Design and Applications of the Hall Network

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July 4, 2011

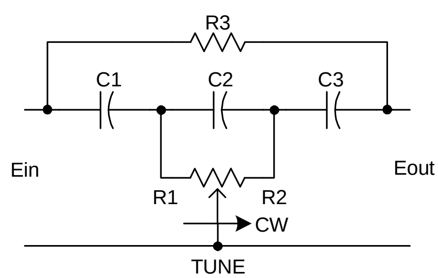
This is the original submission of an article I wrote that was published Jan. 31, 2012 in the online version of Electronic Design. See http://electronicdesign.com/analog/rediscover-truly-tunable-hall-network . This is a deep link and may get moved from time to time. If the link does not work then go to http://www.electronicdesign.com and do a search for "hall network". The published version has a different title and some minor edits in style.

Introduction

The Hall network (Reference 1) is a notch filter related to the twin-tee but has the feature of being tunable over a wide frequency range using a single potentiometer. In its passive form the notch bandwidth is broad but active electronics can narrow the response and extend the network to other applications. This article provides complete design information for the Hall network and presents three one-pot tunable applications; a notch filter, a band-pass filter, and a low-distortion sine-wave oscillator.

Hall network description and information

The Hall network is a third-order RC circuit (Figure 1). R_1 and R_2 are the split portions of the tuning potentiometer which typically includes series end resistors to prevent adjustment all the way to zero resistance where the network ceases to function. Signals at frequencies well below the notch pass through R_3 and signals at frequencies well above the notch pass through the three capacitors. At the notch frequency the separate paths have equal transmission magnitudes but opposite phase and so sum to zero. Figure 2 shows an example frequency response plot for a 1 kHz notch frequency. The phase response is approximately zero degrees at the notch.



1. Original Hall network in its simplest representation
R_1 and R_2 are the left and right portions of the potentiometer

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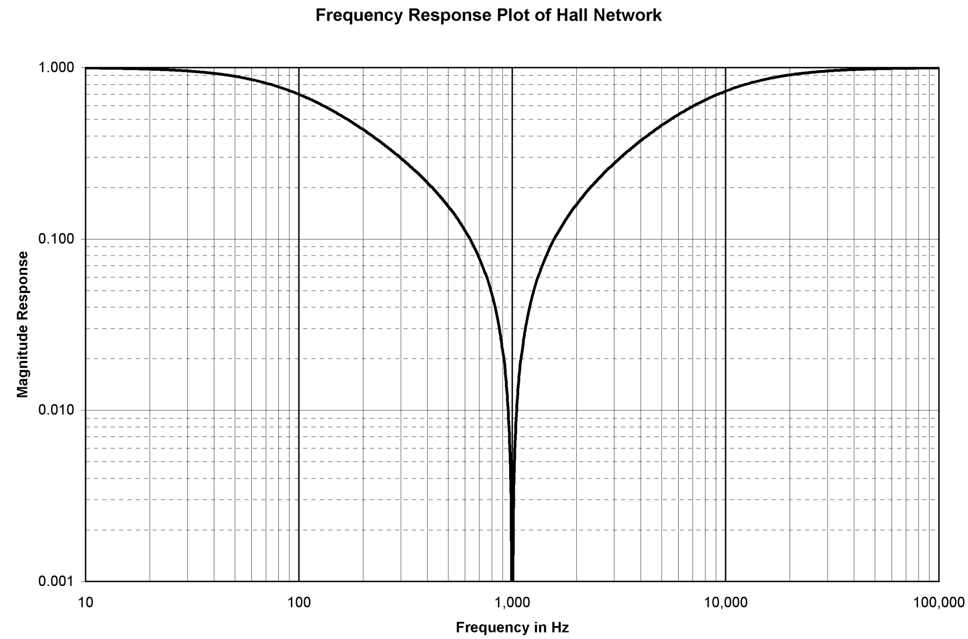


2. Frequency Response of Hall network
Although the overall response is very broad, typical for passive RC circuits, the notch is very deep.

The math (Reference 2) is arduous because the network is third order and has bridged nodes. Patient application of standard circuit methods eventually produces the three important equations. Equation 1 gives the general transfer function of the network.

$$\frac{Eout_{(s)}}{Ein_{(s)}} = \frac{R_1R_2R_3C_1C_2C_3s^3 + R_1R_2(C_1C_2 + C_2C_3 + C_1C_3)s^2 + [R_1(C_1 + C_2) + R_2(C_2 + C_3)]s + 1}{R_1R_2R_3C_1C_2C_3s^3 + [R_1R_2(C_1C_2 + C_2C_3 + C_1C_3) + R_3C_3(R_1C_1 + R_1C_2 + R_2C_2)]s^2 + [R_1(C_1+C_2) + R_2(C_2+C_3) + R_3C_3]s + 1} \qquad (1)$$

Equation 2 gives the notch frequency in Hz and is correct only if R_3 is the required value per Equation 3.

$$F = \frac{1}{2\pi\sqrt{R_1R_2\left(C_1C_2 + C_2C_3 + C_1C_3\right)}} \qquad (2)$$

For the general case of $C_3 = C_1$ and $C_1 = K * C_2$, then to achieve infinite null the required value of R_3 is given by Equation 3. For applications where the notch depth must be as deep as possible use a small series potentiometer to fine tweak R_3 to compensate for component tolerances. Once R_3 is fine tweaked it does not need further adjustment for any tuning setting. Note that R_3 affects tuning but should not be used for tuning.

