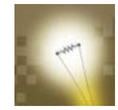

Edited by Brad Thompson

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:

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

where K represents the filter's gain constant. When the input frequency equals ω_O, 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 ω_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, $C=C_1=C_2$, capacitor C_3 equals 2C, 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 circuit's transfer-function response as a notch filter producing output $V_{BR}(t)$:

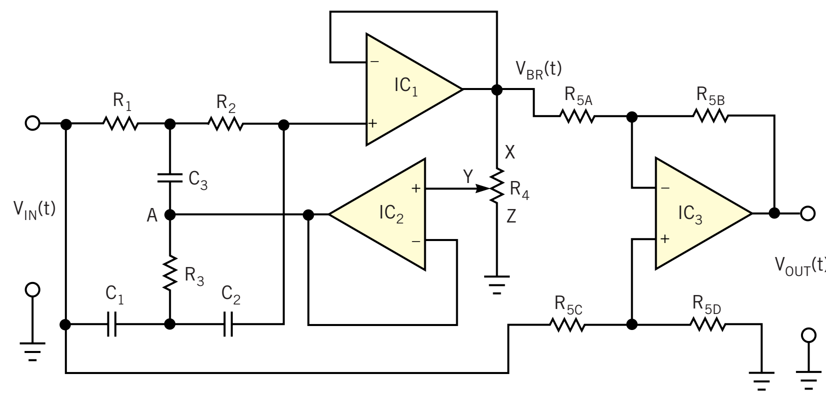

Figure 1 This bandpass active filter features adjustable Q and maximum gain in the passband and consists of a twin-T cell with Q adjustment and a differential output stage. You can also extract a frequency-notch output from the voltage-follower stage.

$$H_{BR}(s)=\frac{V_{BR}(s)}{V_{IN}(s)}=\frac{(RCs)^2+1}{(RCs)^2+4RC(1-m)s+1}. \quad (2)$$

Equation 3 describes the compete circuit's transfer function, a bandpass-filter response with output $V_{OUT}(t)$:

$$H_{BP}(s)=\frac{V_{OUT}(s)}{V_{IN}(s)}=\frac{4RC(1-m)s}{(RCs)^2+4RC(1-m)s+1}, \quad (3)$$

where m represents the twin-T cell's feedback factor. If you designate R_{XY} as the resistance potentiometer R_4'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**:

$$m=\frac{R_{YZ}}{R_{XY}+R_{YZ}}=\frac{R_{YZ}}{R_4}. \quad (4)$$

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

$$\omega_0=\frac{1}{RC}. \quad (5)$$

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

$$Q = \frac{1}{4(1-m)}; \qquad (6)$$

$$K = \frac{1}{Q} = 4(1-m). \qquad (7)$$

The maximum gain, A_{MAX}, at $\omega=\omega_O$, 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 $V_{BR}(t)$ (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, $V_{OUT}(t)$, for the same values of m. In both graphs, frequency f_O 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.□

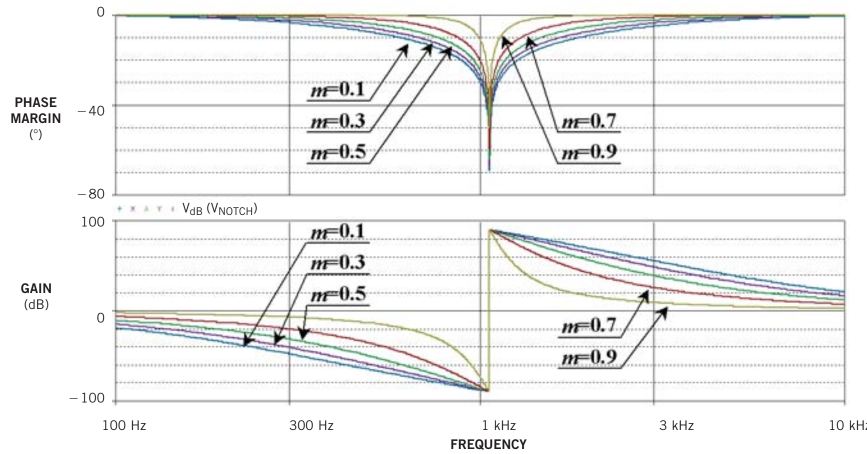


Figure 2 Magnitude and phase Bode plots at the frequency-notch output, $V_{BR}(t)$, show effects of varying twin-T-cell feedback factor, m, from 0.1 to 0.9.

