PH 207
Some Examples and Active Filters

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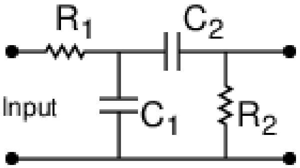
Filters

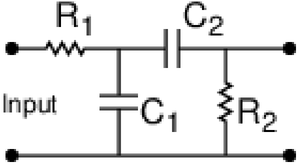
- Used for selecting some frequencies and rejecting others.
- Some common uses:
  - Reducing noise
  - Frequency division multiplexing
  - Enhancing one harmonic of a periodic signal
- Our plan:
  1. Study simple filters or building blocks
  2. Combine these building blocks to make more complex filters



Recall that
$\mathcal{A} = sR_1C_1 + 1 + \frac{R_1}{R_2} + \frac{R_1C_1}{R_2C_2} + \frac{1}{sR_2C_2}$.
So

$$T(s) = \frac{1}{\mathcal{A}} = \frac{\frac{1}{R_1C_1}s}{s^2 + \frac{1+\frac{R_1}{R_2}+\frac{R_1C_1}{R_2C_2}}{R_1C_1}s + \frac{1}{R_1R_2C_1C_2}}$$

Example 1: RCCR BPF $T(s)$ in Standard Form



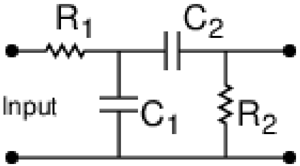
Comparing with the standard form we see that

$$\omega_0 = \frac{1}{\sqrt{R_1 R_2 C_1 C_2}}$$

$$Q = \frac{\sqrt{\frac{R_1 C_1}{R_2 C_2}}}{1 + \frac{R_1}{R_2} + \frac{R_1 C_1}{R_2 C_2}}$$

$$H = \frac{1}{1 + \frac{R_1}{R_2} + \frac{R_1 C_1}{R_2 C_2}}$$

Example 1: Special Case



If $R_1 = R_2 = R$, and $C_1 = C_2 = C$, we have

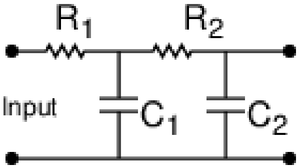
$$\omega_0 = \frac{1}{RC}$$

$$Q = \frac{1}{3}$$

$$H = \frac{1}{3}$$

Not very selective at all!

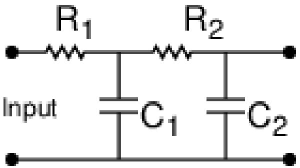
Example 2: RCRC LPF



Compute the ABCD matrix of the network to show that
$\mathcal{A} = s^2 R_1 R_2 C_1 C_2 + (R_1 C_1 + R_1 C_2 + R_2 C_2)s + 1$.
So

$$T(s) = \frac{1}{\mathcal{A}} = \frac{\frac{1}{R_1 R_2 C_1 C_2}}{s^2 + \frac{R_1 C_1 + R_1 C_2 + R_2 C_2}{R_1 R_2 C_1 C_2} s + \frac{1}{R_1 R_2 C_1 C_2}}$$

Example 2: RCRC LPF $T(s)$ in Standard Form



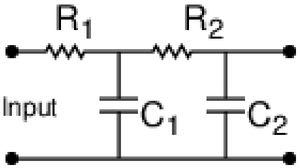
Comparing with the standard form we see that

$$\omega_0 = \frac{1}{\sqrt{R_1 R_2 C_1 C_2}}$$

$$Q = \frac{\sqrt{R_1 R_2 C_1 C_2}}{R_1 C_1 + R_1 C_2 + R_2 C_2}$$

$$H = 1$$

Example 2: Special Case



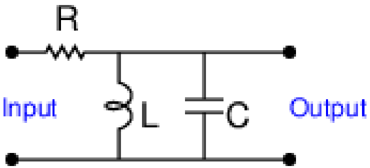
If $R_1 = R_2 = R$, and $C_1 = C_2 = C$, we have

$$\omega_0 = \frac{1}{RC}$$

$$Q = \frac{1}{3}$$

$$H = 1$$

Example 3: BPF from the Parallel RLC Network



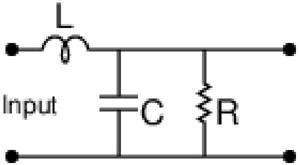
$$T(s) = \frac{\frac{s}{RC}}{s^2 + \frac{s}{RC} + \frac{1}{LC}}$$

Comparing with the standard form we see that $\omega_0 = \frac{1}{\sqrt{LC}}$, $Q = \frac{R}{\sqrt{L/C}}$, and $H = 1$.

Note that the expression for Q here differs from the expression that was derived for the BPF based on the series RLC circuit.

Here Q is proportional to R, there it was inversely proportional to R.

Example 4: LPF from the Parallel RLC Network

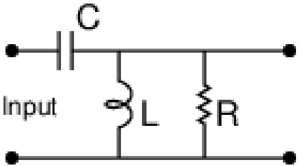


$$T(s) = \frac{\frac{1}{LC}}{s^2 + \frac{s}{RC} + \frac{1}{LC}}$$

Comparing with the standard form we see that $\omega_0 = \frac{1}{\sqrt{LC}}$, $Q = \frac{R}{\sqrt{L/C}}$, and $H = 1$.