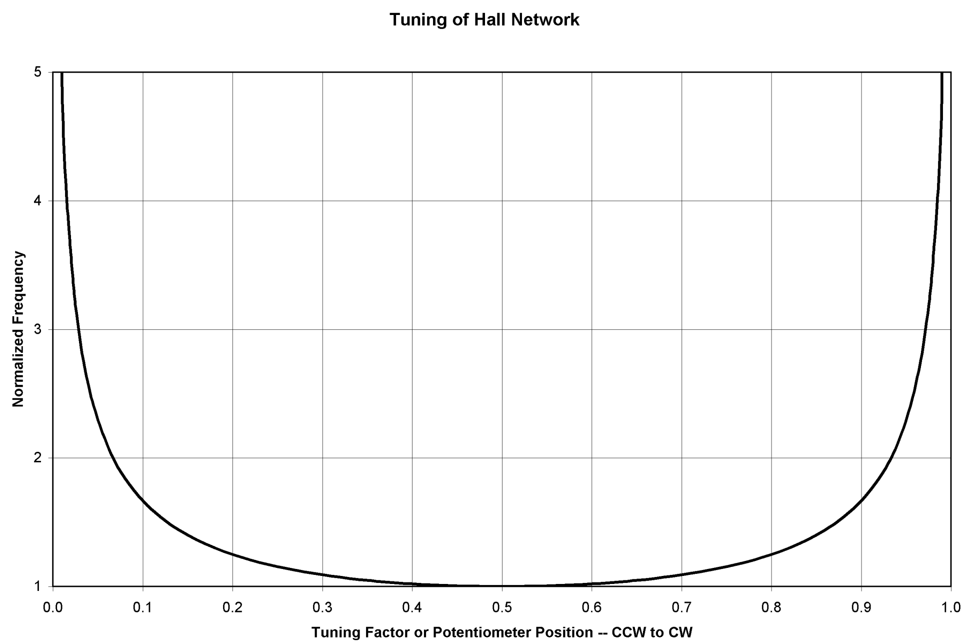
$$R_3 = \frac{(K+2)(K+1)}{K} \times (R_1 + R_2) \tag{3}$$

As a historical reference, in 1955 Mr. Hall only indicated that C_3 equals C_1 (the math confirms that ideally those would be matched as that is important for the circuit to work properly over its tuning range) but said nothing about the relationship of C_2. In his 1968 book, *Waveform Measurements* (Reference 3), Rufus P. Turner illustrated an active band-pass filter using the Hall network with $C_1 = C_3 = 10C_2$. That ratio has been used in a few other places where the Hall network has shown up. In his 1975 article in Ham Radio (Reference 4) concerning two applications of the Hall network, Courtney Hall (no relation to the designer) used all three capacitors the same value and noted that they should all be matched for optimum notch depth. The only known commercial use of the Hall network is as a tunable band-pass filter for the 1961 General Radio product, Type 1232-A Tuned Amplifier and Null Detector (References 5 and 6). An examination of the manual for this instrument reveals that all capacitors are the same value. This product is still sold today by IET Labs, Inc. (http://www.ietlabs.com/) as part of the GenRad 1620 Capacitance Bridge. The minimum notch bandwidth (i.e. highest Q) occurs when all capacitors are the same value (K = 1) and that ratio is used for all examples in this article.

Tuning and Q characteristics

The tuning characteristic (Equation 4 and Figure 3) is very shallow near the middle of the potentiometer and very steep near the ends. This curve was derived by a modified version of Equation 2 where R_1 was replaced with a tuning factor (TF) that varies from 0 to 1 as the pot is rotated from full counter-clockwise to full clockwise and R_2 is replaced with 1 – TF as follows. The 0.5 factor normalizes the minimum frequency at the center of pot rotation to 1.0.

$$F_{normalized} = \frac{0.5}{\sqrt{TF \times (1 - TF)}} \tag{4}$$



3. Hall network tuning characteristic
Note that tuning is symmetric about the middle point of the potentiometer

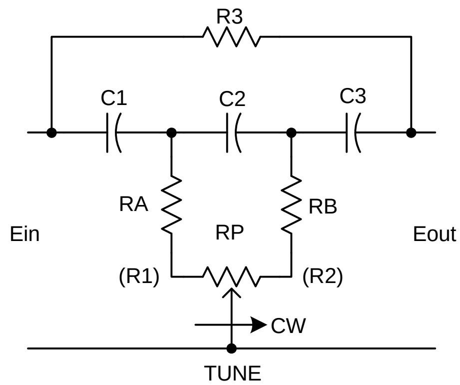
The Q (resonant frequency divided by the -3 dB bandwidth) of any passive RC network is low. With three identical capacitors the Q of the Hall network is a maximum of 0.177 when the tuning factor is 0.5 and decreases to zero when the tuning factor is 0.0 or 1.0 (Equation 5). Although active electronics can increase Q up to a practical maximum of several tens, the best results for applications requiring high Q will be when the tuning factor is in roughly the 0.1 to 0.9 range.

$$Q = 0.25\sqrt{TF} \qquad 0.0 <= TF <= 0.5 \qquad \text{(5a)}$$

$$Q = 0.25\sqrt{1 - TF} \qquad 0.5 <= TF <= 1.0 \qquad \text{(5b)}$$

Design for tuning range

Since the tuning curve is symmetric about the mid point of $R_1 + R_2$ it makes sense to add resistances (Figure 4) to either side of the potentiometer to limit the adjustment range to the region between 0.5 and 1.0. By symmetry the 0.5 to 0.0 region could have been used alternatively but it makes more intuitive sense to the author to pick the right side of the plot.



