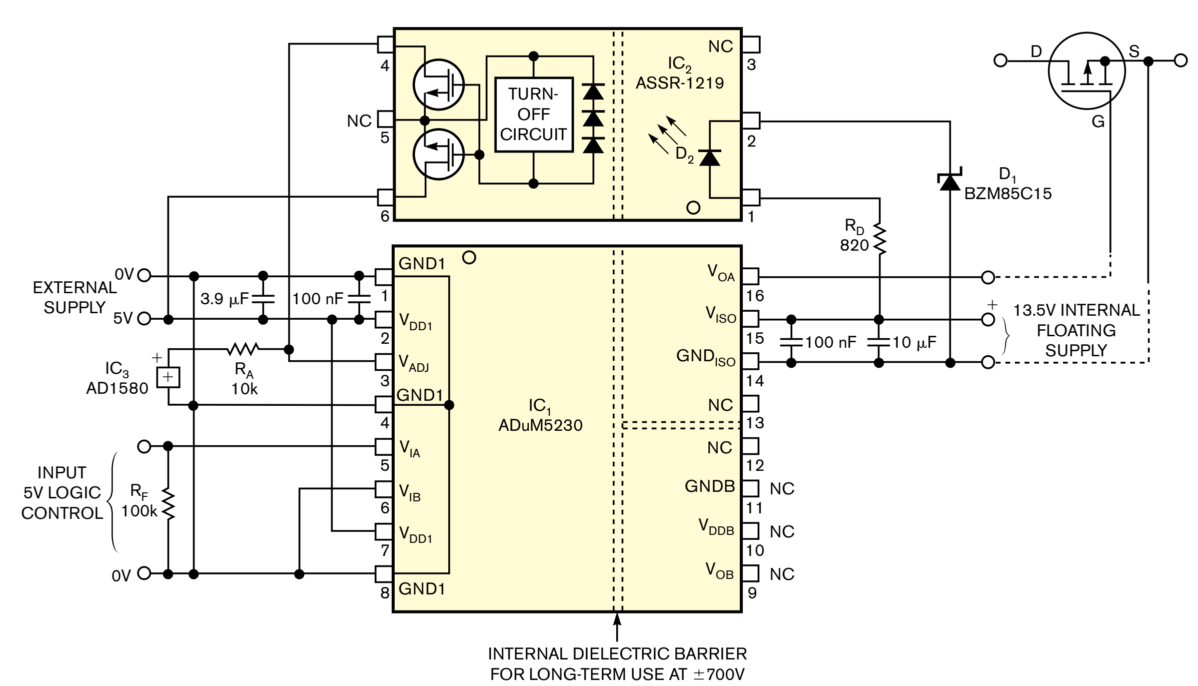


Figure 1 Connecting optical feedback by opto-MOSFET $IC_2$ in the power-MOSFET-driver $IC_1$ stabilizes the high-side output voltage to 13.5V at values of loading current down to 3.7 mA. The power efficiency of the circuit increases for a loading current of less than 7 mA.

from an almost-infinite value to a value of kilohms. The voltage level at the adjust pin then increases, and the duty factor of both the PWM in $IC_1$ decreases. This action establishes an isolated negative-voltage feedback. Thus, the temperatures of both the MOSFET and the LED in $IC_2$ have little effect on the properties of the circuit. At lighter loads, the current drain of the 5V supply is much lower than that of $IC_1$ with its adjust pin open.

Under test, the default supply current of the unloaded $IC_1$ was approximately 94.6 mA. This value decreases to 31.7 mA with the feedback in the circuit. At heavy loading, the high-side output current of $IC_1$ rises to approximately 20 mA, and the duty factor rises automatically to a proper value that's higher than at the default supply current. Thus, the output voltage is roughly 13.5V within the range of approximately 3.7 to 22.6 mA. The power efficiency of the circuit is 20% or greater. At an output current of 4.5 mA, the power efficiency is 20.5%, and the power efficiency for $IC_1$ is approximately 15%. At a current of 3.7 mA, the circuit reaches 20% efficiency, a value that's considerably higher than the 13% in $IC_1$ with its adjust pin open. **EDN**