This circuit is used for impedance matching in industrial applications.

Example 5: HPF from the Parallel RLC Network



$$T(s) = \frac{s^2}{s^2 + \frac{s}{RC} + \frac{1}{LC}}$$

Comparing with the standard form we see that $\omega_0 = \frac{1}{\sqrt{LC}}$, $Q = \frac{R}{\sqrt{L/C}}$, and $H = 1$.

This circuit is also used for impedance matching in industrial applications.

Why RC Filters?

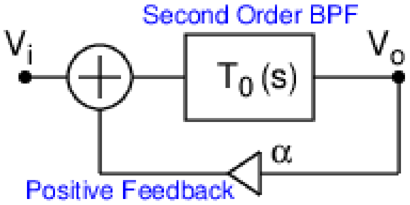
- Inductors are practical at high frequencies.
- At audio frequencies, inductors tend to be bulkier and more expensive compared to resistors and capacitors.
- So there is a desire to make audio frequency filters using resistors and capacitors only.
- But, passive RC second order networks seem to have low Q.
- How can we get more Q?
- The answer is the *active filter*.
- Active filters use amplification to compensate for the losses.

Q-Enhancement using Positive Feedback

Consider a second order BPF whose transfer function is

$$T_0(s) = \frac{H_0 \frac{\omega_{00}}{Q_0} s}{s^2 + \frac{\omega_{00}}{Q_0} s + \omega_{00}^2}$$

The extra 0s in the subscripts are there to indicate original parameters.
Now let us use positive feedback as shown.



What is the new transfer function?

New Transfer Function

$V_o = T_0(s)(V_i + \alpha V_o)$
$V_o(1 - \alpha T_0(s)) = T_0(s) V_i$

$$T(s) = \frac{V_o}{V_i} = \frac{T_0(s)}{1 - \alpha T_0(s)}$$

Substitution of the expression for $T_0(s)$ and simplification gives us

$$T(s) = \frac{H_0 \frac{\omega_{00}}{Q_0} s}{s^2 + (1 - \alpha H_0) \frac{\omega_{00}}{Q_0} s + \omega_{00}^2}$$

New Parameters

Comparing with the standard form

$$T(s) = \frac{H\frac{\omega_0}{Q}s}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2}$$

we see the following.
$\omega_0 = \omega_{00}$. Centre angular frequency does NOT change.

$$Q = \frac{Q_0}{1 - \alpha H_0}$$

$$H = \frac{H_0}{1 - \alpha H_0}$$

Both Q and H are *enhanced* by the factor $\frac{1}{1-\alpha H_0}$.
We need to be careful. If αH_0 exceeds 1, the circuit will oscillate.

Q-Enhancement: Practical Circuits

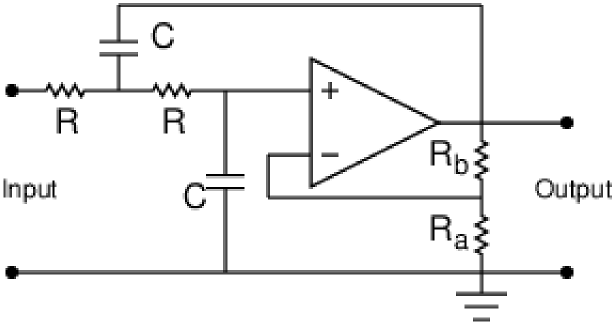
The positive feedback scheme that was described can be implemented using two operational amplifiers.

In practice, only one operational amplifier may be enough.

Not only second order BPF, even second order LPF and HPF circuits can have their Q enhanced using amplifiers.

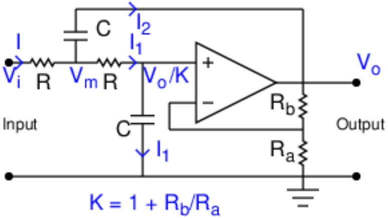
The next circuit is a practical LPF circuit.

The Sallen-Key Lowpass Filter



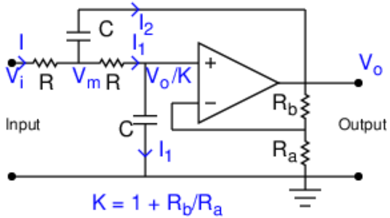
DC Gain is $K = 1 + \frac{R_b}{R_a}$.

The Sallen-Key Lowpass Filter: Analysis



We start with the output voltage V_o.
As we have a non-inverting amplifier of gain $K = 1 + R_b/R_a$, the voltage at the non-inverting input of the amplifier is V_o/K.
$I_1 = sC\frac{V_o}{K}$.
$V_m = \frac{V_o}{K} + I_1 R = (1 + sRC)\frac{V_o}{K}$.
$I_2 = sC(V_m - V_o) = sC(1 - K + sRC)\frac{V_o}{K}$.
$I = I_1 + I_2 = sC(2 - K + sRC)\frac{V_o}{K}$.
$V_i = V_m + RI = \left[1 + (3 - K)sRC + (sRC)^2\right]\frac{V_o}{K}$.