4. Modified Hall network with end resistors to limit tuning range to right hand portion of Figure 3

Depending on the application, the tuning ratio (the maximum divided by the minimum desired frequencies over the full range of the potentiometer) might range from 1.1 or less up to a practical maximum of about 4 with a linear potentiometer. High tuning ratios must utilize the steep portion of the tuning curve (Figure 3**)** and become impractical as tuning is very compressed at the high end. Special taper potentiometers can extend the practical tuning ratio to over ten.

The choice of starting point on the tuning curve affects tuning linearity. For any tuning ratio, starting at 0.5 maximizes the tuning non-linearity. Using higher starting points reduces tuning non-linearity although a point of diminishing returns is reached at around 0.7. With a starting tuning factor of 0.7 tuning ratios less than about 1.2 result in fairly linear tuning and tuning ratios up to around 3 are not excessively compressed at the high end.

The starting point in determining the end resistors is to establish the tuning factor range. For good results, choose the starting or low tuning factor (TF_{low}) to be around 0.7 when the potentiometer wiper is at the full counter-clockwise position and then use Equation 6 to calculate the required high tuning factor (TF_{high}) to achieve the desired tuning ratio when the wiper is at the full clockwise position. Equation 6 was derived from Equation 4 by first normalizing the result to unity for TF_{low} and then solving the resulting quadratic equation for TF_{high} with Equation 4 set to the desired tuning_ratio.

$$TF_{high} = \frac{1+\sqrt{1-\dfrac{4\times TF_{low}\times\left(1-TF_{low}\right)}{tuning\_ratio^2}}}{2} \qquad (6)$$

With the extreme tuning factors known the next step is to determine a good resistance for the potentiometer, R_P. Equation 7 was empirically derived to provide a geometric center estimate for a practical resistance for potentiometer based on experience that for a given frequency there is a range of R and C values that provides good results. Frequency is nominally the geometric center of the tuning range in Hz but can be any frequency within the tuning range as there is broad flexibility in using the result.

$$R_P = \frac{300{,}000}{\sqrt{Frequency}} \tag{7}$$

Choose a convenient value for potentiometer, R_P, between roughly one-fourth to about four times the geometric center value from Equation 7. The end resistors, R_A and R_B, are then calculated as follows.

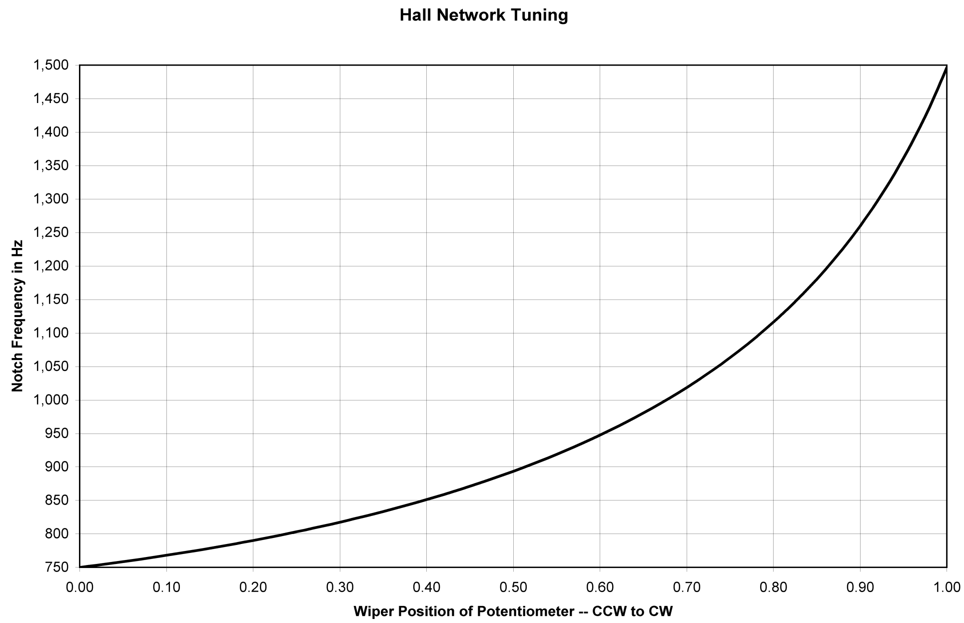
$$R_{total} = \frac{R_P}{TF_{high} - TF_{low}} \tag{8}$$

$$R_A = R_{total} \times TF_{low} \tag{9}$$

$$R_B = R_{total} \times \left(1 - TF_{high}\right) \tag{10}$$

For the examples in this article a Hall network was designed to tune from 750 to 1,500 Hz which is a tuning ratio of 2. TF_{low} was chosen to be 0.7 and TF_{high} computed to be 0.944. The geometric center for R_P computed to be about 9.5-kΩ so a 10-kΩ potentiometer was chosen. R_{total} computed to be 40,917 ohms, R_A computed to be 28,642 ohms, and R_B computed to be 2,291 ohms. R_A was rounded to a standard value of 27-kΩ and R_B was rounded to a standard value of 2.2-kΩ. The actual tuning ratio was computed (Equation 11) to be 2.01 with these values. If the rounded resistor values produce a tuning ratio too different from desired then adjust R_B up or down a standard value as needed. The tuning curve is shown in Figure 5.