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# Synthesize variable resistors with hyperbolic taper

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In adjustable, frequency-selective RC networks, the reciprocal of an RC product, $\omega_C=1/RC$, determines the corner frequencies of the network. If the adjustable elements are potentiometers with a linear-control characteristic—that is, taper—$R(\alpha)=\alpha R_P$, where $\alpha$ is the normalized wiper position, $0\leq\alpha\leq1$, and $R_P$ is the potentiometer's end-to-end resistance, then the corner frequencies are reciprocal functions of the potentiometer's wiper position, and the frequency scale compresses at the high end of the adjustment range. This situation is usually undesirable because it complicates adjustment of the network at the high

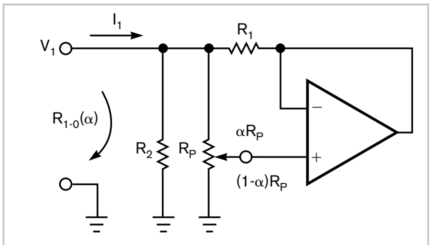


Figure 1 This simple circuit synthesizes a grounded variable resistance with a hyperbolic-control characteristic.

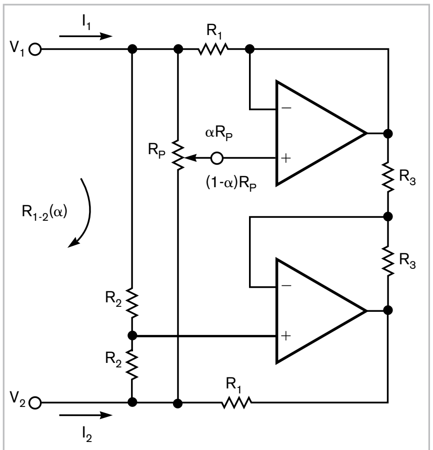


Figure 2 You can realize a floating variable resistance, with hyperbolic taper, with this circuit. Note that fixed resistors with the same number are matched pairs.

end. To make the frequency scale linear requires a control element with a hyperbolic taper—that is, something in the form $R(\alpha)=R_P/(A+\alpha B)$. Such variable resistances are not generally available from manufacturers, but you can synthesize them using a linear taper potentiometer and a few other components.

**Figure 1** shows a simple circuit for producing a ground-referenced variable resistance having the desired hyperbolic-control characteristic. Analysis of this circuit yields the following relationship between the control setting and the resistance from Node 1 to ground: $R_{1\text{-}0}(\alpha)=R_1R_2R_P/(R_1R_2+R_1R_P+\alpha R_2R_P)\,0\leqslant\alpha\leqslant 1$. If you use this resistance in series or in parallel with a capacitor, the resulting corner frequency will be a linear function of $\alpha$: $\omega_C=(R_1R_2+R_1R_P+\alpha R_2R_P)/R_1R_2R_PC$. The minimum and maximum values for $R_{1\text{-}0}$ are $R_{1\text{-}0MIN}=R_1R_2R_P/(R_1R_2+R_1R_P+R_2R_P)$ and $R_{1\text{-}0MAX}=R_2R_P/(R_2+R_P)$.

To design this circuit for specific values of $R_{1\text{-}0MIN}$ and $R_{1\text{-}0MAX}$, choose $R_P>R_{1\text{-}0MAX}$ and then compute $R_1=R_{1\text{-}0MAX}\,R_{1\text{-}0MIN}/(R_{1\text{-}0MAX}-R_{1\text{-}0MIN})$ and $R_2=R_PR_{1\text{-}0MAX}/(R_P-R_{1\text{-}0MAX})$.