The Sallen-Key LPF: Transfer Function



So

$$T(s) = \frac{V_o}{V_i} = \frac{K}{(sRC)^2 + (3-K)sRC + 1} = \frac{K\frac{1}{(RC)^2}}{s^2 + \frac{3-K}{RC}s + \frac{1}{(RC)^2}}$$

Comparing with the standard forms we see that we have a second order LPF with

$\omega_0 = \frac{1}{RC}$,

$Q = \frac{1}{3-K}$,

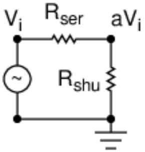
and

$H = K = 1 + R_b/R_a$.

The Sallen-Key LPF: Gain Reduction

The circuit just described has a DC gain of $K = 1 + R_b/R_a$ which exceeds 1.
In many applications, we want a DC gain H_{desired}, which is 1 or less.
This could be achieved by using a voltage divider of division ratio $a = H_{\mathrm{desired}}/K$, followed by a unity gain buffer.
But this can also be achieved by splitting the input resistor into two parts.

Split Input Resistor



We require that the voltage division ratio should be

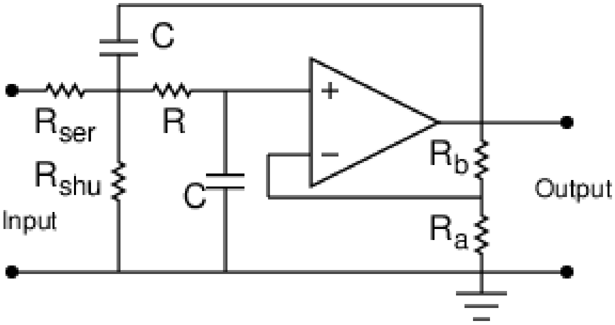
$$\frac{R_{\text{shu}}}{R_{\text{ser}} + R_{\text{shu}}} = a,$$

and the parallel combination of R_{shu} and R_{ser} should be R. So

$$\frac{R_{\text{ser}} R_{\text{shu}}}{R_{\text{ser}} + R_{\text{shu}}} = R.$$

Dividing the second equation by the first, we get
$R_{\text{ser}} = R/a$.
Then solving for R_{shu} we get $R_{\text{shu}} = R/(1-a)$.

The Sallen-Key LPF: Practical Circuit



The Sallen-Key LPF: Design

Specifications: f_0, Q, H_{desired}, and C

Compute $R = \frac{1}{2\pi f_0 C}$

Compute $K = 3 - \frac{1}{Q}$

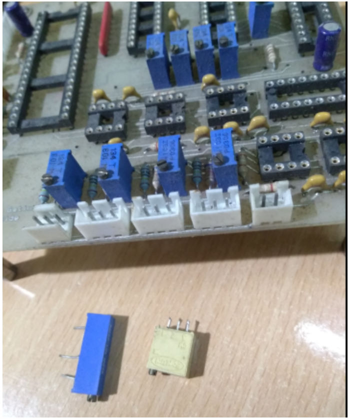
$R_b/R_a = K - 1$. Usually one sets $R_a = 10\,\text{k}\Omega$, and then computes $R_b = (K-1)R_a$.

Compute $a = H_{\text{desired}}/K$.

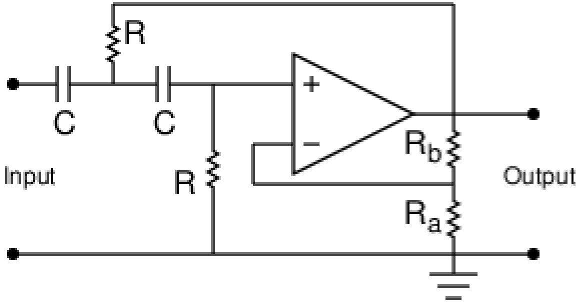
Then compute $R_{\text{ser}} = R/a$ and $R_{\text{shu}} = R/(1-a)$.

Trimmer Potentiometers

Filter design requires nonstandard resistance values. They are usually implemented using *trimmer* potentiometers. The picture shows trimmers in use.



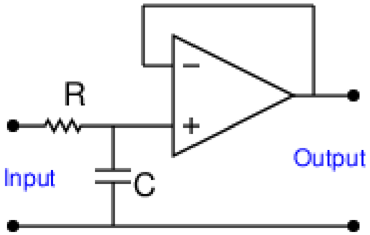
The Sallen-Key Highpass Filter



$$T(s) = \frac{Ks^2}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2}$$

$K = 1 + \frac{R_b}{R_a}$, $\omega_0 = \frac{1}{RC}$, $Q = \frac{1}{3-K}$.
Gain reduction requires a separate voltage divider followed by unity gain buffer.
Splitting of the input capacitor is not practical.

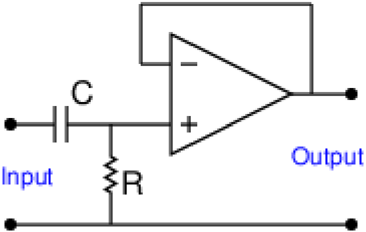
The First Order RC Lowpass Filter



$$T(s) = \frac{\omega_0}{s + \omega_0}$$

$$\omega_0 = \frac{1}{RC}$$

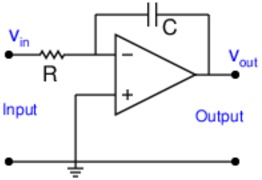
The First Order CR Highpass Filter



$$T(s) = \frac{s}{s + \omega_0}$$

$$\omega_0 = \frac{1}{RC}$$

The Inverting Integrator

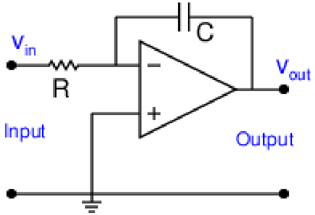


Because of virtual of phantom ground principle, $v_- = 0$, and there is no current flowing into the inverting input terminal of the operational amplifier.
The current from the input is $v_{\rm in}/R$. The current from the output is $C\frac{\mathrm{d}v_{\rm out}}{\mathrm{d}t}$. The sum of these currents must be zero. Or, $v_{\rm in}/R + C\frac{\mathrm{d}v_{\rm out}}{\mathrm{d}t} = 0$
Or,