$$tuning\_ratio = \sqrt{\frac{R_A\left(R_P + R_B\right)}{\left(R_A + R_P\right)R_B}} \tag{11}$$



5. Tuning curve for example Hall network
Although tuning is non-linear, it is not too bad.

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Design for capacitors

For three identical capacitors Equations 2 and 3 for the notch frequency and R_3 simplify to Equations 12 and 13.

$$F = \frac{1}{2\pi\sqrt{3R_1R_2} \times C} \tag{12}$$

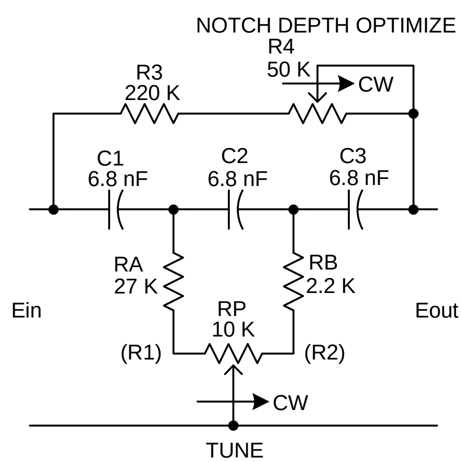
$$R_3 = 6(R_1 + R_2) \tag{13}$$

The required capacitance (Equation 14) for each of the three capacitors is calculated by inverting Equation 12 and noting that at the minimum notch frequency (full counter-clockwise of the potentiometer), $R_1 = R_A$ and $R_2 = R_P + R_B$.

$$C = \frac{1}{2\pi F_{NOTCH\_MINIMUM}\sqrt{3R_A(R_P + R_B)}} \tag{14}$$

For the minimum frequency of 750 Hz the capacitance computed to be 6.75 nF which was rounded to the standard value of 6.8 nF.

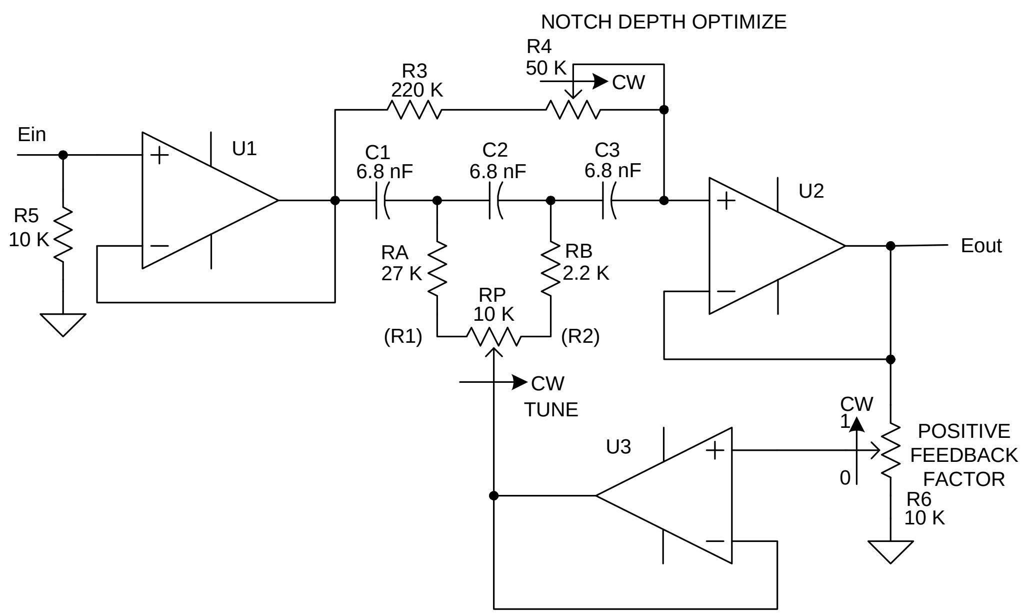
From Equation 13 the value for R_3 for theoretical infinite notch computed to be 6 * (27,000 + 10,000 + 2,200) = 235,000 ohms which could be rounded to the standard value of 240-kΩ for applications where optimum notch depth is not required. For the circuits in this article R_3 was 220-kΩ in series with a 50-kΩ series potentiometer and adjusted for optimum notch depth. The completed design (Figure 6) is now ready to be used in three useful applications.



6. Hall network used for all examples in this article
Tunes from 750 Hz to 1,500 Hz

Active tunable notch filter

Figure 7 shows an active tunable notch filter with the ground terminal of the Hall network bootstrapped with adjustable positive feedback from the output. The positive feedback factor (PFF) varies from 0 when R_6 is at full counter-clockwise to 1.0 when R_6 is at full clockwise. Increasing the positive feedback factor boosts the Q of the circuit thus narrowing the notch bandwidth. The resulting Q is the basic Hall network Q at the given tuning factor (Equation 5) multiplied by the Q boost factor (Equation 15). The practical maximum for Q is around 30 as adjustment for higher values becomes very touchy.



