

## LM13600 Dual Operational Transconductance Amplifiers with Linearizing Diodes and Buffers

### General Description

The LM13600 series consists of two current controlled transconductance amplifiers each with differential inputs and a push-pull output. The two amplifiers share common supplies but otherwise operate independently. Linearizing diodes are provided at the inputs to reduce distortion and allow higher input levels. The result is a 10 dB signal-to-noise improvement referenced to 0.5 percent THD. Controlled impedance buffers which are especially designed to complement the dynamic range of the amplifiers are provided.

### Features

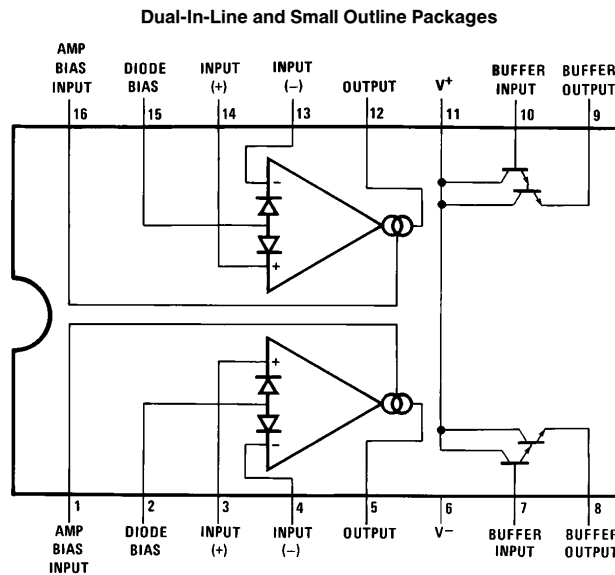
- $g_m$  adjustable over 6 decades
- Excellent  $g_m$  linearity

- Excellent matching between amplifiers
- Linearizing diodes
- Controlled impedance buffers
- High output signal-to-noise ratio

### Applications

- Current-controlled amplifiers
- Current-controlled impedances
- Current-controlled filters
- Current-controlled oscillators
- Multiplexers
- Timers
- Sample and hold circuits

### Connection Diagram



TL/H/7980-2

Order Number LM13600M, LM13600N or LM13600AN  
See NS Package Number M16A or N16A

## Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (Note 1)		
LM13600	36 V <sub>DC</sub> or ±18V	
LM13600A	44 V <sub>DC</sub> or ±22V	
Power Dissipation (Note 2) T <sub>A</sub> = 25°C	570 mW	
Differential Input Voltage	±5V	
Diode Bias Current (I <sub>D</sub> )	2 mA	
Amplifier Bias Current (I <sub>ABC</sub> )	2 mA	
Output Short Circuit Duration	Continuous	
Buffer Output Current (Note 3)	20 mA	

Operating Temperature Range	0°C to +70°C
DC Input Voltage	+V <sub>S</sub> to -V <sub>S</sub>
Storage Temperature Range	-65°C to +150°C
Soldering Information	
Dual-In-Line Package	
Soldering (10 seconds)	260°C
Small Outline Package	
Vapor Phase (60 seconds)	215°C
Infrared (15 seconds)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" for other methods of soldering surface mount devices.

## Electrical Characteristics (Note 4)

Parameter	Conditions	LM13600			LM13600A			Units
		Min	Typ	Max	Min	Typ	Max	
Input Offset Voltage (V <sub>OS</sub> )	Over Specified Temperature Range I <sub>ABC</sub> = 5 μA		0.4	4		0.4	1	mV
			0.3	4		0.3	2	mV
							1	mV
V <sub>OS</sub> Including Diodes	Diode Bias Current (I <sub>D</sub> ) = 500 μA		0.5	5		0.5	2	mV
Input Offset Change	5 μA ≤ I <sub>ABC</sub> ≤ 500 μA		0.1	3		0.1	1	mV
Input Offset Current			0.1	0.6		0.1	0.6	μA
Input Bias Current	Over Specified Temperature Range		0.4	5		0.4	5	μA
			1	8		1	7	μA
Forward Transconductance (g <sub>m</sub> )	Over Specified Temperature Range	6700	9600	13000	7700	9600	12000	μmho
		5400			4000			μmho
g <sub>m</sub> Tracking			0.3			0.3		dB
Peak Output Current	R <sub>L</sub> = 0, I <sub>ABC</sub> = 5 μA		5		3	5	7	μA
	R <sub>L</sub> = 0, I <sub>ABC</sub> = 500 μA	350	500	650	350	500	650	μA
	R <sub>L</sub> = 0, Over Specified Temp Range	300			300			μA
Peak Output Voltage	R <sub>L</sub> = ∞, 5 μA ≤ I <sub>ABC</sub> ≤ 500 μA	+12	+14.2		+12	+14.2		V
		-12	-14.4		-12	-14.4		V
Supply Current	I <sub>ABC</sub> = 500 μA, Both Channels		2.6			2.6		mA
V <sub>OS</sub> Sensitivity	ΔV <sub>OS</sub> /ΔV+		20	150		20	150	μV/V
			20	150		20	150	μV/V
CMRR		80	110		80	110		dB
Common Mode Range		±12	±13.5		±12	±13.5		V
Crosstalk	Referred to Input (Note 5) 20 Hz < f < 20 kHz		100			100		dB
Differential Input Current	I <sub>ABC</sub> = 0, Input = ±4V		0.02	100		0.02	10	nA
Leakage Current	I <sub>ABC</sub> = 0 (Refer to Test Circuit)		0.2	100		0.2	5	nA

## Electrical Characteristics (Note 4) (Continued)

Parameter	Conditions	LM13600			LM13600A			Units
		Min	Typ	Max	Min	Typ	Max	
Input Resistance		10	26		10	26		k $\Omega$
Open Loop Bandwidth			2			2		MHz
Slew Rate	Unity Gain Compensated		50			50		V/ $\mu$ s
Buffer Input Current	(Note 5), Except $I_{ABC} = 0 \mu\text{A}$		0.2	0.4		0.2	0.4	$\mu\text{A}$
Peak Buffer Output Voltage	(Note 5)	10			10			V

**Note 1:** For selections to a supply voltage above  $\pm 22\text{V}$ , contact factory.

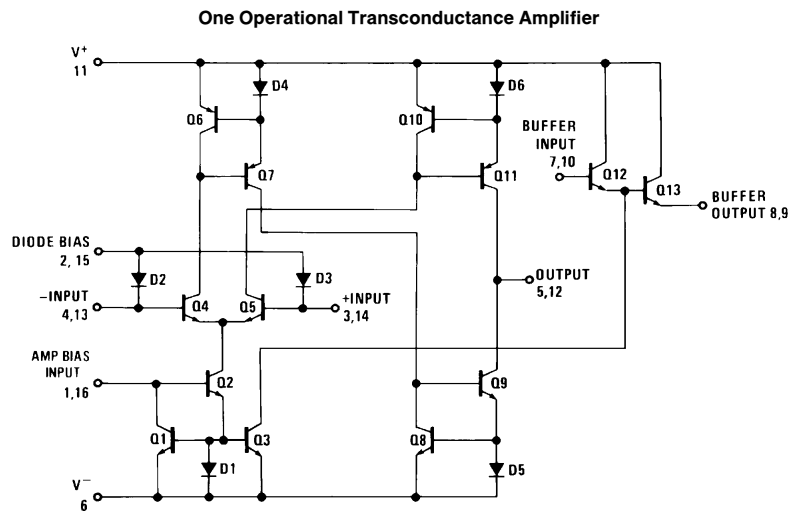
**Note 2:** For operating at high temperatures, the device must be derated based on a  $150^\circ\text{C}$  maximum junction temperature and a thermal resistance of  $175^\circ\text{C}/\text{W}$  which applies for the device soldered in a printed circuit board, operating in still air.