$$\frac{\mathrm{d}v_{\rm out}}{\mathrm{d}t} = -\frac{v_{\rm in}}{RC}.$$

The derivative of the output is proportional to the input, or, the output is an integral of the input. Hence the name. The initial charge on the capacitor determines the constant of integration.

The Inverting Integrator: Transfer Function



$$T(s) = -\frac{1/(sC)}{R} = -\frac{1}{sRC} = -\frac{s_0}{s},$$

where,

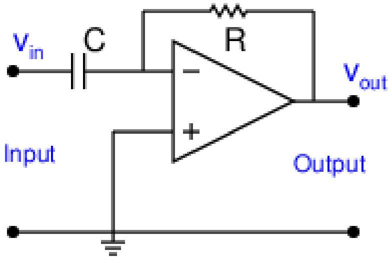
$$s_0 = \frac{1}{RC}.$$

Remember to put a high resistance in parallel with the capacitor if the integrator is used without any other negative feedback.
Due to the negative sign in the transfer function, this circuit is called the *inverting* integrator. The adjective 'inverting' is usually omitted.

The Integrator: Uses

- Frequently used as a building block in filter design
- Integration: Induced EMF to flux
- Waveform generation: Square wave to triangular wave

The (Inverting) Differentiator

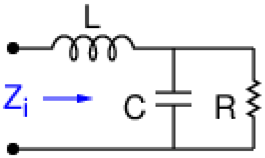


Interchanging the R and C in an integrator results in a *differentiator*.

$$v_{\mathrm{out}} = -RC\frac{\mathrm{d}v_{\mathrm{in}}}{\mathrm{d}t}.$$

This circuit is not much used. This is because differentiation enhances high frequencies.

Example of Impedance Matching

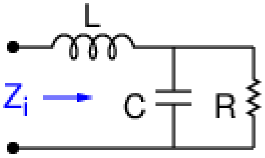


At what frequency is Z_i real?
What is its value at that frequency?

$$Z_i(s) = sL + \frac{R}{1 + sRC}$$

$$Z_i(j\omega) = j\omega L + \frac{R}{1 + j\omega RC} = j\omega L + \frac{R(1 - j\omega RC)}{1 + (\omega RC)^2}$$

Impedance Matching Analysis

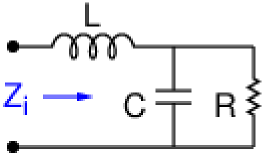


For $Z_i(j\omega)$ to be real, we require

$$\omega L = \frac{\omega R^2 C}{1 + (\omega RC)^2}$$

$$(\omega RC)^2 + 1 = \frac{R^2 C}{L}$$

Impedance Matching Results



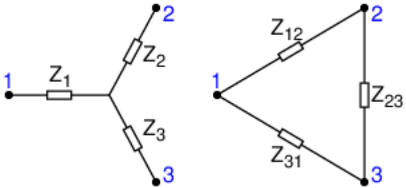
$$\omega = \frac{1}{\sqrt{LC}}\sqrt{1 - \frac{L/C}{R^2}}$$

At this ω, Z_i is

$$\frac{L/C}{R}$$

This value may be small enough for the generator to deliver more power to the load.

Star to Delta Transformation



Requirements for the two networks to be equivalent:

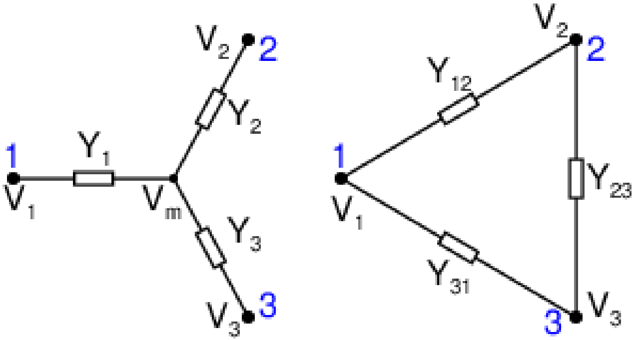
$$Z_{23} = Z_2 + Z_3 + \frac{Z_2 Z_3}{Z_1},$$

$$Z_{31} = Z_3 + Z_1 + \frac{Z_3 Z_1}{Z_2},$$

$$Z_{12} = Z_1 + Z_2 + \frac{Z_1 Z_2}{Z_3}.$$

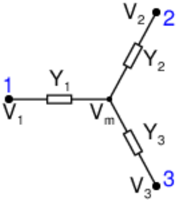
How do we show this?