7. Active notch filter
Positive feedback is applied to bootstrap ground end of Hall network to increase Q.

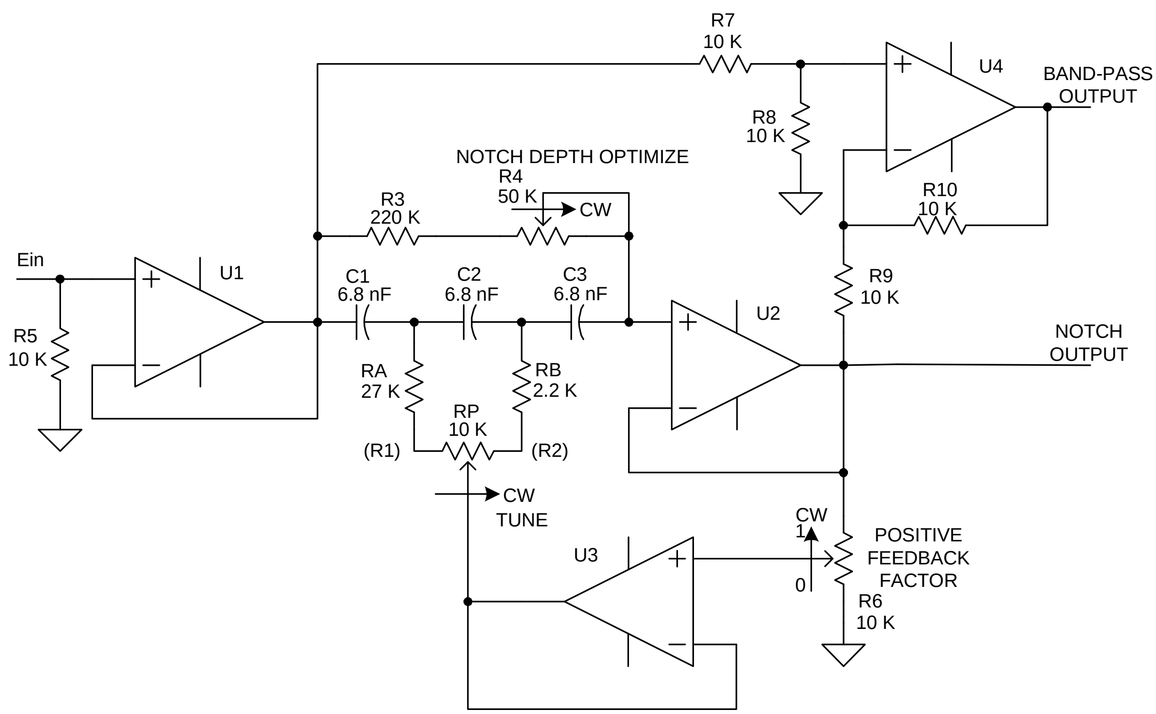
$$Q_{BOOST\_FACTOR} = \frac{1}{1 - PFF} \tag{15}$$

Unfortunately, notch depth in a physical circuit becomes shallower with increasing Q even with the best setting of the optional Notch Depth Optimize adjustment. The mathematical model does not show this effect. Although not proven, it is suspected that small phase lags in U_2 and U_3 are a contributing factor and some kind of phase compensation would improve operation. For the prototype circuit the measured null depth at 1 kHz with PFF = 0 was -54 dB and was -45 dB with PFF = 0.9.

The operational amplifiers used for this example were TL084 although most any op-amp would work fine. When the Positive Feedback Factor is zero (R_6 at full counter-clockwise) the circuit is simply the standard Hall network with unity gain buffers. As the potentiometer is turned clockwise the notch width narrows until at unity positive feedback the circuit has no filtering and might oscillate. Except for the notch region this circuit has unity gain for any setting of the Tuning control or Positive Feedback Factor.

Active tunable band-pass filter

A band-pass filter is formed by subtracting the output of the notch filter from the input (Figure 8). For best results resistors R_7, R_8, R_9, and R_{10} should be matched.



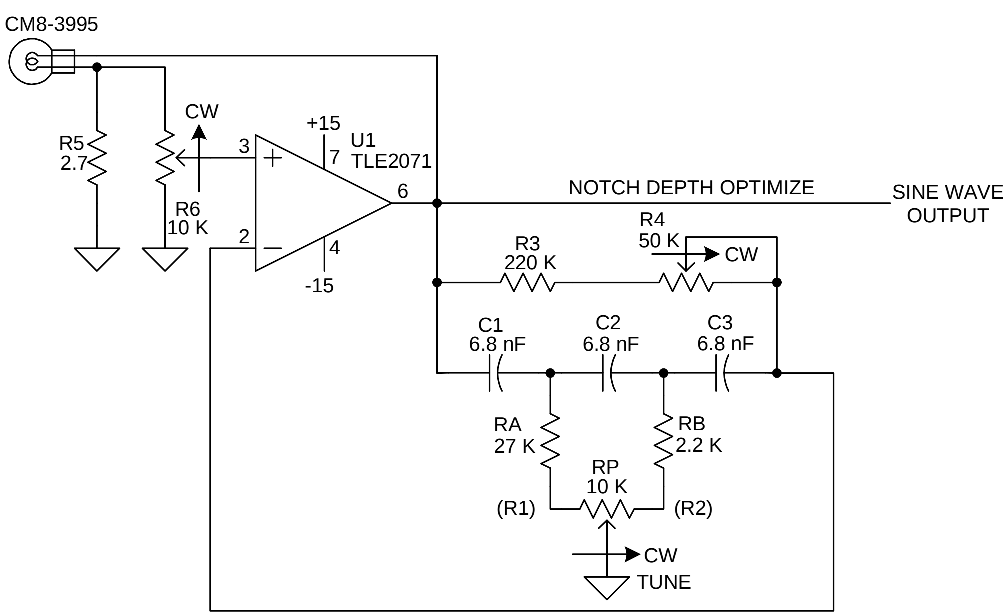
8. Band-pass and Notch Filter
Note that the band-pass is formed by subtracting the notch output from the input.