**Note 3:** Buffer output current should be limited so as to not exceed package dissipation.

**Note 4:** These specifications apply for  $V_S = \pm 15\text{V}$ ,  $T_A = 25^\circ\text{C}$ , amplifier bias current ( $I_{ABC}$ ) =  $500 \mu\text{A}$ , pins 2 and 15 open unless otherwise specified. The inputs to the buffers are grounded and outputs are open.

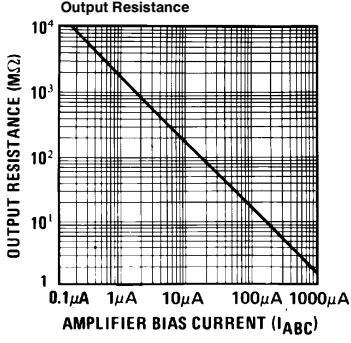
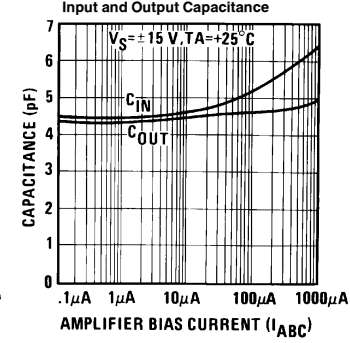
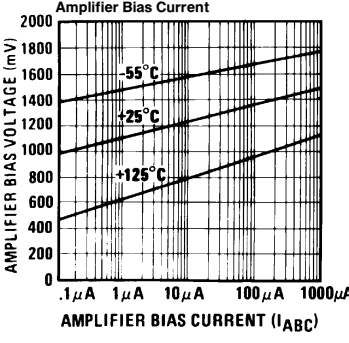
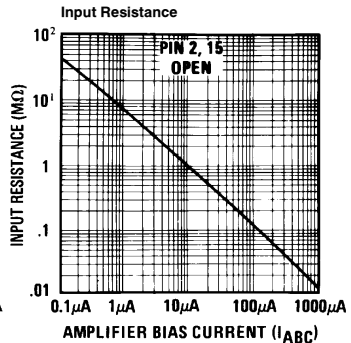
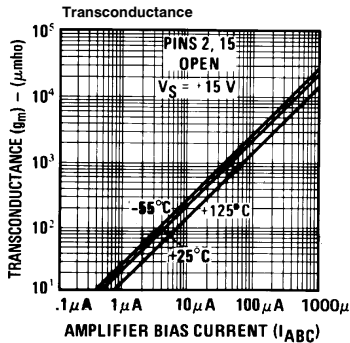
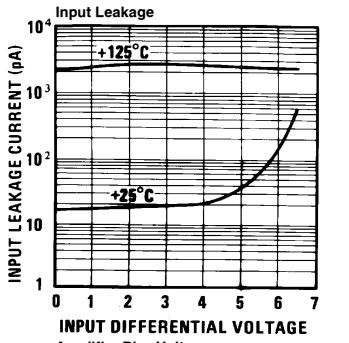
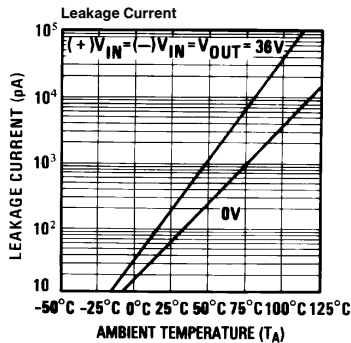
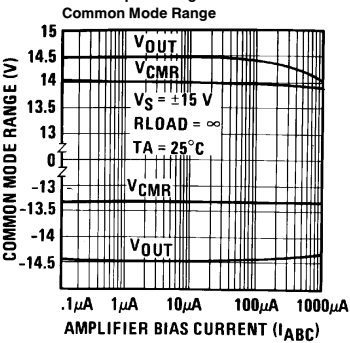
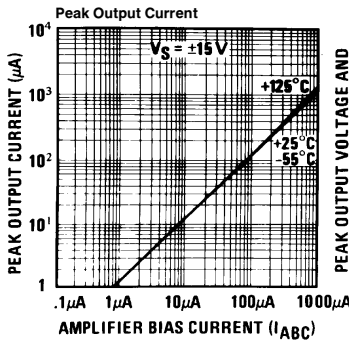
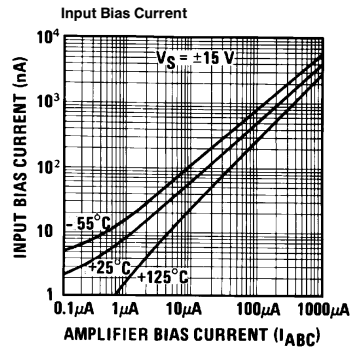
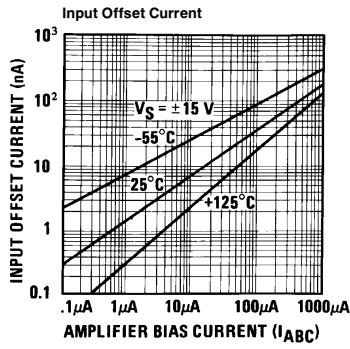
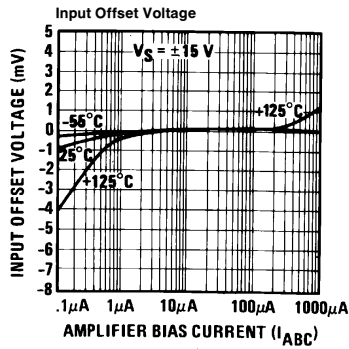
**Note 5:** These specifications apply for  $V_S = \pm 15\text{V}$ ,  $I_{ABC} = 500 \mu\text{A}$ ,  $R_{OUT} = 5 \text{ k}\Omega$  connected from the buffer output to  $-V_S$  and the input of the buffer is connected to the transconductance amplifier output.

## Schematic Diagram



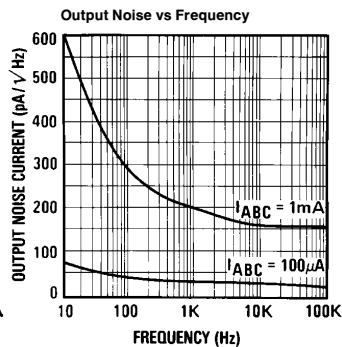
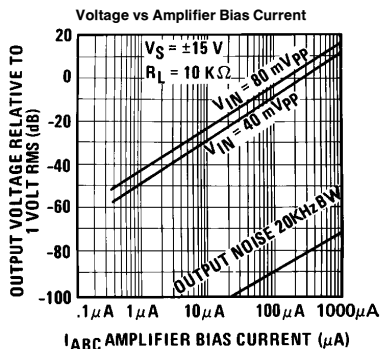
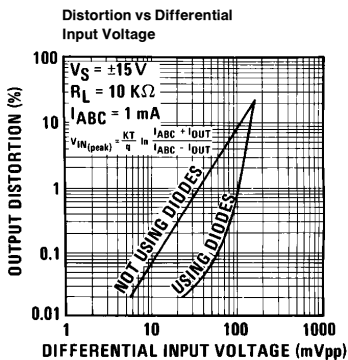
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# Typical Performance Characteristics



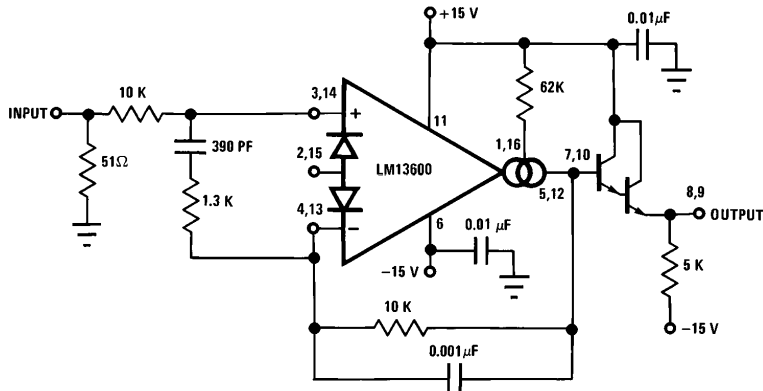
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# Typical Performance Characteristics (Continued)