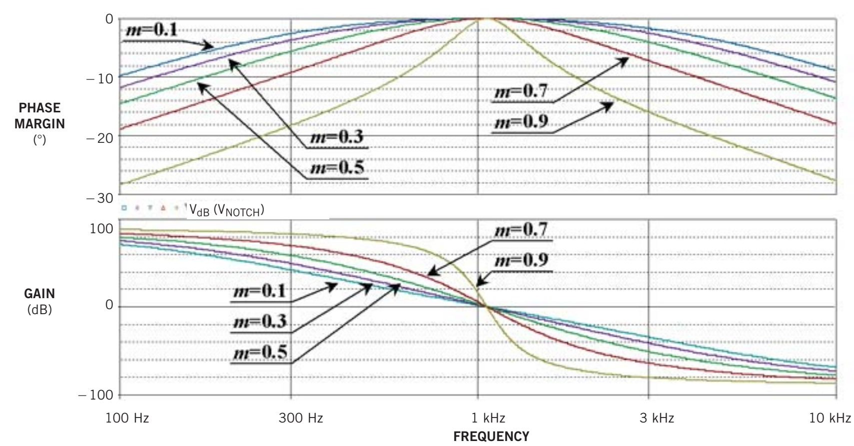


Figure 3 Magnitude and phase Bode plots at the bandpass output, $V_{OUT}(t)$, show effects of varying twin-T-cell feedback factor, m, from 0.1 to 0.9.

Moving-coil meter measures low-level currents

Kevin Bilke, Maxim Integrated Products Inc, Hook Hants, UK

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, IC_1 provides a separate supply-voltage terminal for its internal circuitry. In operation, IC_1's output current, I_{OUT}, equals V_{SENSE}/100V, where V_{SENSE} is the voltage across R_{SENSE1}.

This Design Idea uses IC_1 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_{SENSE1}, which IC_1 converts to a maximum output current of 10 mA that produces a maximum voltage of 1V across R_1. Operational amplifier IC_2 and transistor Q_1 form a voltage-controlled current sink that draws current through meter M_1. A full-scale reading of 15 mA develops 1V across 66Ω resistor R_{SENSE2}. 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.□