A useful characteristic of this band-pass filter is that the peak response at resonance is 1.0 independent of Q as set by the Positive Feedback Factor. This means that as Q is adjusted signals at the resonant frequency remain unchanged in amplitude while the amplitude of signals at frequencies away from resonance varies. The notch output is simultaneously available. The Q of the band-pass is identical to that of the notch.

One-pot tunable low-distortion sine wave oscillator

A low-distortion one-pot tunable sine wave oscillator (Figure 9) is constructed by combining the Hall network with the light bulb from Bill Hewlett's famous oscillator (References 7 and 8). The resistance of the lamp filament increases with the thermally averaged rms voltage across it (Figure 10) and is used to control the positive feedback gain to the correct level. The thermal time constant of the filament is long compared to a half-cycle of oscillation so its resistance remains constant throughout the oscillation waveform – an important characteristic for low distortion. Before oscillation commences there is no voltage across the lamp filament so its cold resistance is relatively low. Thus, the positive feedback path voltage divider gain to the U_1 non-inverting input is relatively high. The negative feedback path gain at the notch frequency is very low. Since the positive feedback path gain exceeds the negative feedback path gain the system poles are in the right-half s-plane and oscillation grows exponentially. The lamp filament then heats with the increasing oscillation voltage across it. Heating causes its resistance to increase thus reducing the positive feedback path gain and the system poles shift leftwards.

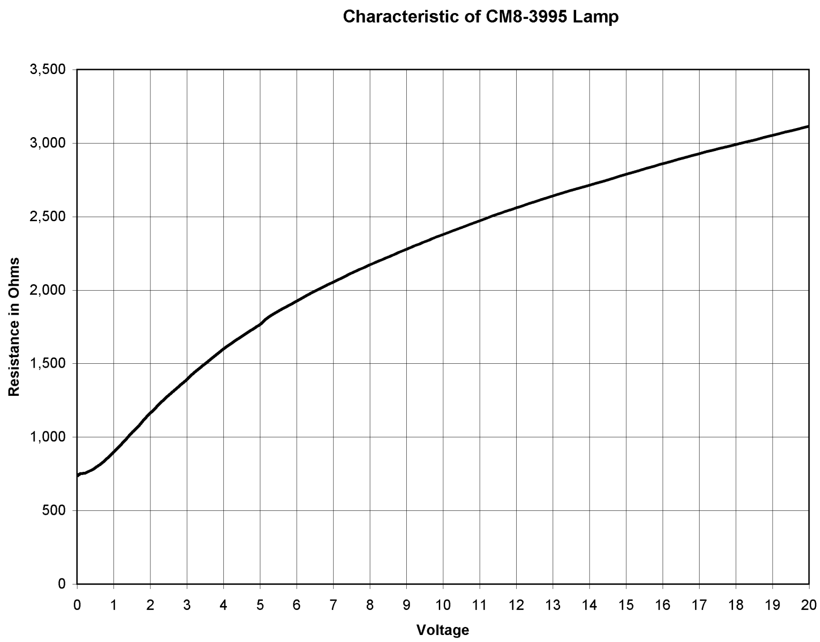
There are two effects of this thermal control loop. (1) Oscillation stabilizes at the amplitude where the positive and negative feedback path gains are equal and the system poles are exactly on the jw axis. (2) The amplitude is regulated by the filament resistance-voltage characteristic as there is a unique voltage that results in equal feedback gains.



9. Low-distortion sine wave oscillator
The lamp filament resistance forms a voltage divider with R_5 in the positive feedback path that increases voltage division with oscillation signal.

Bill determined that a good operating point on the lamp resistance-voltage curve is in the lower power region where the change in resistance with applied voltage is high as this provides high stabilization gain. A further advantage of this operating level is that the thermal time constant of the filament is long compared to higher operating voltages so distortion is lower. His oscillator used a 120 volt, 3-watt lamp operating at about 7 Vrms. Correspondingly, the lamp (Chicago Miniature 8-3995, 130 V, 20 mA in a common T-

3¼ miniature bayonet base) in this application was chosen to operate at about 7 Vrms. The lamp produces no visible light at this low power level.



10. Plot of resistance versus voltage for the CM8-3995 lamp
Note that the change in resistance with applied voltage is high at low voltages and decreases for higher voltages

A Texas Instruments TLE2071 was chosen for U_1 because it is capable of higher than typical output current to drive the lamp and also features wider than typical bandwidth. The amplifier gain at the oscillation frequency is many hundreds so the gain-bandwidth product of the amplifier must be sufficiently high – typically over one thousand times the oscillation frequency.