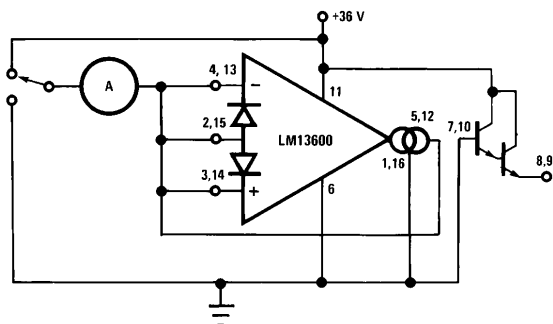
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**Unity Gain Follower**

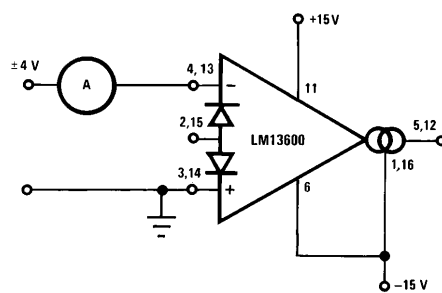


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**Leakage Current Test Circuit**



**Differential Input Current Test Circuit**



TL/H/7980-7

TL/H/7980-6

## Circuit Description

The differential transistor pair  $Q_4$  and  $Q_5$  form a transconductance stage in that the ratio of their collector currents is defined by the differential input voltage according to the transfer function:

$$V_{IN} = \frac{kT}{q} \ln \frac{I_5}{I_4} \quad (1)$$

where  $V_{IN}$  is the differential input voltage,  $kT/q$  is approximately 26 mV at 25°C and  $I_5$  and  $I_4$  are the collector currents of transistors  $Q_5$  and  $Q_4$  respectively. With the exception of  $Q_3$  and  $Q_{13}$ , all transistors and diodes are identical in size. Transistors  $Q_1$  and  $Q_2$  with Diode  $D_1$  form a current mirror which forces the sum of currents  $I_4$  and  $I_5$  to equal  $I_{ABC}$ :

$$I_4 + I_5 = I_{ABC} \quad (2)$$

where  $I_{ABC}$  is the amplifier bias current applied to the gain pin.

For small differential input voltages the ratio of  $I_4$  and  $I_5$  approaches unity and the Taylor series of the ln function can be approximated as:

$$\frac{kT}{q} \ln \frac{I_5}{I_4} \approx \frac{kT}{q} \frac{I_5 - I_4}{I_4} \quad (3)$$

$$I_4 \approx I_5 \approx \frac{I_{ABC}}{2}$$

$$V_{IN} \left[ \frac{I_{ABC} q}{2kT} \right] = I_5 - I_4 \quad (4)$$

Collector currents  $I_4$  and  $I_5$  are not very useful by themselves and it is necessary to subtract one current from the

other. The remaining transistors and diodes form three current mirrors that produce an output current equal to  $I_5$  minus  $I_4$  thus:

$$V_{IN} \left[ \frac{I_{ABC} q}{2kT} \right] = I_{OUT} \quad (5)$$

The term in brackets is then the transconductance of the amplifier and is proportional to  $I_{ABC}$ .

## Linearizing Diodes

For differential voltages greater than a few millivolts, Equation 3 becomes less valid and the transconductance becomes increasingly nonlinear. *Figure 1* demonstrates how the internal diodes can linearize the transfer function of the amplifier. For convenience assume the diodes are biased with current sources and the input signal is in the form of current  $I_S$ . Since the sum of  $I_4$  and  $I_5$  is  $I_{ABC}$  and the difference is  $I_{OUT}$ , currents  $I_4$  and  $I_5$  can be written as follows:

$$I_4 = \frac{I_{ABC}}{2} - \frac{I_{OUT}}{2}, I_5 = \frac{I_{ABC}}{2} + \frac{I_{OUT}}{2}$$

Since the diodes and the input transistors have identical geometries and are subject to similar voltages and temperatures, the following is true:

$$\begin{aligned} \frac{kT}{q} \ln \frac{\frac{I_D}{2} + I_S}{\frac{I_D}{2} - I_S} &= \frac{kT}{q} \ln \frac{\frac{I_{ABC}}{2} + \frac{I_{OUT}}{2}}{\frac{I_{ABC}}{2} - \frac{I_{OUT}}{2}} \\ \therefore I_{OUT} &= I_S \left( \frac{2I_{ABC}}{I_D} \right) \quad \text{for } |I_S| < \frac{I_D}{2} \end{aligned} \quad (6)$$

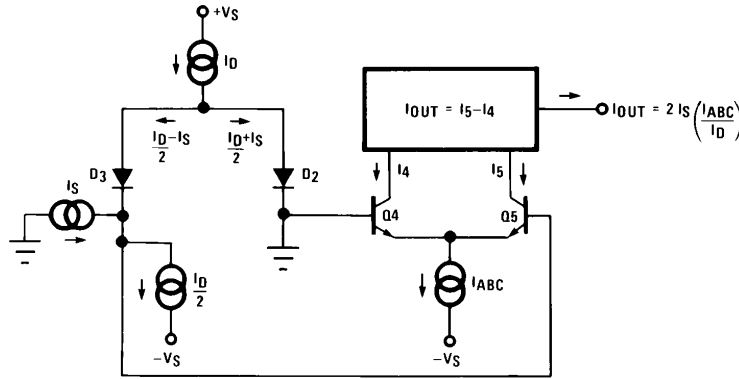


FIGURE 1. Linearizing Diodes

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## Linearizing Diodes (Continued)

Notice that in deriving Equation 6 no approximations have been made and there are no temperature-dependent terms. The limitations are that the signal current not exceed  $I_D/2$  and that the diodes be biased with currents. In practice, replacing the current sources with resistors will generate insignificant errors.

## Controlled Impedance Buffers

The upper limit of transconductance is defined by the maximum value of  $I_{ABC}$  (2 mA). The lowest value of  $I_{ABC}$  for which the amplifier will function therefore determines the overall dynamic range. At very low values of  $I_{ABC}$ , a buffer which has very low input bias current is desirable. An FET follower satisfies the low input current requirement, but is somewhat non-linear for large voltage swing. The controlled impedance buffer is a Darlington which modifies its input bias current to suit the need. For low values of  $I_{ABC}$ , the buffer's input current is minimal. At higher levels of  $I_{ABC}$ , transistor  $Q_3$  biases up  $Q_{12}$  with a current proportional to  $I_{ABC}$  for fast slew rate. When  $I_{ABC}$  is changed, the DC level of the Darlington output buffer will shift. In audio applications where  $I_{ABC}$  is changed suddenly, this shift may produce an audible "pop". For these applications the LM13700 may produce superior results.

## Applications—Voltage Controlled Amplifiers

Figure 2 shows how the linearizing diodes can be used in a voltage-controlled amplifier. To understand the input biasing, it is best to consider the 13 k $\Omega$  resistor as a current source and use a Thevenin equivalent circuit as shown in Figure 3. This circuit is similar to Figure 1 and operates the same. The potentiometer in Figure 2 is adjusted to minimize the effects of the control signal at the output.

For optimum signal-to-noise performance,  $I_{ABC}$  should be as large as possible as shown by the Output Voltage vs. Amplifier Bias Current graph. Larger amplitudes of input signal also improve the S/N ratio. The linearizing diodes help here by allowing larger input signals for the same output distortion as shown by the Distortion vs. Differential Input Voltage graph. S/N may be optimized by adjusting the magnitude of the input signal via  $R_{IN}$  (Figure 2) until the output distortion is below some desired level. The output voltage swing can then be set at any level by selecting  $R_L$ .

