Wireless World Circard

Wien-bridge bandpass filter

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

\[ Q = \frac{1}{2 - (R_f / R_1)} \]

Voltage gain, \( V_{\text{out}} = 2Q V_{\text{in}} \)

To find \( Q \) measure voltage gain and divide by 2.

Circuit description
The Wien network—Fig. 1—gives a maximum response at \( f_0 = 1/2\pi RC \text{ Hz} \). 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 \( R_2 \) or both \( C_1 \), 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. Suitable for moderate \( Q \) values (5–25) at frequencies in the audio range. High \( Q \) is obtainable if only short-term stability is required.

Series 1: Basic active filters—1

Typical performance at 1kHz
IC: 741; supplies ±15V
\( C = 10\mu F \pm 1\% \)
\( R = 150k\Omega \pm 5\% \)
\( (R_2 + R_3) = 10k\Omega \)
Source resistance: 60Ω
\( V_{\text{out}} (\text{max.}) = 9-2V \text{ r.m.s.} \)
Max. load current: 12mA
\( Q \) is constant for supplies between ±2 and ±18V if clipping is avoided.
\( Z_{\text{in}} = 1/Q; Z_{\text{out}} = Q \).

Component changes
- Using ±1% capacitors and ±5% resistors, \( f_0 \) will typically be within ±5% of theoretical value up to approximately 15kHz and within ±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 < \( R \) for predictable \( f_0 \) and < \( R/15 \) for predictable \( Q \) within 5%.
- Circuit will oscillate when \( R_2/R_1 > 2 \).
- 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 ±5% of theoretical value up to about 22kHz.

Wireless World Circard

Wien-bridge all-pass network

\[ 180^\circ \text{ phase shift occurs at } f_0, \text{ where} \]

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

\[ \lambda = R_2 \]

\[ k = \frac{5\lambda - 1}{\lambda + 1} \]

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

Series 1: Basic active filters—2

Typical performance at 1kHz
IC: 741; supplies ±15V
\( C = 10\mu F \pm 1\% \)
\( R = 150k\Omega \pm 5\% \)
\( (R_1 + R_2) = 1k\Omega \)
\( (R_1 + R_2) = 10k\Omega \)
Source resistance: 60Ω
With unity voltage gain \( V_{\text{out}} (\text{max.}) = 26V \text{ pk-pk} \).
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 \( f_0(\text{max.}) \) of about 10kHz, with unity voltage gain.
- Reducing compensation capacitor to 3.3pF extends range of predictable performance up to \( f_0(\text{max.}) \) of about 20kHz.
- With \( R = 150k\Omega \) and \( (R_1 + R_2) = 10k\Omega \), \( (R_1 + R_2) \) may be increased to approximately 22kΩ without significant change in performance.
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

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

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Voltage-controlled filters

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

where \( R = \frac{R_1 + VVR_1}{R_2 + VVR_2} \)

ICs 741; supplies ±6V

VVRs: 1/4 \( \times \) CD4016AE

\( C_1 = C_2 = 0.1 \mu \text{F} \pm 2\% \)

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

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

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

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

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 \( R_0 \)) 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 \( R_0/(R_3 + VVR) \).

Typical performance at 1kHz

\( V_{DD} - V_{SS} = +6V \)

\( V_C = 0 \text{ to } +6V \)

\( V_{VIRS} \) variable within the range approximately 680 to 2800.

\( V_{C3} = 4.32V \) to produce

\( Q \approx 10 \) when \( V_{C1} \)

\( V_{C2} = 0V \).

Component changes

- Supplied 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 \( V_{VIRS}/R_n \), \( n = 1 \) or 2, the wider the range of \( f_0 \) variation for given values of \( C \) within the audio band.
- For a given value of \( R_3 \), \( Q \) variation is greatest when \( R_3 \rightarrow 0 \).

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Wireless World Circard

Low-pass Sallen & Key filter

\[ f_0 = \frac{1}{2\pi \sqrt{C_1 C_2 R_1 R_2}} \text{ Hz} \]

MC 1741 supply: ±15V

Drive: 1V r.m.s.

\( R_1 = R_2 = \frac{22k\Omega \pm 5\%}{2} \)

\( C_1 = 940p\text{F} \pm 1\% \)

\( f_0 = 10\text{kHz} \)

For \( C_3/C_1 \) 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 \( f_0 \) (→100kHz), distortion minimized with

\( R_1, R_2 = 6.8k\Omega \) (min.)

and \( C_1 = 2C_2 = 300p\text{F} \).