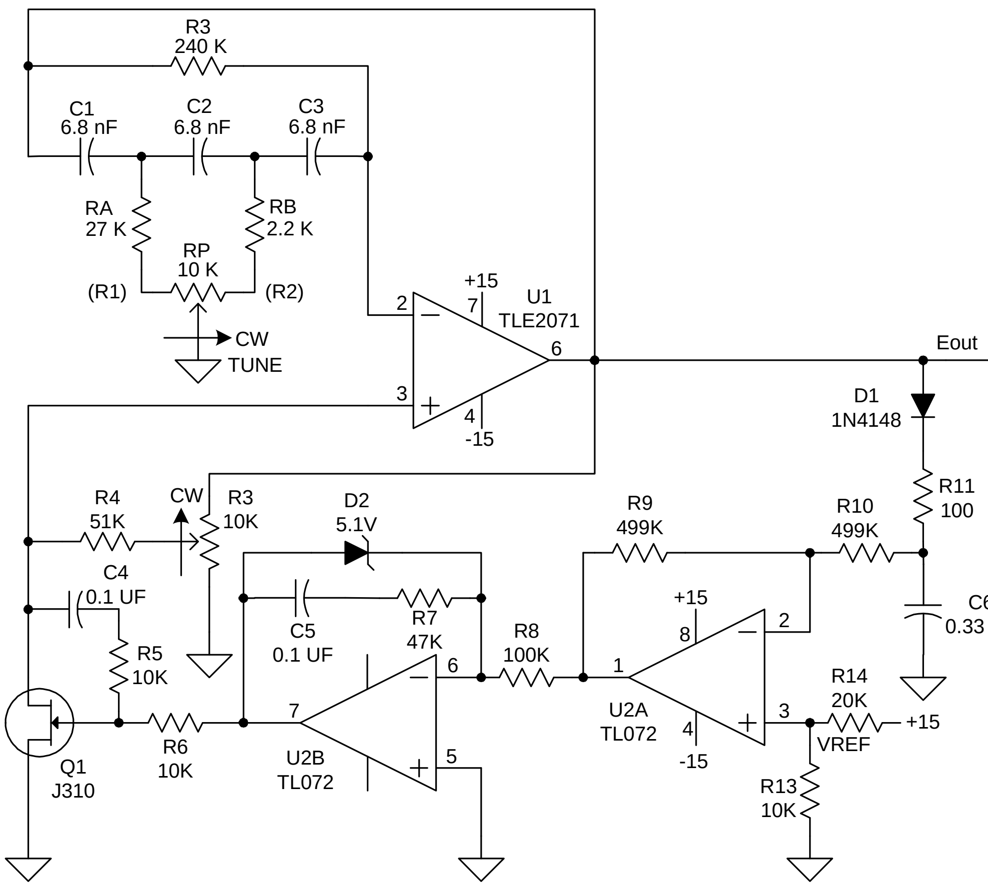
Because of the low resistances in the positive feedback path it is important that the ground impedance in the circuit be very low. Otherwise the high gain at the oscillation frequency will amplify small voltage drops across the ground to the point that it may be difficult to set the oscillation amplitude. Depending on the particular ground impedance and connections the circuit might oscillate to the full rail-rail output voltage of U_1 independent of the setting of R_6 or might have a hysteresis effect where the circuit either oscillates or not depending on the setting of R_6 but no setting will stabilize the amplitude at a desired level. One quick (but should not be permanent) remedy short of improving the ground system should these problems occur is to tweak R_4 off of the theoretical ideal value as this will make the notch shallower thus requiring less gain for oscillation. Each of these problems was experienced with various prototype versions of the oscillator. However, the circuit works very well if constructed properly.

The initial setting of R_6 should be full clockwise as this will insure that oscillation starts although the waveform will be clipped at the op-amp rails (use a higher resistance for R_5 if oscillation does not start.). Then adjust R_6 counter-clockwise until the output amplitude is a sine wave about 20 Vpp. The thermal servo system of the filament is under-damped so the amplitude will fluctuate for a number of seconds before stabilizing. The ultra low distortion sine wave (less than 0.003%) is worth the wait.

For different designs, choose a lamp with a rated voltage around twenty, give or take, times the target rms output voltage. Choose a lamp with the lowest rated current for a given voltage to ease current drive requirement of the op-amp. Then determine the resistance of the filament at the target operating voltage. Choose R_5 to be roughly in the range of one three-hundredth to one thousandth of that filament resistance. Use R_6 to fine adjust the operating point. Be sure the op-amp has sufficient output current to drive the lamp peak current and a large signal gain-bandwidth product at least a thousand times higher than the frequency of oscillation.

An undesirable characteristic of this circuit is that the sine wave amplitude varies with tuning. This is because the transfer gain of the Hall network at the notch frequency varies with tuning. For stable oscillation the positive feedback factor must also vary by the same amount. This means that as the Hall network transfer gain decreases the oscillation amplitude must increase to cause the positive feedback factor to decrease by the same amount – i.e. the resistance of the filament must increase. Depending on the tuning range and specific values of the components the amplitude variation with tuning could range from ceasing to oscillate to clipping at the op-amp power rails. R_6 can always be adjusted to compensate but that is inconvenient. The lamp stabilized oscillator is attractive for its simplicity if the tuning range is narrow or to be permanently set at a specific frequency.

The only way to obtain stable oscillation amplitude over a wide frequency range is to use an electronic servo system to control the gain instead of the lamp. This approach achieves almost the low distortion of the lamp circuit while rapidly settling to the amplitude set point. Because of the much higher impedances in the positive feedback path, this circuit is practically immune to the ground impedance issues of the lamp version. A sample circuit (Reference 9 and Figure 11) was adapted from the amplitude control for the Hewlett-Packard model 239A oscillator.



11. JFET controlled sine-wave oscillator
The JFET is used in its ohmic region as a voltage controlled resistor to manage positive feedback.