Although the noise contribution of the linearizing diodes is negligible relative to the contribution of the amplifier's internal transistors,  $I_D$  should be as large as possible. This minimizes the dynamic junction resistance of the diodes ( $r_d$ ) and maximizes their linearizing action when balanced against  $R_{IN}$ . A value of 1 mA is recommended for  $I_D$  unless the specific application demands otherwise.

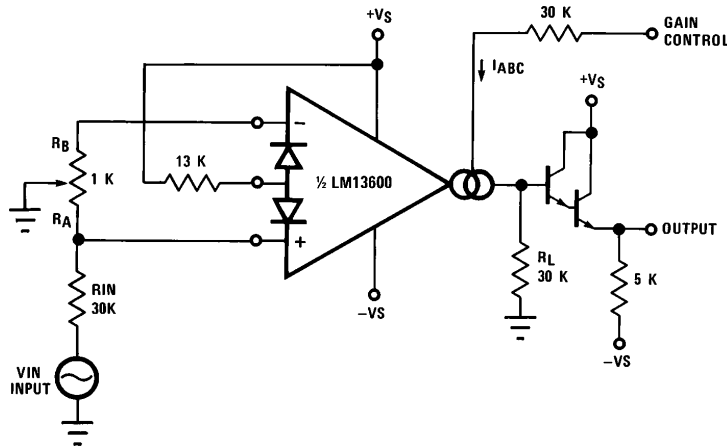


FIGURE 2. Voltage Controlled Amplifier

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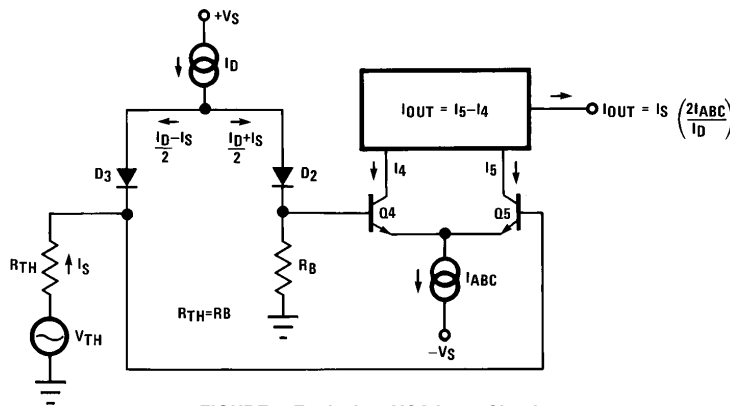


FIGURE 3. Equivalent VCA Input Circuit

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## Stereo Volume Control

The circuit of Figure 4 uses the excellent matching of the two LM13600 amplifiers to provide a Stereo Volume Control with a typical channel-to-channel gain tracking of 0.3 dB.  $R_p$  is provided to minimize the output offset voltage and may be replaced with two 510  $\Omega$  resistors in AC-coupled applications. For the component values given, amplifier gain is derived for Figure 2 as being:

$$\frac{V_O}{V_{IN}} = 940 \times I_{ABC}$$

If  $V_C$  is derived from a second signal source then the circuit becomes an amplitude modulator or two-quadrant multiplier as shown in Figure 5, where:

$$I_O = \frac{-2I_S}{I_D} (I_{ABC}) = \frac{-2I_S}{I_D} \frac{V_{IN2}}{R_C} - \frac{2I_S}{I_D} \frac{(V^- + 1.4V)}{R_C}$$

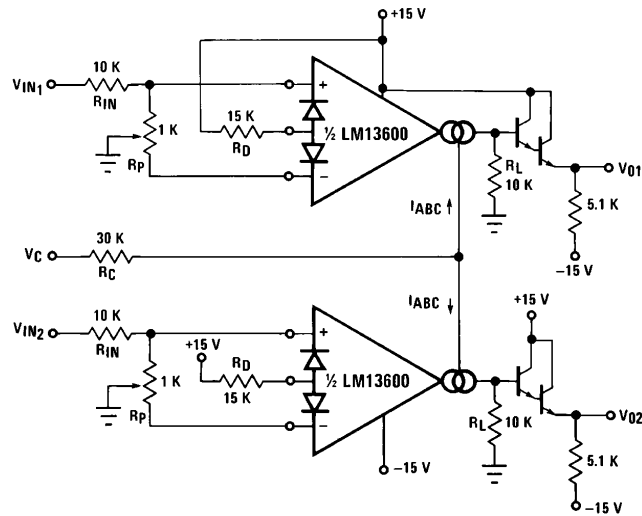


FIGURE 4. Stereo Volume Control

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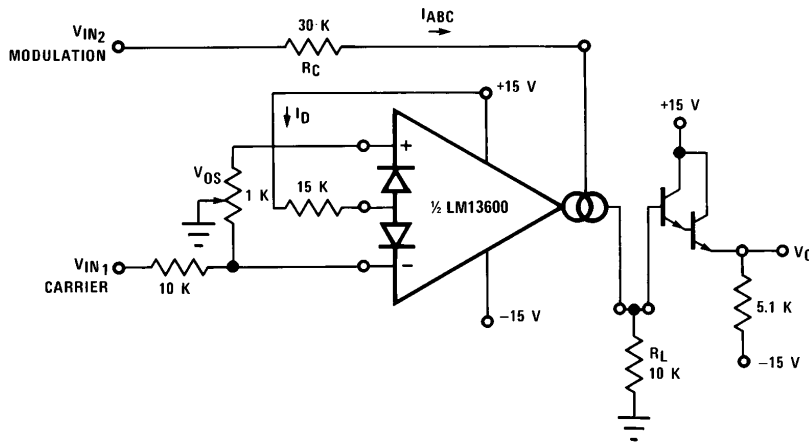


FIGURE 5. Amplitude Modulator

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## Stereo Volume Control (Continued)

The constant term in the above equation may be cancelled by feeding  $I_S \times I_D R_C / 2 (V^- + 1.4V)$  into  $I_O$ . The circuit of *Figure 6* adds  $R_M$  to provide this current, resulting in a four-quadrant multiplier where  $R_C$  is trimmed such that  $V_O = 0V$  for  $V_{IN2} = 0V$ .  $R_M$  also serves as the load resistor for  $I_O$ .

Noting that the gain of the LM13600 amplifier of *Figure 3* may be controlled by varying the linearizing diode current  $I_D$  as well as by varying  $I_{ABC}$ , *Figure 7* shows an AGC Amplifier using this approach. As  $V_O$  reaches a high enough amplitude (3  $V_{BE}$ ) to turn on the Darlington transistors and the linearizing diodes, the increase in  $I_D$  reduces the amplifier gain so as to hold  $V_O$  at that level.

## Voltage Controlled Resistors

An Operational Transconductance Amplifier (OTA) may be used to implement a Voltage Controlled Resistor as shown

in *Figure 8*. A signal voltage applied at  $R_X$  generates a  $V_{IN}$  to the LM13600 which is then multiplied by the  $g_m$  of the amplifier to produce an output current, thus:

$$R_X = \frac{R + R_A}{g_m R_A}$$

where  $g_m \approx 19.2 I_{ABC}$  at 25°C. Note that the attenuation of  $V_O$  by  $R$  and  $R_A$  is necessary to maintain  $V_{IN}$  within the linear range of the LM13600 input.

*Figure 9* shows a similar VCR where the linearizing diodes are added, essentially improving the noise performance of the resistor. A floating VCR is shown in *Figure 10*, where each "end" of the "resistor" may be at any voltage within the output voltage range of the LM13600.