For designed values of \( f_0 = 100kHz \), 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, \( C_1 = 2C_2 = 0.1\mu\text{F} \), \( R_1 = R_2 = 2.2 \) to 500k\Omega. Range 2: 400Hz to 94kHz, \( C_1 = 2C_2 = 940p\text{F} \), \( R_1 = R_2 = 2.2 \) to 500k\Omega.
Circuit modifications

- By scaling \( R_1 = R_2/2 \), \( C_1 = 2C_2 \) and setting \( R_3 = R_4 \) 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.t.s, for example). Possibilities include Cds photocells, temperature-sensitive resistances like copper, platinum, and thermistors where signal swing does 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: \[ f_0 = \frac{1}{2\pi RC} \sqrt{\frac{R_1}{R_1 + VVR}} \] Hz.

In Fig. 5: \[ f_0 = \frac{1}{2\pi RC} \sqrt{\frac{VVR}{R_1 + VVR}} \] Hz.

Further reading

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.e. 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


Cross references

Series 1, cards 6 & 7.

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Wireless World Circard

High-pass Sallen & Key filter

Circuit description
High-pass filter is obtained from low-pass unit by interchanging capacitors and resistors. Gain is unity at 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 resistance, and wide bandwidth e.g. op-amps connected as voltage followers or specially designed i.c. voltage followers like the LM310.

Typical performance
- $R_1 = R_2 = 100k\Omega \pm 5\%$
- $C_1 = 2C_2 = 0.022\mu F \pm 10\%$
- $f_0 = 110Hz$

$R_0$ maintained for supply voltage changes from $\pm 6$ to $\pm 18V$.

Component changes
- $R_1 = R_2$: 3.9k$\Omega$ to 1M$\Omega$
- $C_1 = 2C_2$: 0.02 to 0.2$\mu F$
- Amplifiers N5741V, LM310.

Wide-range filter: use LM310 with twin-gang potentiometer, connecting booster terminal via 1$\Omega$ resistor to supply. See graph for typical performance. Range 1: 1.25 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$\kappa F$, $C_1 = C_2 = 0.02\mu F$.

Wireless World Circard

Low, high & band-pass triple amplifier

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, band-pass at B and modified high-pass at C. Where all four amplifiers are included in a single i.e. as above, a restricted performance is offered by each amplifier individually, e.g. lower voltage gain, and the output resistance of a single 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 $R_4$ of approximately 10k$\Omega$, with $R_4 = 50R_3$ for correct bias. 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 $R_1 = 6.8k\Omega$.

$Q = R_1/R_2 = 0.68$

Drive signal 2V pk-pk.

Output approx. 50mV pk-pk.
Circuit modifications
- Single-ended supply operation possible for both op-amps and transistors, replacing shunt resistor by potential divider with \( R_1 R_3 / (R_1 + R_3) = R \) (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 \( R_3 \) and \( C_3 \) in the original have to be re-adjusted if optimum performance is to be achieved near the cut-off point.

Further reading

Cross references
Series 1, cards 6 & 7.

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Op-amp triple with phase compensation

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 slow-rate limiting changes the amplifier response leading to the possibility of sustained oscillation.

Typical data
\[ R_1 = 3.3k\Omega \pm 5\% \]
\[ C_1 = 0.1\mu F \pm 10\% \]
\[ R_3 = 330k\Omega \pm 5\% \]
\[ f_0 = \frac{1}{2\pi R_3 C_1} = 482\text{Hz} \]
\[ Q_0 = R_3/R_1 = 100 \]
\[ f_0(\text{obs}) = 470\text{Hz} \]
\[ Q(\text{obs}) = 105 \]
\[ V_{\text{out}}/V_{\text{in}} \approx Q \text{ at } f = f_0 \]

Component changes
\[ R_1 = 470\Omega \text{ to } 1\Omega \]
\[ R_3 = 470\Omega \text{ to } \infty \]
\[ C_1 = 330pF \text{ to } 10\mu F \text{ (non-polarized)} \]
Amplifiers: N5741V, SN72741P, MC1741, etc.
Supply ±5 to ±18V (±3V in some cases).
Alternative amplifiers: any op-amp compensated for integrator operation.
\[ f_0 = < 1\text{Hz to } 100\text{kHz} \]
\[ Q < 1 \text{ to } > 100 \]

Multi-feedback filter

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
\[ P = 1.5k\Omega \text{ to } 1M\Omega \]
\[ R_3 = 1k\Omega \text{ to } 100k\Omega \]
\[ C_1, C_2 = 150pF \text{ to } 10\mu F \]
\[ Q = 1 \text{ to } 50 \text{ (At high } Q \text{ values, sensitivity to component variations is excessive)} \]
\[ f_0 < 1\text{Hz to } > 100\text{kHz} \]

