Bandpass filter features adjustable Q and constant maximum gain

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Applications such as audio equalizers require bandpass filters with a constant maximum gain that's independent of the filter's quality factor, Q. However, all of the well-known filter architectures—Sallen-Key, multiple-feedback, state-variable, and Tow-Thomas—suffer from altered maximum gain when Q varies. Equation 1 expresses the second-order bandpass transfer function of a bandpass filter:

\[ H_{BP}(s) = \frac{K \left( \frac{s}{\omega_0} \right)^2 + 1}{\omega_0^2 + \frac{1}{Q} \left( \frac{s}{\omega_0} \right) + 1} \],

where K represents the filter’s gain constant. When the input frequency equals \( \omega_0 \), the filter’s gain, \( A_{MAX} \), is proportional to the product, KQ. Thus, modifying the quality factor alters the gain and vice versa.

This Design Idea describes a filter structure in which \( K \) is inversely proportional to Q. Altering Q also modifies K, producing a magnitude-plot set in which the curves maintain the same maximum gain at the central frequency \( \omega_0 \)—that is, KQ remains constant. Figure 1 shows the filter, which comprises a twin T cell with an adjustable quality factor and a differential stage. The differential stage comprises op amp IC\(_3\) and resistors R\(_{5A}\) through R\(_{5D}\). This stage outputs the difference between the filter’s input signal and the twin-T network’s output. Capacitors C\(_1\) and C\(_2\) are of equal value, C\(_1\) = C\(_2\), capacitor C\(_3\) equals 2C\(_1\), resistors R\(_1\) and R\(_2\) are also equal and of value R = R\(_1\) = R\(_2\), and R\(_3\) equals R/2. Equation 2 describes the twin-T cell’s transfer-function response as a notch filter producing output \( V_{BR}(t) \):

\[ H_{BR}(s) = \frac{V_{BR}(s)}{V_{IN}(s)} = \frac{(RC_3)^2 + 1}{(RC_3)^2 + 4RC(1-m)s + 1} \]  

where \( m \) represents the twin-T cell’s feedback factor. If you designate R\(_{XY}\) as the resistance potentiometer R\(_X\)’s upper terminal, Point X; the rotor as Point Y; and R\(_{YZ}\) as the resistance between the rotor and the bottom terminal, Point Z, you can express \( m \) as the quotient of Equation 4:

\[ m = \frac{R_{YZ}}{R_{XY} + R_{YZ}} = \frac{R_{YZ}}{R_4} \]  

Comparing Equation 3 with the respective normalized transfer functions of a bandpass filter, Equation 1, Equation 5 expresses the central frequency of the filter, \( \omega_0 \), coincident with the transmission zero of the twin-T network:

\[ \omega_0 = \frac{1}{RC} \]  

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Equations 6 and 7, respectively, give quality factor Q and gain constant K:

\[
Q = \frac{1}{4(1-m)}; \quad (6)
\]

\[
K = \frac{1}{Q} = 4(1-m). \quad (7)
\]

The maximum gain, \( A_{\text{MAX}} \) at \( \omega = \omega_0 \), always remains constant and equal to 1 (0 dB) and is independent of Q. The minimum quality factor is 1/4 for \( m = 0 \), which corresponds to the potentiometer’s rotor connected to ground. The maximum gain is theoretically infinite, but, in practice, it’s difficult to achieve a quality factor beyond 50. In most applications, Q ranges from 1 to 10.

Figure 2 shows the filter’s magnitude and phase Bode plots for the frequency-notch output \( V_{\text{BR}}(t) \) (available at IC 1’s output) for values of m from 0.1 to 0.9. Figure 3 shows Bode plots for the filter’s bandpass output, \( V_{\text{OUT}}(t) \), for the same values of m. In both graphs, frequency \( f_0 \) equals 1061 Hz. To minimize frequency-response variations and improve response accuracy, you can build the filter with precision metal-film resistors of 1% or better tolerance. Likewise, use close-tolerance mica, polycarbonate, polyester, polystyrene, polypropylene, or Teflon capacitors. For best performance, avoid carbon resistors and electrolytic, tantalum, or ceramic capacitors.

**Figure 2** Magnitude and phase Bode plots at the frequency-notch output, \( V_{\text{BR}}(t) \), show effects of varying twin-T-cell feedback factor, m, from 0.1 to 0.9.

**Figure 3** Magnitude and phase Bode plots at the bandpass output, \( V_{\text{OUT}}(t) \), show effects of varying twin-T-cell feedback factor, m, from 0.1 to 0.9.

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Moving-coil meter measures low-level currents

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Although an analog moving-coil meter may lack the resolution and accuracy that a digital readout provides, a meter remains the display of choice for certain applications. A digital readout simply cannot provide information about a measurement’s rate of change, and tracking a reading’s trend is easier on an analog meter.