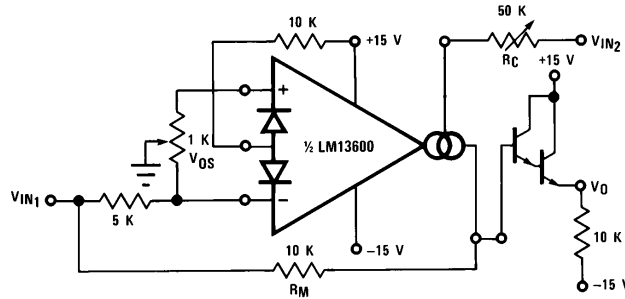


FIGURE 6. Four-Quadrant Multiplier

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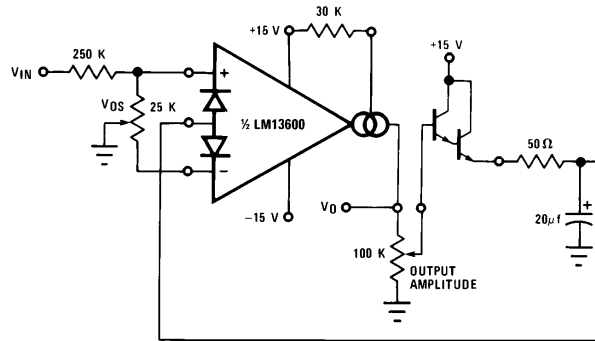


FIGURE 7. AGC Amplifier

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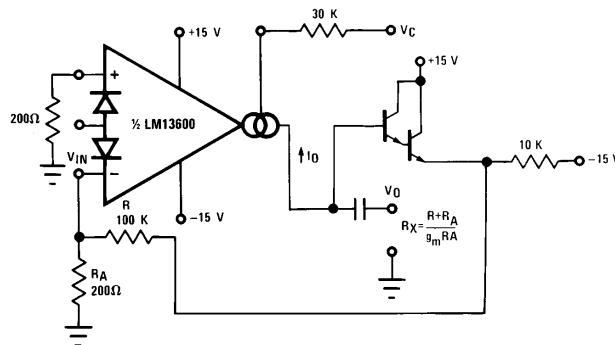


FIGURE 8. Voltage Controlled Resistor, Single-Ended

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## Voltage Controlled Filters

OTA's are extremely useful for implementing voltage controlled filters, with the LM13600 having the advantage that the required buffers are included on the I.C. The VC Lo-Pass Filter of Figure 11 performs as a unity-gain buffer amplifier at frequencies below cut-off, with the cut-off frequency being the point at which  $X_C/g_m$  equals the closed-loop gain of  $(R/R_A)$ . At frequencies above cut-off the circuit provides a single RC roll-off (6 dB per octave) of the input signal amplitude with a  $-3$  dB point defined by the given equation,

where  $g_m$  is again  $19.2 \times I_{ABC}$  at room temperature. Figure 12 shows a VC High-Pass Filter which operates in much the same manner, providing a single RC roll-off below the defined cut-off frequency.

Additional amplifiers may be used to implement higher order filters as demonstrated by the two-pole Butterworth Lo-Pass Filter of Figure 13 and the state variable filter of Figure 14. Due to the excellent  $g_m$  tracking of the two amplifiers and the varied bias of the buffer Darlington's, these filters perform well over several decades of frequency.

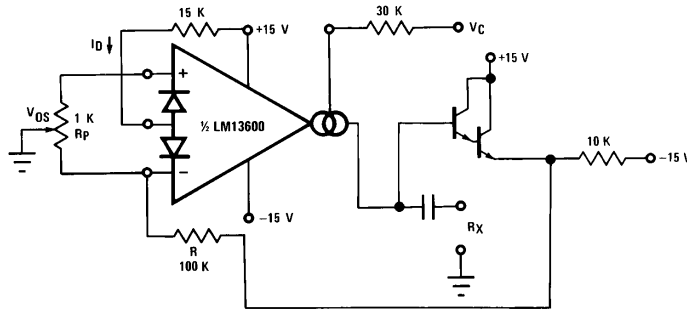


FIGURE 9. Voltage Controlled Resistor with Linearizing Diodes

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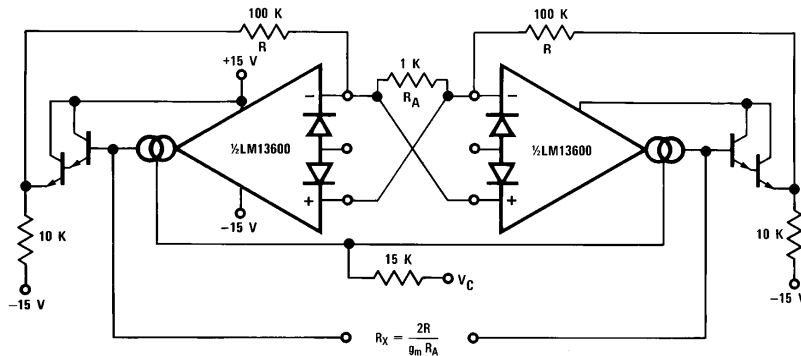


FIGURE 10. Floating Voltage Controlled Resistor

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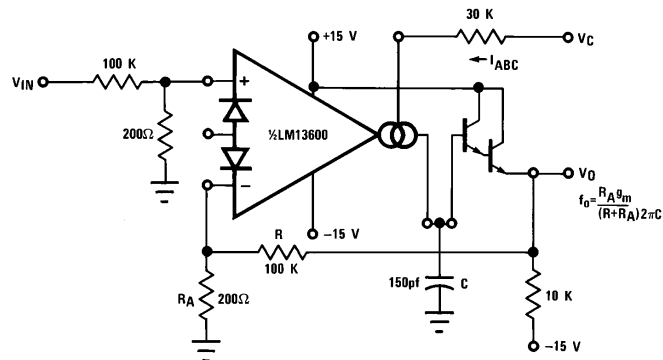
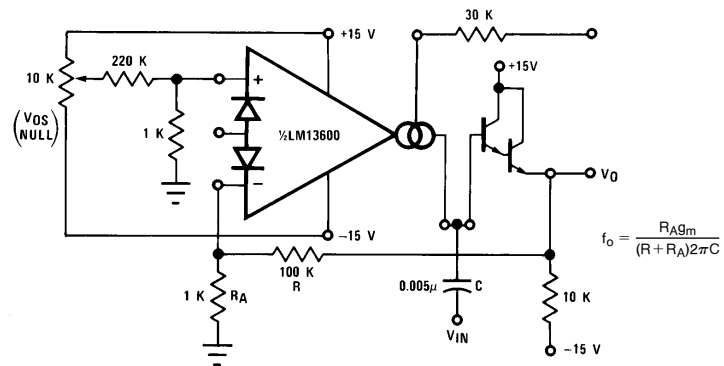


FIGURE 11. Voltage Controlled Low-Pass Filter

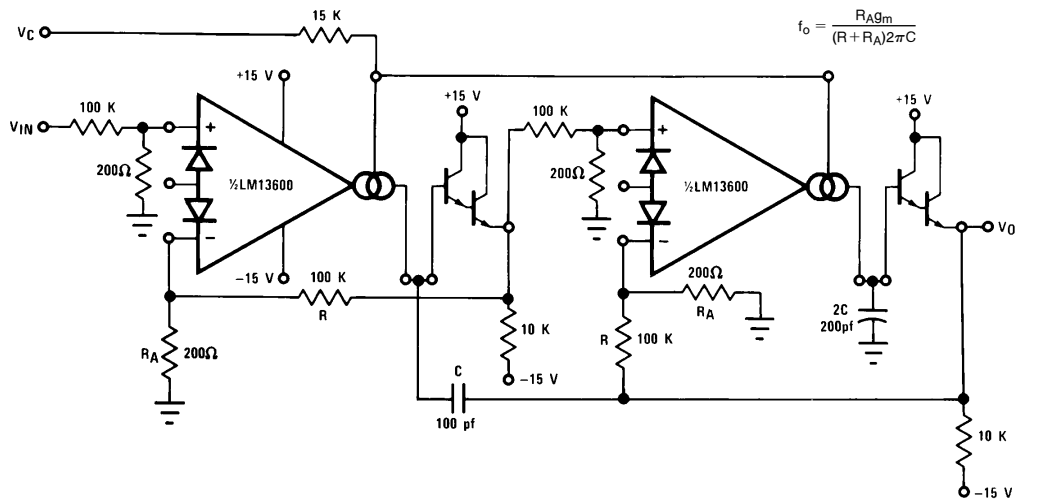
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## Voltage Controlled Filters (Continued)