The peak of the sine wave is detected by components D_1, R_{11}, and C_6. R_{11} limits peak capacitor charging current from the op-amp to prevent distortion. The output of U_{2A} is the control error voltage which is the detected peak voltage minus twice the reference voltage (VREF = 5 volts for this example) at the non-inverting input. Thus, at zero error the peak sine wave output voltage is twice the reference voltage plus the diode drop or about 10.6 volts for this example. U_{2B} is a proportional plus integral controller that applies a variable gate bias to JFET Q_1 which operates in its ohmic region as a voltage controlled resistor. The integral gain is $1/(R_8{*}C_5)$ and the proportional gain is R_7/R_8. R_7 is used to adjust the damping and is a compromise between not too under-damped and low distortion. Ideally, the channel resistance of Q1 would be constant throughout a cycle of oscillation but the resistance is modulated by the ripple voltage on C_6 amplified by the proportional gain. The lowest distortion occurs when R_7 is as small as can be tolerated (i.e. lowest proportional gain and lowest damping) so that the settling time when changing frequencies is not too oscillatory. Zener D_2 limits the voltage swing of U_{2B} to the gate control range of Q_1. The zener voltage may need to be adjusted depending on the characteristics of a particular JFET.

A type J310 JFET was used for this example but most any JFET should work well with appropriate adjustments to R_4 and perhaps D_2. The channel resistance is high when the gate-source voltage is near pinch-off and low when the gate-source voltage is near zero. The required attenuation factor in the positive feedback path of U_1 is many hundreds and is divided into two parts. Part one consists of the voltage division of potentiometer, R_3, and part two consists of the voltage division between R_4 and the resistance of the JFET channel. Breaking the attenuation into two parts enables a single relatively low resistance potentiometer value to be used to scale R_4 up to whatever feedback resistance is required. If the oscillation amplitude is below the set-point then the error output voltage of U_{2A} is positive which then drives the output of U_{2B} negative resulting in a high channel resistance for Q_1. This makes the positive feedback factor relatively high thus causing the oscillation amplitude to increase until the error voltage is zero.

In the ohmic region of operation, the resistance of the JFET channel is fairly linear for small voltages across the channel less than a few tenths of a volt positive or negative. C_4, R_5, and R_6 form an AC voltage divider that couples half of the small signal at the drain of the JFET to its gate to significantly improve linearity for low distortion (Reference 10).

The adjustment range of potentiometer, R_3, is divided into three regions:

- Oscillation will not occur if R_3 is too counter-clockwise. U_{2B} will drive the JFET gate bias towards pinch-off in an attempt to maximize the gain.

- Oscillation amplitude will not be controlled and the peaks may go into clipping if R_3 is too clockwise. U_{2B} will drive the JFET gate bias positive in an attempt to minimize the gain.

- Between the excessive conditions is a zone where the oscillation amplitude is controlled and sine wave amplitude is independent of the R_3 setting. The optimum setting for lowest distortion is at the onset of where the output of U_{2B} starts to quickly go negative. This should be checked over the entire tuning range and R_3 adjusted for best compromise so that the output of U_{2B} is always near zero volts preferably not positive more than a tenth of a volt or so. The resistance of R_4 has to be selected based on the characteristics of the particular JFET. The useful adjustment range will be very narrow if R_4 is too small. The circuit will not oscillate if R_4 is too large. Although there is broad leeway, a good value for R_4 is such that the optimum adjustment of R_3 is about one-third to one-half up from the ground end. This value has to be experimentally determined.

The distortion of both circuits was measured at 1 kHz using an H-P model 3580A spectrum analyzer. The amplitude of the fundamental was adjusted to be at the 0 dB reference level. The sweep range was 0 to 10 kHz and the resolution bandwidth was 30 Hz. With the lamp version no harmonic was visible above the instrument noise floor 90 dB below the reference level. Thus, distortion is less than 0.003 percent. The JFET version showed a slight indication at the second harmonic.

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There is some interesting history associated with these two oscillator circuits. Bill used the lamp filament to increase negative feedback to stabilize his Wien bridge oscillator while the lamp filament in the Hall circuit decreases positive feedback to stabilize the oscillation. It is also interesting to note that the lamp control is from H-P's first oscillator and the JFET control is from H-P's last purely analog oscillator before digital synthesis took over in the 1980s.

One question that arises concerning notch type oscillators is how they work at all. Considering that at the notch frequency the transfer gain is theoretically zero even an infinite gain op-amp would be insufficient. The solution to this apparent dilemma is that the net zero phase operating frequency including the small time delay through the op-amp is not precisely the notch frequency. Thus, there is a finite, albeit small, transmission through the network. The op-amp then amplifies that small voltage by many hundreds to produce the sine-wave output.

Conclusion

The Hall network accomplishes the often sought after feat over many years of being truly tunable over a wide frequency range using a single potentiometer. It is a mystery to the author why the Hall network did not attain popularity many years ago. Perhaps this article will provide the Hall network with a new opportunity to become widely known.

Acknowledgement

I would like to give a big personal thanks to the designer of the Hall network, Mr. Henry P. Hall, now retired from a distinguished career at General Radio and recipient of the IEEE Joseph F. Keithley Award in Instrumentation and Measurement in 2004, who graciously provided me with helpful information, inspiration, and encouragement.

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Author's Biography

Kenneth Kuhn is a senior electrical engineer at Xylem Inc. | OI Analytical / CMS Field Products in Pelham, Alabama and designs electronics for chemical analysis equipment. He obtained his B.S.E.E. at Auburn University and his M.S.E.E. at the University of Alabama, Birmingham where he is also on the adjunct faculty teaching evening courses in electronics. His home electronics shop includes a large collection of vintage Hewlett-Packard test equipment which can be seen on his virtual H-P museum at http://www.kennethkuhn.com/hpmuseum. Other interests include taking care of his pet cats, investing, and listening to and composing modern classical music.