Large moving-coil meters may require significant amounts of current for full-scale deflection, and using a shunt resistor may prove impractical when the meter current is larger than the current you are measuring. You can solve the problem by driving the meter from a separate power supply (Figure 1). In this example, an 8-in. moving-coil meter that requires 15 mA for full-scale deflection displays a current range of 0 to 1A dc. This technique can also simplify specifying or fabricating shunt resistors for custom current ranges. Unlike other current-sense amplifiers that derive operating power from the current you are measuring, IC₁ provides a separate supply-voltage terminal for its internal circuitry. In operation, IC₁’s output current, \( I_{\text{OUT}} \), equals \( \frac{V_{\text{SENSE}}}{100} \), where \( V_{\text{SENSE}} \) is the voltage across \( R_{\text{SENSE}} \).

This Design Idea uses IC₁ rather than the many current-sense amplifiers available because it provides a separate supply-voltage terminal for the internal circuitry, whereas other devices take power from the current you are measuring. In this application, a full-scale current of 1A develops 1V across \( R_{\text{SENSE}} \), which IC₁ converts to a maximum output current.
of 10 mA that produces a maximum voltage of 1V across R1. Operational amplifer IC2 and transistor Q1 form a voltage-controlled current sink that draws current through meter M1. A full-scale reading of 15 mA develops 1V across 66Ω resistor RSENSE2. You can adjust the resistor’s value to calibrate the meter or to alter the full-scale current range.

This circuit also allows separation of the measurement point and meter location. Moving-coil meters are not intended for applications that require precision measurement, and you can use relaxed-accuracy passive components. Bypass the instrument-supply voltage with decoupling capacitors that the electrical-noise environment requires.

**Figure 1**

This circuit displays current on a moving-coil meter that consumes a substantial fraction of the current you are measuring.

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**MOSFET enhances voltage regulator’s overcurrent protection**

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The classic LM317 adjustable-output linear voltage regulator offers a relatively high, if package-dependent, current-handling capability. In addition, the LM317 features current limiting and thermal-overload protection. With the addition of a few components, you can enhance an LM317-based voltage regulator by adding a high-speed short-circuit current limiter (*Figure 1*). Under normal operation, resistors R2 and R3 apply V_{GS} bias to power MOSFET Q1, an IRF4905S, which fully conducts and presents an on-resistance of a few milliohms. The voltage drop across current-sampling resistor R1 is proportional to IC,‘s input current and provides base drive for bipolar transistor Q2.

As load current increases, the voltage across R1 increases, biasing Q2 into conduction and decreasing Q1’s gate bias. As Q1’s gate bias decreases, its on-resistance increases, limiting the current into IC1, according to \( I_{MAX} = \frac{V_{BE}Q2}{R1} \), or approximately 0.6V/1Ω.

Diodes D1 and D2, respectively, protect against capacitive-load discharge and polarity reversal. Depending on the circuit’s requirements, IC1 and Q1 may require heat dissipators.

**Figure 1**

A few added components extend this linear regulator’s overcurrent protection.
Digitally programmable resistor serves as test load
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FIGURE 1 ILLUSTRATES a digitally programmable precision resistance that can serve as a microprocessor-driven power-supply load in custom-designed ATE (automatic-test equipment). An 8-bit current-output DAC, IC1, a DAC08, drives current-to-voltage converter IC2A, which in turn drives the gate of power MOSFET Q1. The device under test connects to J1 and J2. In operation, current from the device under test develops a voltage across sampling resistors R8A and R8B. Amplifier IC2B drives IC1's reference input and closes the feedback path. Transistor Q2 provides overcurrent protection by diverting gate drive from Q1 when the voltage drop across R8A and R8B reaches Q2's VBE(ON). V0 and I0 represent the output voltage and current, respectively; N represents the decimal equivalent of the binary input applied to IC1; and A represents the gain of the amplifier stage IC2B. R1 comprises the parallel combination of R1A and R1B. Equation 1 describes the circuit’s load current:

\[
\frac{V_0}{I_0} = \frac{A \times R_e}{R_6} \times \frac{N}{256}.
\]

Solving Equation 2 yields the circuit’s output resistance:

\[
V_0 = \frac{A \times R_e \times N}{256}.
\]

Using the component values shown, the circuit’s equivalent resistance ranges from approximately 5.5Ω for N=0, an all-zero binary input, to 255Ω for N=255, an all-one binary input.

You can modify circuit values to cover other resistance ranges. Replacing the 8-bit DAC08 with a 10-bit D/A converter increases resistance resolution. To increase the circuit’s power-handling capability, replace Q1 with a higher power MOSFET and an appropriately sized heat dissipator. Capacitors C3 and C4 control the circuit’s bandwidth.