Star to Delta Transformation



To derive the conditions, we first label the star and delta networks in terms of admittances. Here $Y_1 = 1/Z_1$, $Y_2 = 1/Z_2$, $Y_3 = 1/Z_3$, $Y_{23} = 1/Z_{23}$, $Y_{31} = 1/Z_{31}$, and $Y_{12} = 1/Z_{12}$. We have also indicated three arbitrary voltages V_1, V_2, and V_3, at the three terminals. The voltage where the three star elements meet is indicated as V_m. V_m will be first be determined in terms of V_1, V_2, and V_3.

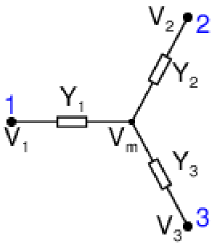
Star to Delta Transformation



In the star network, the current entering Terminal 1 is $I_1 = Y_1(V_1 - V_m)$. Likewise, the current entering Terminal 2 is $I_2 = Y_2(V_2 - V_m)$. and the current entering Terminal 3 is $I_3 = Y_3(V_3 - V_m)$. The sum of these currents $I_1 + I_2 + I_3 = 0$ by KCL. Or, $Y_1(V_1 - V_m) + Y_2(V_2 - V_m) + Y_3(V_3 - V_m) = 0$.
So

$$V_m = \frac{Y_1 V_1 + Y_2 V_2 + Y_3 V_3}{Y_1 + Y_2 + Y_3}.$$

Star to Delta Transformation



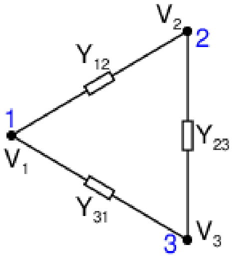
Since V_m is now known in terms of V_1, V_2, and V_3, it is possible to compute

$$I_1 = Y_1(V_1 - V_m) = Y_1\left(V_1 - \frac{Y_1V_1 + Y_2V_2 + Y_3V_3}{Y_1 + Y_2 + Y_3}\right)$$

$$= (V_1 - V_2)\frac{Y_1Y_2}{Y_1 + Y_2 + Y_3} + (V_1 - V_3)\frac{Y_1Y_3}{Y_1 + Y_2 + Y_3}.$$

Similarly, expressions for I_2, and I_3 can be obtained.

Star to Delta Transformation



Here the current entering Terminal 1 is seen to be

$$I_1 = (V_1 - V_2)Y_{12} + (V_1 - V_3)Y_{31}.$$

If the delta network is to be equivalent to the star network, then this expression should agree with the expression for I_1 on the previous slide.

Star to Delta Transformation

Then it is required that

$$Y_{12} = \frac{Y_1 Y_2}{Y_1 + Y_2 + Y_3}.$$

Similar considerations for the equalities of I_2, and I_3 show that

$$Y_{31} = \frac{Y_3 Y_1}{Y_1 + Y_2 + Y_3},$$

and

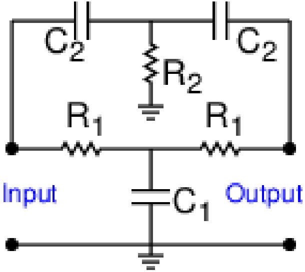
$$Y_{23} = \frac{Y_2 Y_3}{Y_1 + Y_2 + Y_3}.$$

Expressions in the impedance form:

$$Z_{12} = \frac{1}{Y_{12}} = \frac{Y_1 + Y_2 + Y_3}{Y_1 Y_2} = \frac{1}{Y_2} + \frac{1}{Y_1} + \frac{Y_3}{Y_1 Y_2} = Z_2 + Z_1 + \frac{Z_1 Z_2}{Z_3} = Z_1 + Z_2 + \frac{Z_1 Z_2}{Z_3},$$

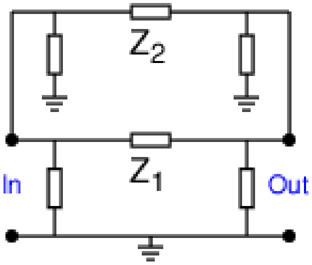
etc. This completes the derivation of the requirements stated.

The Twin-T Notch Filter



At what frequency is the output zero?

Twin-T Analysis

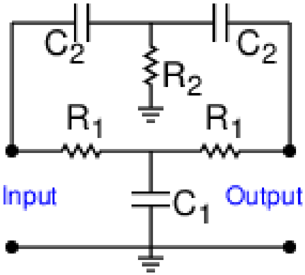


$$Z_1 = 2R_1 + sR_1^2 C_1$$

$$Z_2 = \frac{2}{sC_2} + \frac{1}{s^2 R_2 C_2^2}$$

For no transmission, we need $Z_1 = -Z_2$.

