Q remaining constant. Changing R_3 also varies the centre frequency keeping the bandwidth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

Component changes

R_1: 1.5kΩ to 1MΩ
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.



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. $R_5 = 10k\Omega$ pot., $R_3 = 100k\Omega$.
- Summing outputs at three amplifiers with conventional virtual earth summer gives more general transfer function.
- Low-pass output available at C (normally used with $Q = 0{\cdot}7$ 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.

Further reading

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

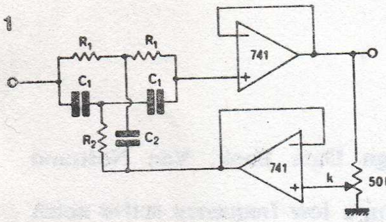
Cross references

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

Wireless World Circard

Series 1: Basic active filters–9

Adjustable-Q twin-T notch filter

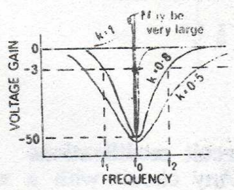


$$f_0 = \frac{\sqrt{n}}{2\pi R_1 C_1}$$

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

Typical performance

$R_2 = 8.1$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 ($k \rightarrow 1$) may give unsatisfactory results.

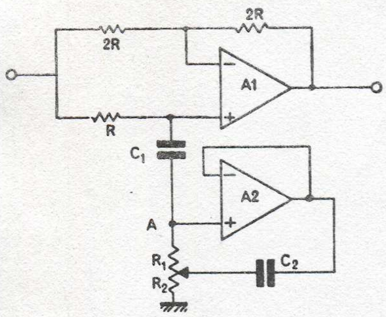
Component changes

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.

Wireless World Circard

Series 1: Basic active filters–10

Easily-tuned notch filter



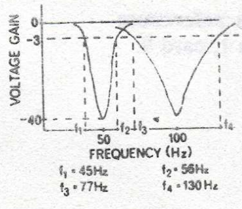
Q of resonant branch is

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

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

Typical performance

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

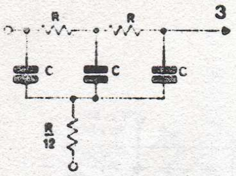
The impedance between A and ground is equivalent to an inductor, $C_2R_1R_2$, in series with a resistor, R_1+R_2. At a specific frequency C_1 and the equivalent inductance series resonate, leaving the equivalent resistance to bring the bridge into balance provided $(R_1 + R_2) = R$ i.e. an overall notch 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.



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

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 applications notes.

Cross references

Series 1, cards 3 & 10.

Further reading

National Semiconductor applications notes.

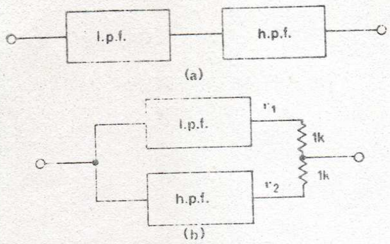
Cross references

Series 1, card 9.

Wireless World Circard

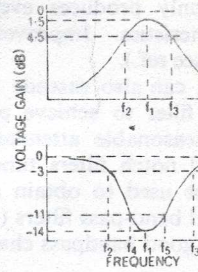
Series 1: Basic active filters–11

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, a 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.

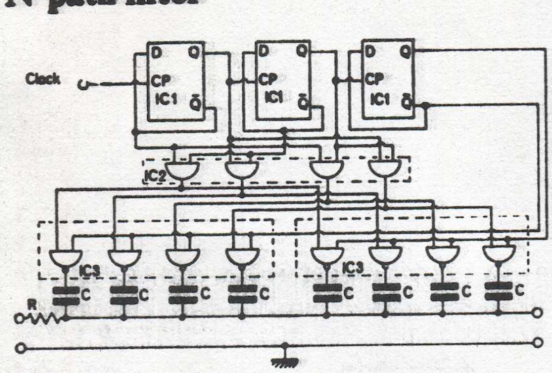
Component changes

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

Wireless World Circard

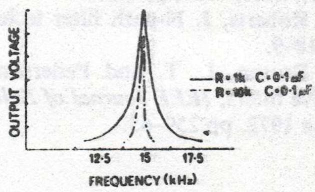
Series 1: Basic active filters–12

N-path filter



IC_1 ½ 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 ±1%

C: 0.1μF±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 ±10%, Q = 200.
IC_1 ½MC7479; 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).

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.

Further reading

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

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.