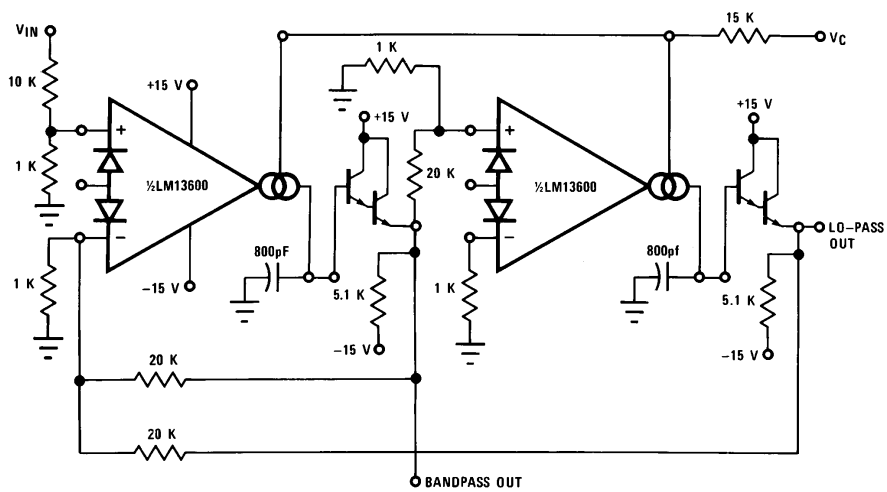
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FIGURE 12. Voltage Controlled Hi-Pass Filter



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FIGURE 13. Voltage Controlled 2-Pole Butterworth Lo-Pass Filter



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FIGURE 14. Voltage Controlled State Variable Filter

## Voltage Controlled Oscillators

The classic Triangular/Square Wave VCO of Figure 15 is one of a variety of Voltage Controlled Oscillators which may be built utilizing the LM13600. With the component values shown, this oscillator provides signals from 200 kHz to below 2 Hz as  $I_C$  is varied from 1 mA to 10 nA. The output amplitudes are set by  $I_A \times R_A$ . Note that the peak differential input voltage must be less than 5V to prevent zenering the inputs.

A few modifications to this circuit produce the ramp/pulse VCO of Figure 16. When  $V_{O2}$  is high,  $I_F$  is added to  $I_C$  to

increase amplifier A1's bias current and thus to increase the charging rate of capacitor C. When  $V_{O2}$  is low,  $I_F$  goes to zero and the capacitor discharge current is set by  $I_C$ .

The VC Lo-Pass Filter of Figure 11 may be used to produce a high-quality sinusoidal VCO. The circuit of Figure 16 employs two LM13600 packages, with three of the amplifiers configured as lo-pass filters and the fourth as a limiter/inverter. The circuit oscillates at the frequency at which the loop phase-shift is  $360^\circ$  or  $180^\circ$  for the inverter and  $60^\circ$  per filter stage. This VCO operates from 5 Hz to 50 kHz with less than 1% THD.