You can extend the basic circuit of **Figure 1** to produce a floating variable resistance with hyperbolic taper (**Figure 2**). The value of the floating resistance between nodes 1 and 2 is $R_{1\text{-}2}(\alpha)=2R_1R_2R_P/(2R_1R_2+R_1R_P+2\alpha R_2R_P)\,0\leqslant\alpha\leqslant 1$, and the minimum and maximum values for $R_{1\text{-}2}$ are $R_{1\text{-}2MIN}=2R_1R_2R_P/(2R_1R_2+R_1R_P+2R_2R_P)$ and $R_{1\text{-}2MAX}=2R_2R_P/(2R_2+R_P)$. To design the circuit of **Figure 2** for specific values of $R_{1\text{-}2MIN}$ and $R_{1\text{-}2MAX}$, choose $R_P>R_{1\text{-}2MAX}$ and then compute $R_1=R_{1\text{-}2MAX}R_{1\text{-}2MIN}/(R_{1\text{-}2MAX}-R_{1\text{-}2MIN})$ and $R_2=\frac{1}{2}R_PR_{1\text{-}2MAX}/(R_P-R_{1\text{-}2MAX})$. Note that the value of the $R_3$ resistors does not directly affect the value of $R_{1\text{-}2}(\alpha)$. You should choose resistors that are large enough to not excessively load the op-amp outputs.

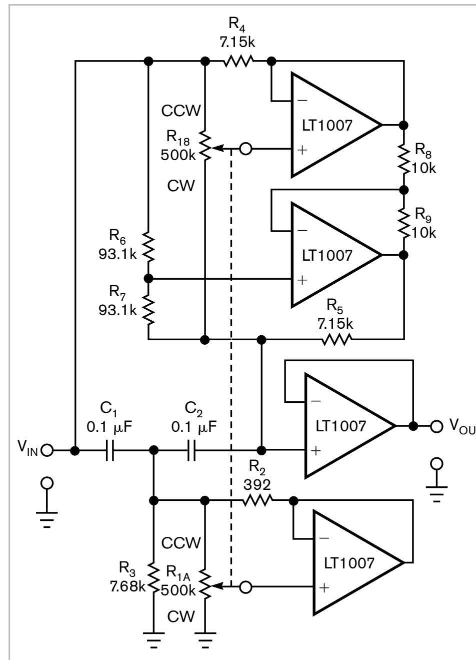


Figure 3 The basic circuits of figures 1 and 2 have been used in the design of a bridged-T notch filter with a variable notch center frequency and a linear frequency scale.

**Figure 3** illustrates the application of the circuits in **figures 1** and **2** to the design of an adjustable bridged-T notch filter with a linear frequency scale. The filter has a notch center frequency that is adjustable from 50 to 1000 Hz and a notch depth of −20 dB. These requirements and the choice of 0.1-µF capacitors for $C_1$ and $C_2$ dictate that $R_{1\text{-}0}$ varies from 375 to 7503Ω and that $R_{1\text{-}2}$ varies from 6752 to 135,047Ω. (A side benefit of using this technique is that it frees the designer from the restrictions of the limited number of standard end-to-end resistance values that potentiometer manufacturers offer.)

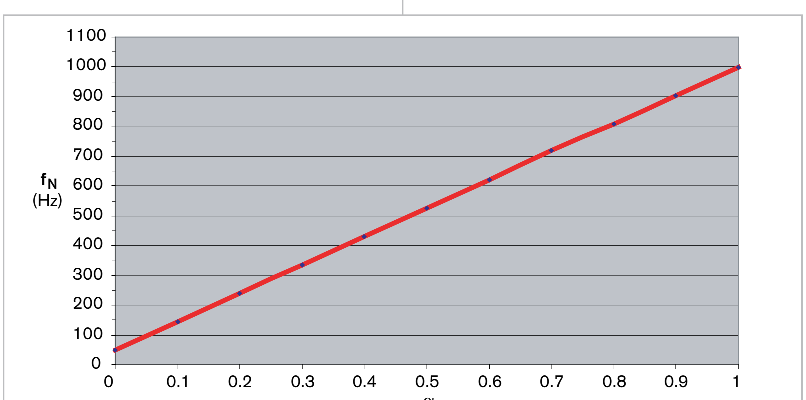


Figure 4 The Spice-simulated notch center frequency for the circuit of Figure 3 versus the normalized wiper position shows that the notch center frequency is a linear function of the control position.

**Figure 4** plots the Spice-simulated notch center frequency for the circuit of **Figure 3** against the normalized wiper position. The notch center frequency is a linear function of the control position. EDN