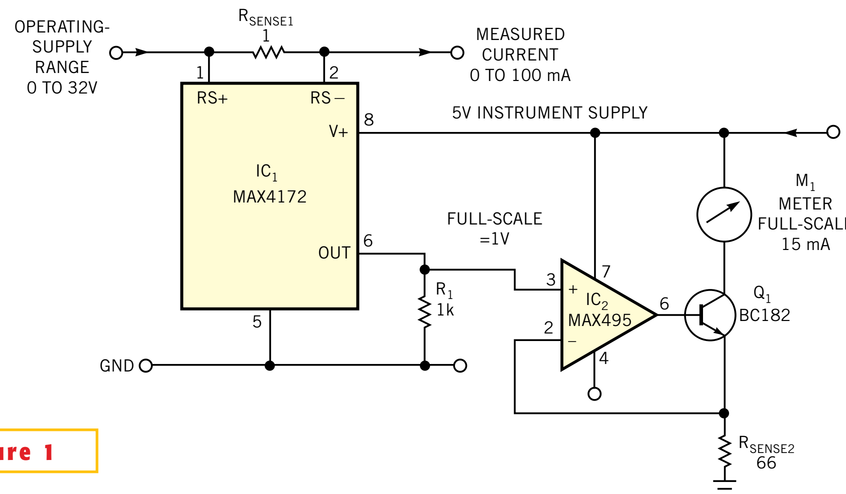


Figure 1

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

MOSFET enhances voltage regulator's overcurrent protection

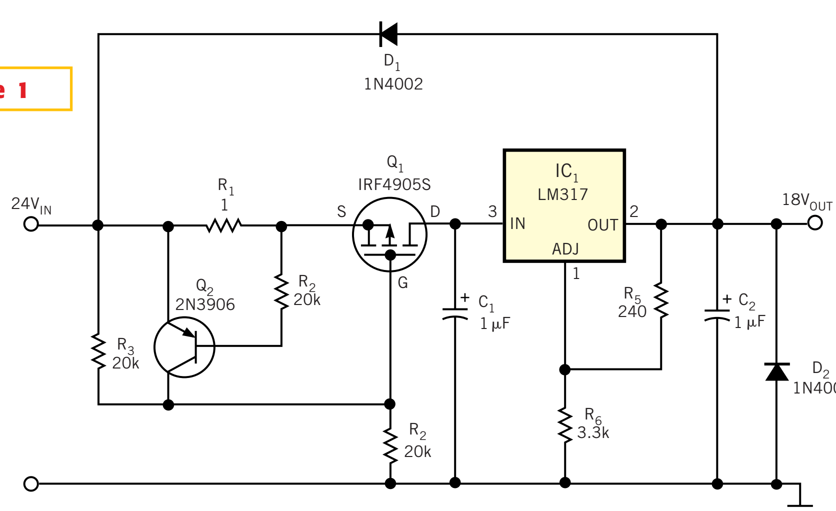
Herb Hulseman, Eagle Test Systems, Mundelein, IL

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 R_2 and R_3 apply V_{GS} bias to power MOSFET Q_1, an IRF4905S, which fully conducts and presents an on-resistance of a few milliohms. The voltage drop across current-sampling resistor R_1 is proportional to IC_1's input current and provides base drive for bipolar transistor Q_2.

As load current increases, the voltage across R_1 increases, biasing Q_2 into conduction and decreasing Q_1's gate bias. As Q_1's gate bias decreases, its on-resistance increases, limiting the current into IC_1, according to $I_{MAX}=V_{BE}Q_2/R_1$, or approximately 0.6V/1Ω.

Resistors R_5 and R_6 set IC_1's output voltage, as the LM317's application notes describe. By varying the value of R_1, you can adjust the circuit's limiting current from milliamperes to the LM317's maximum current-handling capability. Diodes D_1 and D_2, respectively, protect against capacitive-load discharge and polarity reversal. Depending on the circuit's requirements, IC_1 and Q_1 may require heat dissipators.□

Figure 1



A few added components extend this linear regulator's overcurrent protection.

Digitally programmable resistor serves as test load

Francesc Casanellas, Aiguafreda, Spain

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, IC_1, a DAC08, drives current-to-voltage converter IC_{2A}, which in turn drives the gate of power MOSFET Q_1. The device under test connects to J_1 and J_2. In operation, current from the device under test develops a voltage across sampling resistors R_{8A} and R_{8B}. Amplifier IC_{2B} drives IC_1's reference input and closes the feedback path. Transistor Q_2 provides overcurrent protection by diverting gate drive from Q_1 when the voltage drop across R_{8A} and R_{8B} reaches Q_2's $V_{BE(ON)}$. V_O and I_O represent the output voltage and current, respectively; N represents the decimal equivalent of the binary input applied to IC_1; and A represents the gain of the amplifier stage IC_{2B}. R_1 comprises the parallel combination of R_{1A} and R_{1B}. **Equation 1** describes the circuit's load current:

$$\frac{V_0}{R_1} = I_{OUT} = \frac{V_I}{R_6}\times\frac{N}{256} = \frac{I_0\times R_8\times A}{R_6}\times\frac{N}{256}. \qquad (1)$$