Twin-T Analysis



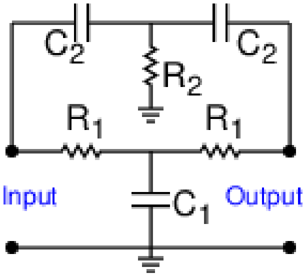
This requires

$$\omega_0^2 = \frac{2}{R_1^2 C_1 C_2},$$

and

$$\omega_0^2 = \frac{1}{2 R_1 R_2 C_2^2}.$$

Twin-T Analysis



Or,

$$R_1 C_1 = 4 R_2 C_2.$$

One way of achieving this is to set $R_1 = R$, $R_2 = R/2$, $C_1 = 2C$, and $C_2 = C$, so that

$$\omega_0 = \frac{1}{RC}.$$

Twin-T Filter: Uses

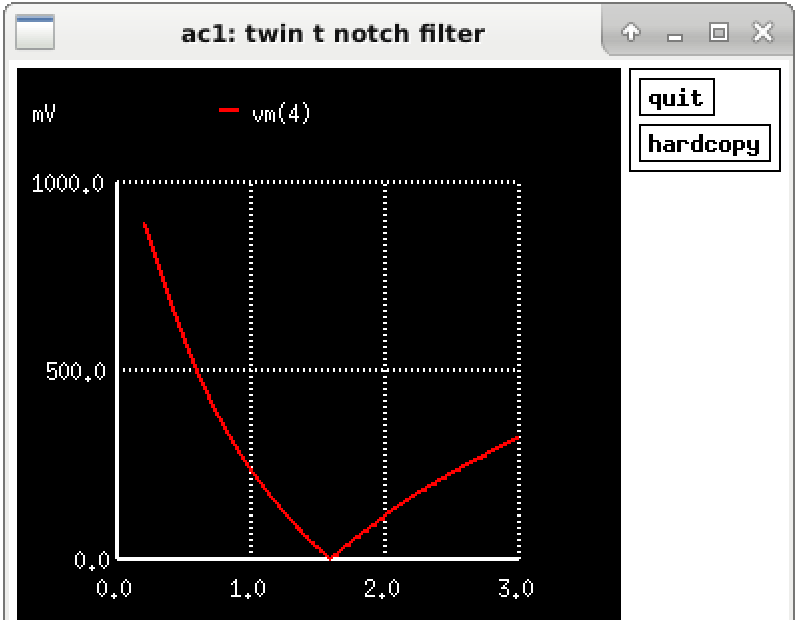
- At frequency $f_0 = 1/(2\pi RC)$, there is no transmission.
- Can be used as a notch filter.
- If used in the negative feedback path, can be part of a narrow band filter.
- This is a third order filter, not part of the mainstream.
- Much used in various forms.

Twin-T Filter: SPICE Code

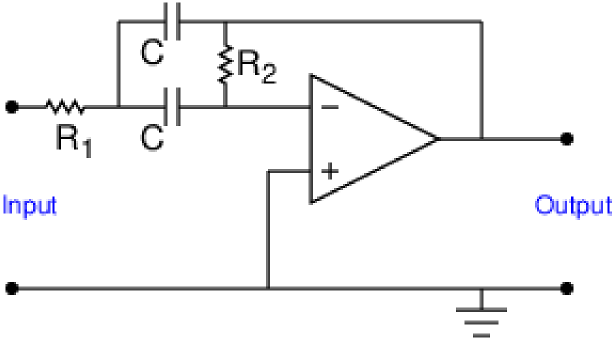
```
Twin T Notch filter
*****************************
VIN 1 0 AC 1
R1A 1 2 10k
R1B 2 4 10k
R2 3 0 5k
C1 2 0 20n
C2A 1 3 10n
C2B 3 4 10n
.AC LIN 1000 0.2k 3.0k

.CONTROL
run
plot vp(4)
plot vm(4)
.ENDCONTROL
```

Twin-T Filter: Magnitude Plot

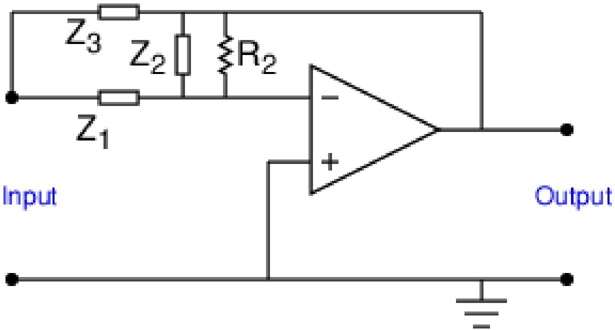


The Single Amplifier Biquad (SAB)



- This circuit acts as a bandpass filter.
- Developed by Delyiannis and Friend.

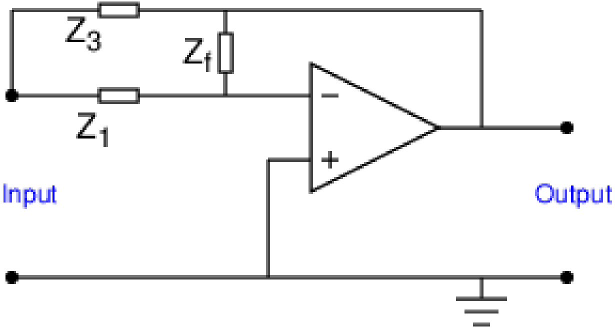
SAB Analysis



$$Z_1 = 2R_1 + \frac{1}{sC} = \frac{1 + 2sR_1C}{sC}$$

$$Z_2 = \frac{2}{sC} + \frac{1}{s^2C^2R_1} = \frac{1 + 2sR_1C}{s^2R_1C^2}$$

SAB Transfer Function



$$Z_f = Z_2 || R_2 = \frac{(1 + 2sR_1C)R_2}{1 + 2sR_1C + s^2R_1R_2C^2}$$

$$T(s) = -\frac{Z_f}{Z_1} = \frac{-sR_2C}{1 + 2sR_1C + s^2R_1R_2C^2} = \frac{-\frac{1}{R_1C}s}{s^2 + \frac{2}{R_2C}s + \frac{1}{R_1R_2C^2}}$$