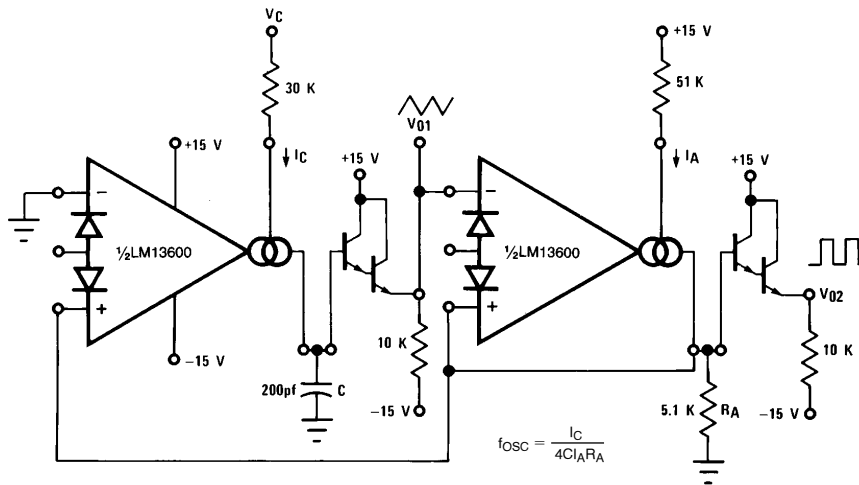


FIGURE 15. Triangular/Square-Wave VCO

TL/H/7980-22

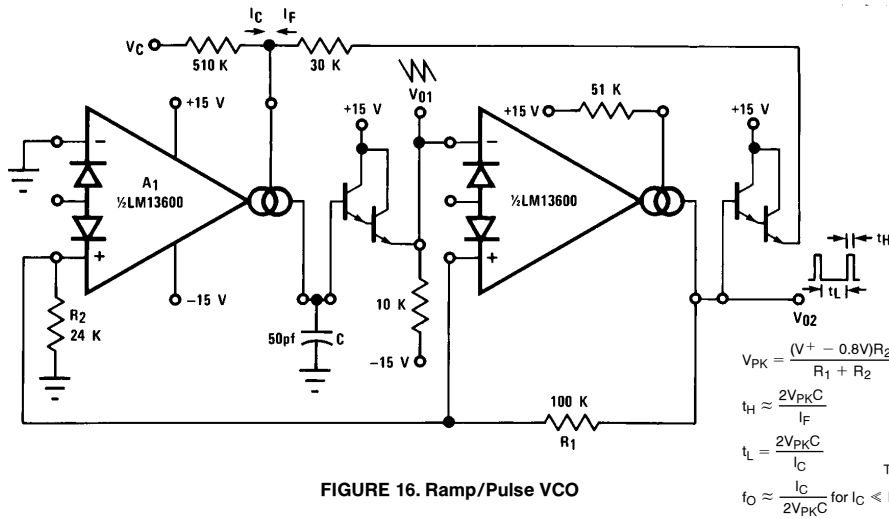


FIGURE 16. Ramp/Pulse VCO

TL/H/7980-23

## Voltage Controlled Oscillators (Continued)

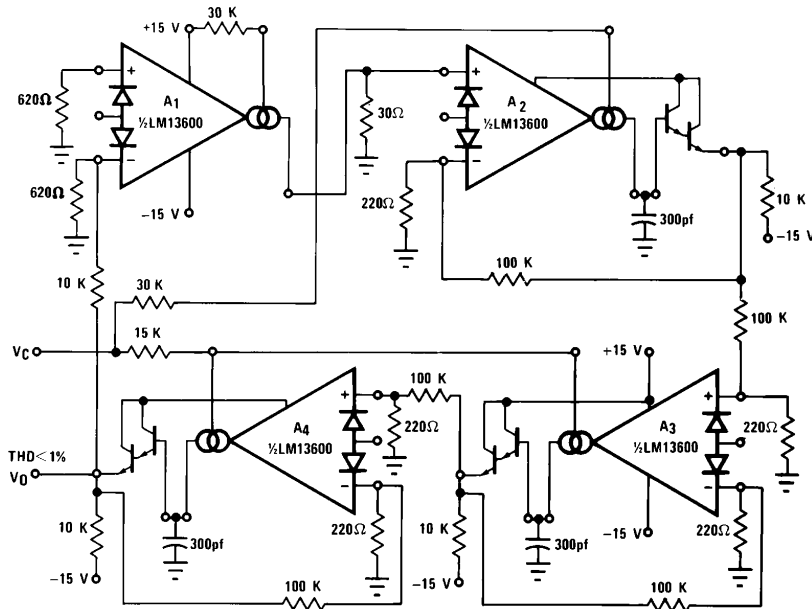


FIGURE 17. Sinusoidal VCO

TL/H/7980-24

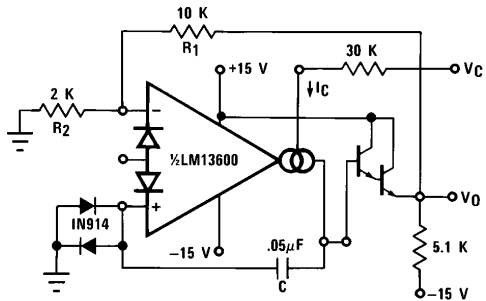


FIGURE 18. Single Amplifier VCO

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Figure 18 shows how to build a VCO using one amplifier when the other amplifier is needed for another function.

## Additional Applications

Figure 19 presents an interesting one-shot which draws no power supply current until it is triggered. A positive-going trigger pulse of at least 2V amplitude turns on the amplifier through R<sub>B</sub> and pulls the non-inverting input high. The amplifier regenerates and latches its output high until capacitor C charges to the voltage level on the non-inverting input. The output then switches low, turning off the amplifier and discharging the capacitor. The capacitor discharge rate is increased by shorting the diode bias pin to the inverting input so that an additional discharge current flows through D<sub>1</sub> when the amplifier output switches low. A special feature of this timer is that the other amplifier, when biased from V<sub>O</sub>, can perform another function and draw zero stand-by power as well.

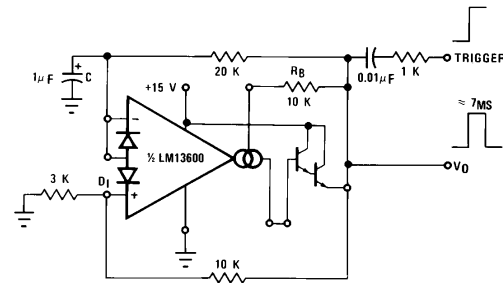


FIGURE 19. Zero Stand-By Power Timer

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## Additional Applications (Continued)

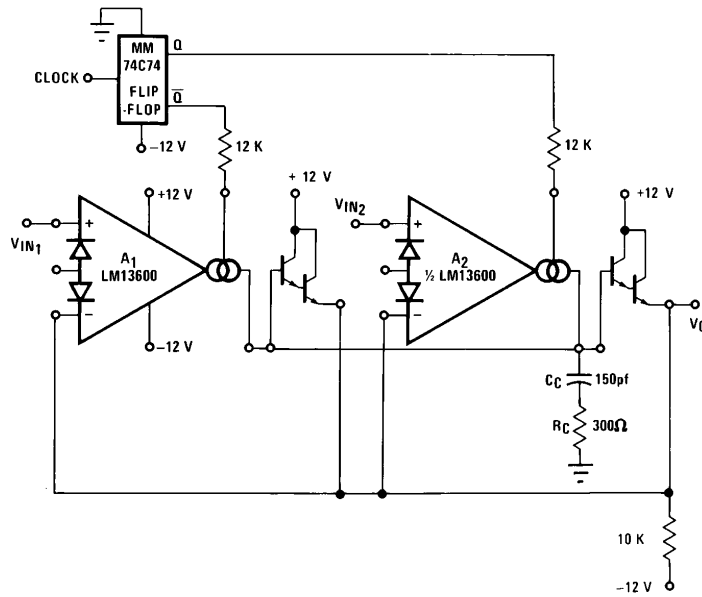


FIGURE 20. Multiplexer

TL/H/7980-27

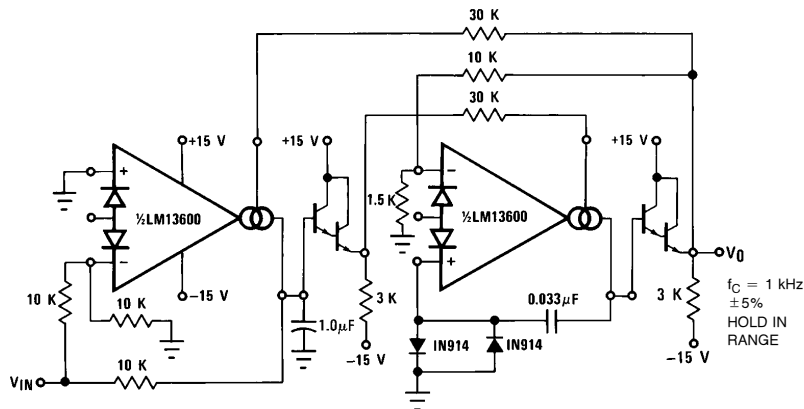


FIGURE 21. Phase Lock Loop

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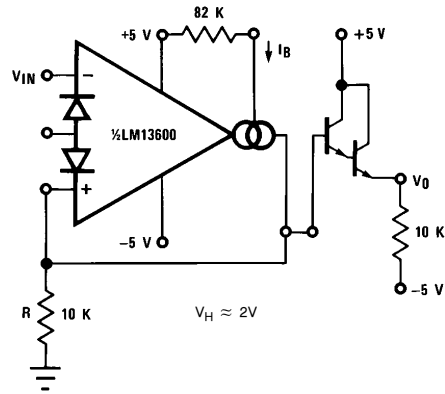
The Schmitt Trigger of Figure 22 uses the amplifier output current into R to set the hysteresis of the comparator; thus  $V_H = 2 \times R \times I_B$ . Varying  $I_B$  will produce a Schmitt Trigger with variable hysteresis.

Figure 23 shows a Tachometer or Frequency-to-Voltage converter. Whenever A1 is toggled by a positive-going input, an amount of charge equal to  $(V_H - V_L) C_f$  is sourced into  $C_f$  and  $R_f$ . This once-per-cycle charge is then balanced by the current of  $V_O/R_f$ . The maximum  $f_{IN}$  is limited by the amount of time required to charge  $C_f$  from  $V_L$  to  $V_H$  with a current of  $I_B$ , where  $V_L$  and  $V_H$  represent the maximum low and maxi-

imum high output voltage swing of the LM13600. D1 is added to provide a discharge path for  $C_f$  when A1 switches low.

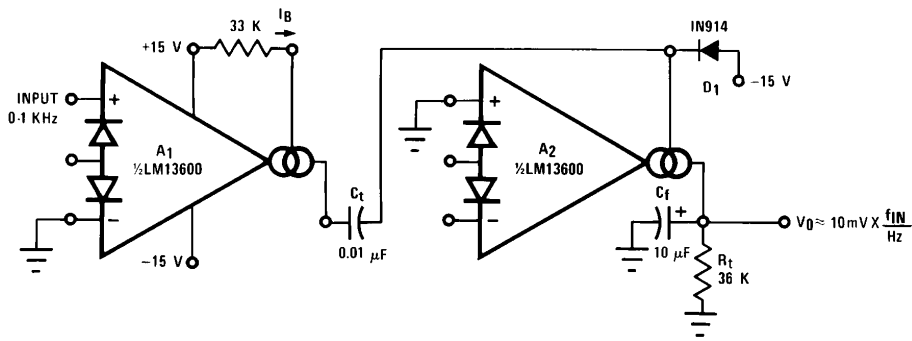
The Peak Detector of Figure 24 uses A2 to turn on A1 whenever  $V_{IN}$  becomes more positive than  $V_O$ . A1 then charges storage capacitor C to hold  $V_O$  equal to  $V_{INPK}$ . One precaution to observe when using this circuit: the Darlington transistor used must be on the same side of the package as A2 since the A1 Darlington will be turned on and off with A1. Pulling the output of A2 low through D1 serves to turn off A1 so that  $V_O$  remains constant.