For \( C_1 = C_2 = C \)
\[ A_0 = R_4/R_1 \]
\[ C = 0.01\mu F \pm 10\% \]
\[ R_1 = R_3 = 1.5k\Omega \pm 5\% \]
\[ R_2 = 470k\Omega \pm 5\% \]
Amplifier: 741
\[ f_0(\text{obs}) = 845\text{Hz} \]
\[ Q(\text{obs}) = 11.3 \]

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 10k\Ω 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. $R_3 = 10k\Omega$ pot., $R_1 = 100k\Omega$.
- Summing outputs at three amplifiers with conventional virtual earth summer gives more general transfer function.
- Low-pass output available at C (normally used with $Q = 0.7$ 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 inverter—alternative method of neutralizing phase-shift.

Further reading


Cross references

Series 1, cards 1, 3, 6, 8, 11 & 12.
Wireless World Circard

Adjustable-Q twin-T notch filter

\[ f_0 = \frac{\sqrt{n}}{2\pi R_1 C_1} \]
where \( n = 2C_1/C_2 = R_1/2R_2 \) (usually)
\( f_0 = 1000\,\text{Hz} \) with
\( C_1 = 10\,\text{nF} \)
\( C_2 = 20\,\text{nF} \) (trimmed)
\( R_1 = 16.2\,\text{k}\Omega \)

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 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.

Typical performance
\( R_2 = 8.1\,\text{k}\Omega \) (trimmed)
\( f_1 = 680\,\text{Hz} \)
\( f_2 = 1650\,\text{Hz} \)
\( f_0 = 50\,\text{Hz} \) with
\( C_1 = 10\,\text{nF} \)
\( C_2 = 20\,\text{nF} \) (trimmed)
\( R_1 = 324\,\text{k}\Omega \)
\( R_2 = 162\,\text{k}\Omega \) (trimmed)
\( f_1 = 33\,\text{Hz} \)
\( f_2 = 76\,\text{Hz} \)

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 \( C_1 = 270\,\text{pF} \), \( R_1 = 600\,\text{k}\Omega \), \( n = 1 \), \( f_0 = 1\,\text{kHz} \). Very large resistors (>10M\Omega) should be avoided.

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Easily-tuned notch filter

\[ Q \text{ of resonant branch is } \frac{1}{\sqrt{\frac{C_2 R_1 R_2}{R_1 + R_2} \times \frac{1}{C_1}}} \]
\( f_0 = 1/2\pi \sqrt{C_1 C_2 R_1 R_2} \)
\( f_0 = 50\,\text{Hz} \) with
\( R_1, R_2 : 50\,\text{k}\Omega \)
\( C_1 : 4\,\text{nF} \)
\( C_2 : 1\,\mu\text{F} \)
\( R : 100\,\text{k}\Omega \)

Circuit description
The impedance between \( A \) and ground is equivalent to an inductor, \( C_2R_2 \), 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 \( C \) changes the notch frequency with no theoretical change in \( n \) as 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, \( A_2 \) 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).

Typical performance

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

Component variations
\( R : 50\,\text{k}\Omega \) to 1M\Omega
\( C_2 : 10\,\text{pF} \) to 0.1\muF
\( C_1 : 40\,\text{nF} \) to 100pF

Circuit modification
\( R_1 \) and \( R_2 \) may be kept as fixed resistors if a variable capacitor for \( C_1 \) is available.
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. 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
2. N. B. Rowe, Designing a low frequency active notch filter, Electronic Engineering, April 1972.

Cross references
Series 1, cards 3 & 10.
Wireless World Circard

Compound filters

(a) provides the bandpass characteristic shown. Low and high-pass filters used are described in cards 4 and 5 with cut-off frequencies 10kHz and 2.35kHz respectively.

\[ f_1 = 5kHz, \quad f_2 = 1.9kHz \text{ and} \quad f_3 = 13kHz. \]

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

\[ \omega_0 = (\omega_1 + \omega_2)/2, \quad f_1 = 3kHz, \quad f_2 = 0.76kHz, \quad f_3 = 12kHz, \quad f_4 = 2.1kHz \text{ and} \quad f_5 = 4.5kHz. \]

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, at 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.

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N-path filter

IC3: 74HC404 D-type flip-flop
IC2: off SN4040
connected as two-input AND gates
IC1: 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 \( f_0 = \text{clock frequency}/N \)
Bandwidth: \( 2/\pi RC \) Hz,
\( N = \text{number of low-pass filter sections} = 8. \)
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. \( R = 10kΩ \pm 10\% \), \( Q = 200. \)
IC1: 74HC74; IC3 IC501; IC5 IC7403.

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.1).
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

Further reading