SAB Parameters

We see that the SAB is a BPF.
Comparing with the standard form we get

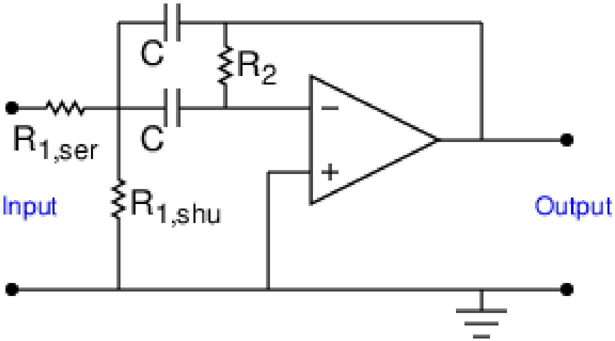
$$\omega_0 = \frac{1}{\sqrt{R_1 R_2} C}$$

$$Q = \frac{1}{2}\sqrt{\frac{R_2}{R_1}}$$

$$H = -\frac{R_2}{2R_1} = -2Q^2$$

Note that $R_2/R_1 = 4Q^2$.

SAB With Gain Reduction



SAB Design

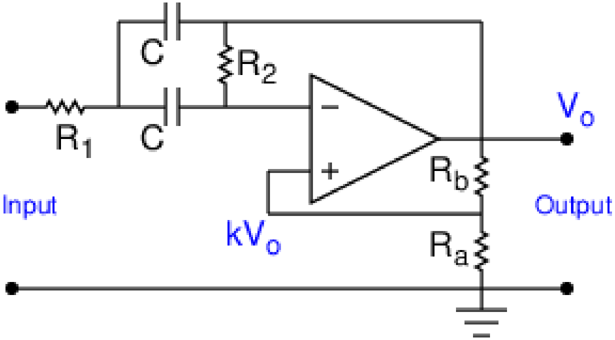
Specification: f_0, Q, H_{desired}, and C are specified.
H_{desired} should be negative.
Steps:

1. $R_2 = Q/(\pi f_0 C)$.
2. $R_1 = R_2/(4Q^2)$.
3. $a = H_{\mathrm{desired}}/(-2Q^2)$.
4. $R_{1,\mathrm{ser}} = R_1/a$.
5. $R_{1,\mathrm{shu}} = R_1/(1-a)$.

SAB Disadvantage

- $R_2/R_1 = 4Q^2$ can be quite large.
- Hard to get such high ratio inside integrated circuits.
- The remedy is to first design a low Q SAB, and then enhance its Q using positive feedback.

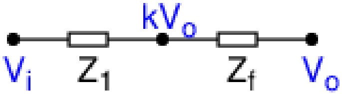
Q-Enhanced SAB



Positive feedback is used.

$$k = \frac{R_a}{R_a + R_b}$$

Q-Enhanced SAB Analysis



$$\frac{V_i - kV_o}{Z_1} = \frac{kV_o - V_o}{Z_f}$$

$$V_i - kV_o = \frac{kV_o - V_o}{Z_f/Z_1} = \frac{kV_o - V_o}{-T_0(s)} = \frac{(1-k)V_o}{T_0(s)}$$

where, $T_0(s) = -Z_f/Z_1$ is the transfer function of the original SAB.

$$V_i = kV_o + \frac{(1-k)V_o}{T_0(s)} = V_o \frac{1 - k + kT_0(s)}{T_0(s)}$$

Q-Enhanced SAB Transfer Function

So the transfer function of the Q-Enhanced SAB is

$$T(s) = \frac{V_o}{V_i} = \frac{T_0(s)}{1 - k + kT_0(s)} = \frac{\frac{1}{1-k}T_0(s)}{1 + \frac{k}{1-k}T_0(s)}$$

Let

$$\alpha = \frac{k}{1 - k}$$

so that

$$k = \frac{\alpha}{1 + \alpha}$$

and

$$\frac{1}{1 - k} = 1 + \alpha$$

Q-Enhanced SAB Transfer Function

So we have

$$T(s) = \frac{(1+\alpha)T_0(s)}{1+\alpha T_0(s)}$$

Let us rewrite the transfer function of the original SAB, $T_0(s)$ as

$$T_0(s) = \frac{-2Q_0^2 \frac{\omega_0}{Q_0} s}{s^2 + \frac{\omega_0}{Q_0} s + \omega_0^2}$$

the original Q being rewritten as Q_0 and H being written as $-2Q_0^2$. Substitution and simplification results in

$$T(s) = \frac{-2(1+\alpha)Q_0^2 \frac{\omega_0}{Q_0} s}{s^2 + (1 - 2\alpha Q_0^2)\frac{\omega_0}{Q_0} s + \omega_0^2}$$

Q-Enhanced SAB Parameters

We see that for the Q-Enhanced SAB, ω_0 is unchanged,

$$Q = \frac{Q_0}{1 - 2\alpha Q_0^2}$$

and

$$H = \frac{-2Q_0^2(1 + \alpha)}{1 - 2\alpha Q_0^2}$$

Selecting Q_0 and Computing α

It is recommended to use

$$Q_0 \approx 1.5$$

Compute $R_2 = Q_0/(\pi f_0 C)$, and $R_1 = R_2/(4Q_0^2)$.
Next, compute α using

$$\alpha = \frac{1 - Q_0/Q}{2Q_0^2}$$

Then compute k using

$$k = \frac{\alpha}{1 + \alpha}$$

We have $R_b/R_a = 1/k - 1$.