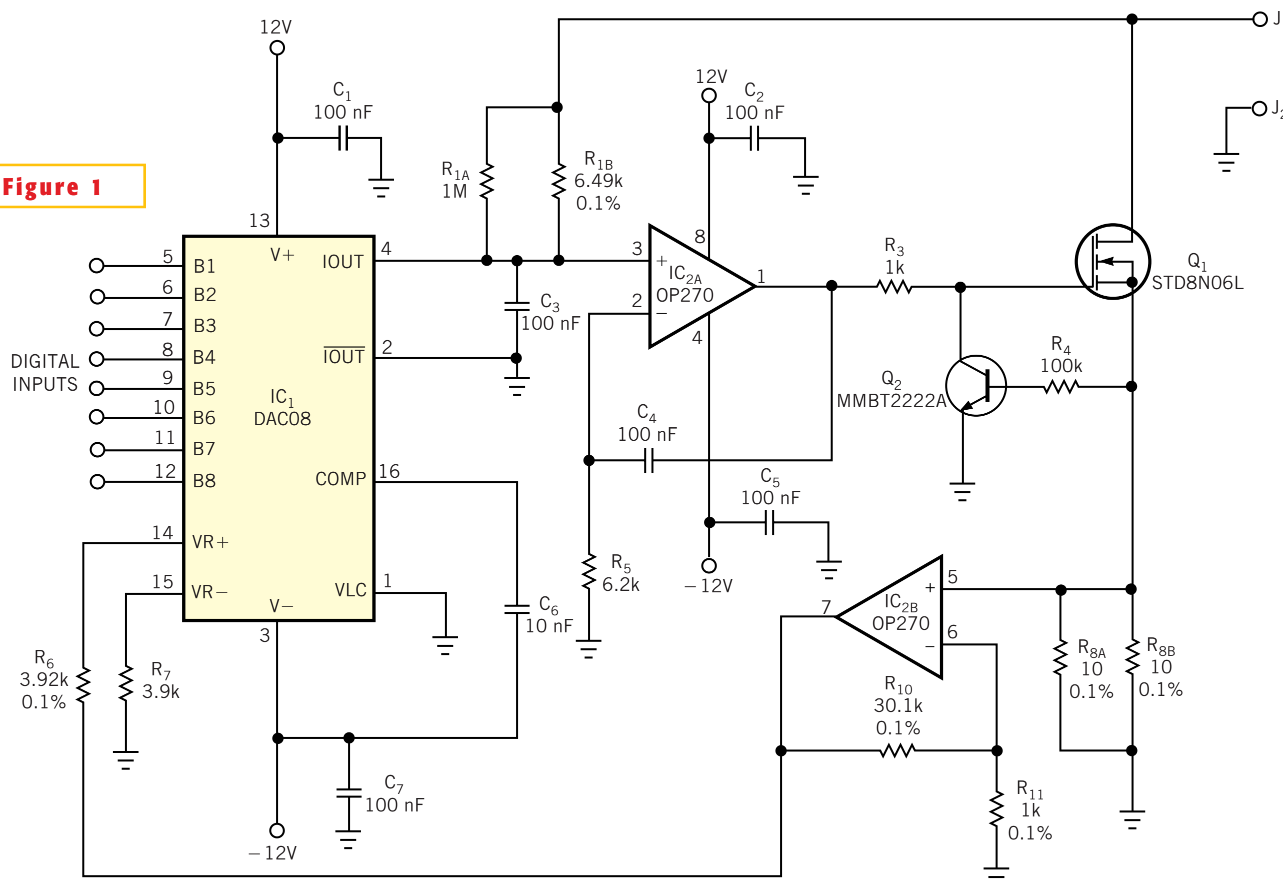
Solving **Equation 2** yields the circuit's output resistance:

$$\frac{V_0}{I_0} = \frac{A\times R_8}{R_6}\times\frac{N}{256}. \qquad (2)$$

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 Q_1 with a higher power MOSFET and an appropriately sized heat dissipator. Capacitors C_3 and C_4 control the circuit's bandwidth.□

Figure 1



This digitally programmable resistor features low component count and inexpensive parts.