Gain Reduction

Compute $H = \dfrac{-2Q_0^2(1+\alpha)}{1-2\alpha Q_0^2}$.

Then compute

$$a = H_{\text{desired}}/H$$

Then compute

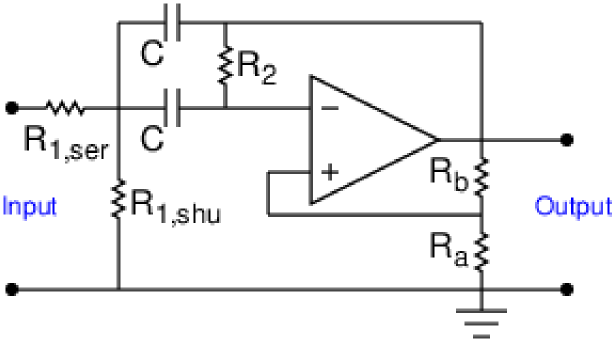
$$R_{1,\text{ser}} = R_1/a$$

and

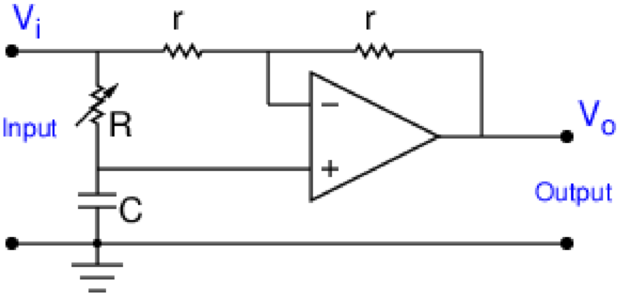
$$R_{1,\text{shu}} = R_1/(1-a)$$

The final circuit is on the next slide.

Q-Enhanced SAB with Gain Reduction

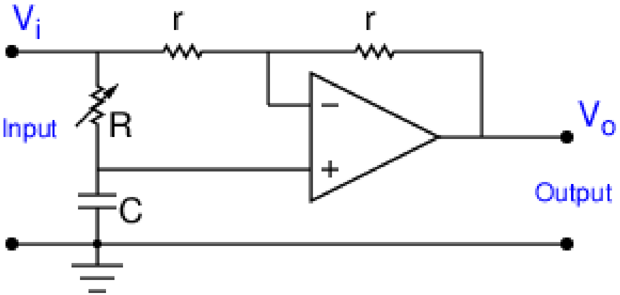


An All-Pass Filter



Used as a phase shifter.
One could make this from the difference of the transfer functions of the RC LPF and the CR HPF after suitable buffering and subtraction, but this circuit uses only one operational amplifier.

All-Pass Filter Analysis



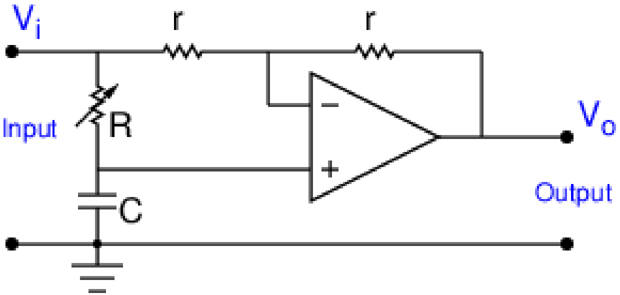
The voltage at the input pins is $(V_i + V_o)/2$.
So

$$\frac{V_i - \frac{V_i+V_o}{2}}{R} = sC\frac{V_i + V_o}{2}$$

Or,

$$\frac{V_i - V_o}{2} = sRC\frac{V_i + V_o}{2}$$

All-Pass Filter Transfer Function



Or,

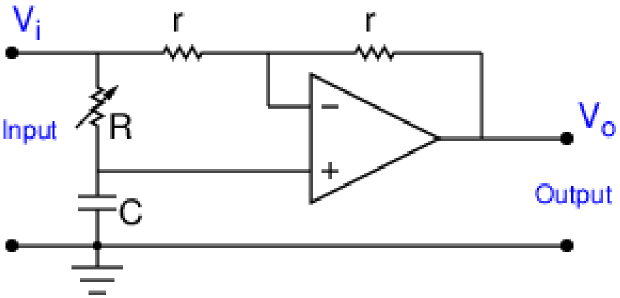
$$V_i(1 - sRC) = V_o(1 + sRC)$$

So,

$$T(s) = \frac{V_o}{V_i} = \frac{1 - sRC}{1 + sRC} = \frac{\omega_0 - s}{\omega_0 + s}$$

where $\omega_0 = \frac{1}{RC}$.

All-Pass Filter as a Phase Shifter



We have

$$|T(j\omega)| = 1$$

and

$$\angle T(j\omega) = -2\arctan(\omega/\omega_0)$$

As R changes from 0 to ∞, the phase lag changes from 0 to π.