**Additional Applications** (Continued)



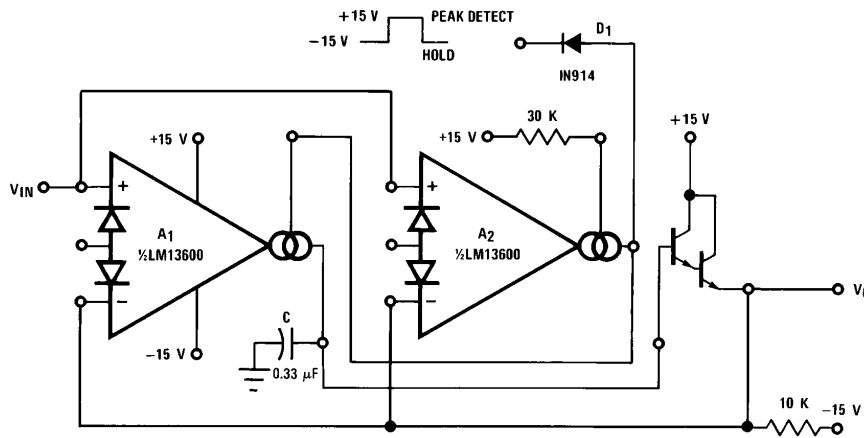
**FIGURE 22. Schmitt Trigger**

TL/H/7980-29



**FIGURE 23. Tachometer**

TL/H/7980-30



**FIGURE 24. Peak Detector and Hold Circuit**

TL/H/7980-31

## Additional Applications (Continued)

The Sample-Hold circuit of Figure 25 also requires that the Darlington buffer used be from the other (A2) half of the package and that the corresponding amplifier be biased on continuously. The Ramp-and-Hold of Figure 26 sources  $I_B$  into capacitor C whenever the input to A1 is brought high, giving a ramp-rate of about 1 V/ms for the component values shown.

The true-RMS converter of Figure 27 is essentially an automatic gain control amplifier which adjusts its gain such that the AC power at the output of amplifier A1 is constant. The output power of amplifier A1 is monitored by squaring amplifier A2 and the average compared to a reference voltage with amplifier A3. The output of A3 provides bias current to the diodes of A1 to attenuate the input signal. Because the output power of A1 is held constant, the RMS value is constant and the attenuation is directly proportional to the RMS value of the input voltage. The attenuation is also proportional to the diode bias current. Amplifier A4 adjusts the ratio of currents through the diodes to be equal and therefore the voltage at the output of A4 is proportional to the RMS value of the input voltage. The calibration potentiometer is set such that  $V_O$  reads directly in RMS volts.

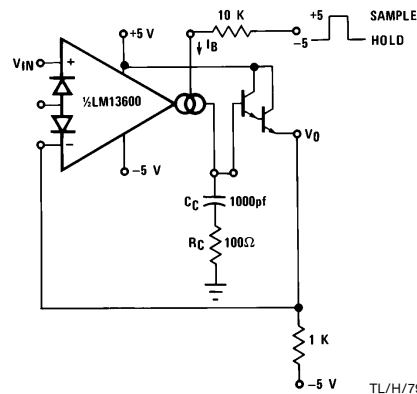


FIGURE 25. Sample-Hold Circuit

TL/H/7980-32

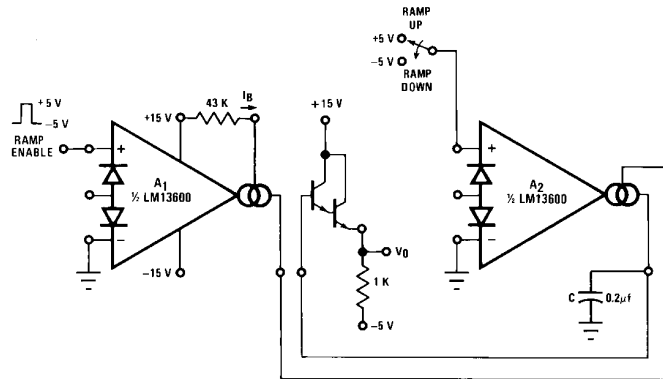


FIGURE 26. Ramp and Hold

TL/H/7980-33

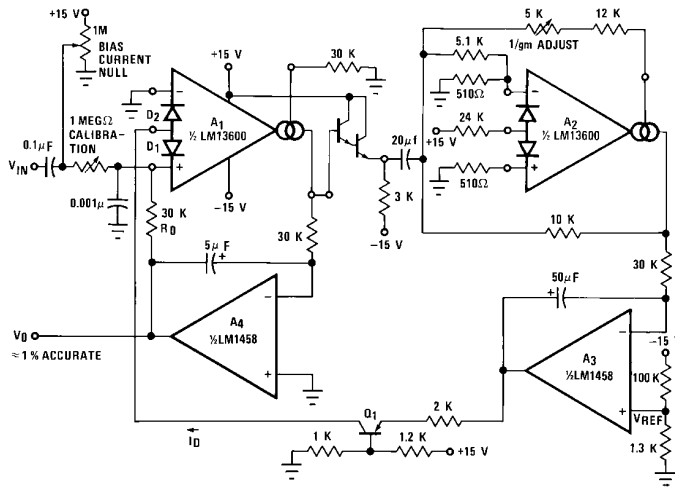


FIGURE 27. True RMS Converter

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## Additional Applications (Continued)

The circuit of *Figure 28* is a voltage reference of variable temperature coefficient. The 100 kΩ potentiometer adjusts the output voltage which has a positive TC above 1.2V, zero TC at about 1.2V and negative TC below 1.2V. This is accomplished by balancing the TC of the A2 transfer function against the complementary TC of D1.

The log amplifier of *Figure 29* responds to the ratio of currents through buffer transistors Q3 and Q4. Zero temperature dependence for  $V_{OUT}$  is ensured because the TC of the A2 transfer function is equal and opposite to the TC of the logging transistors Q3 and Q4.

The wide dynamic range of the LM13600 allows easy control of the output pulse width in the Pulse Width Modulator of *Figure 30*.

For generating  $I_{ABC}$  over a range of 4 to 6 decades of current, the system of *Figure 31* provides a logarithmic current out for a linear voltage in.

Since the closed-loop configuration ensures that the input to A2 is held equal to 0V, the output current of A1 is equal to  $I_3 = -V_C/R_C$ .

The differential voltage between Q1 and Q2 is attenuated by the R1, R2 network so that A1 may be assumed to be

operating within its linear range. From equation (5), the input voltage to A1 is:

$$V_{IN1} = \frac{-2kT I_3}{q I_2} = \frac{2kT V_C}{q I_2 R_C}$$

The voltage on the base of Q1 is then

$$V_{B1} = \frac{(R_1 + R_2) V_{IN1}}{R_1}$$

The ratio of the Q1 and Q2 collector currents is defined by:

$$V_{B1} = \frac{kT}{q} \ln \frac{I_{C2}}{I_{C1}} \approx \frac{kT}{q} \ln \frac{I_{ABC}}{I_1}$$

Combining and solving for  $I_{ABC}$  yields:

$$I_{ABC} = I_1 \exp \left[ \frac{2(R_1 + R_2) V_C}{R_1 I_2 R_C} \right]$$

This logarithmic current can be used to bias the circuit of *Figure 4* provide a temperature independent stereo attenuation characteristic.

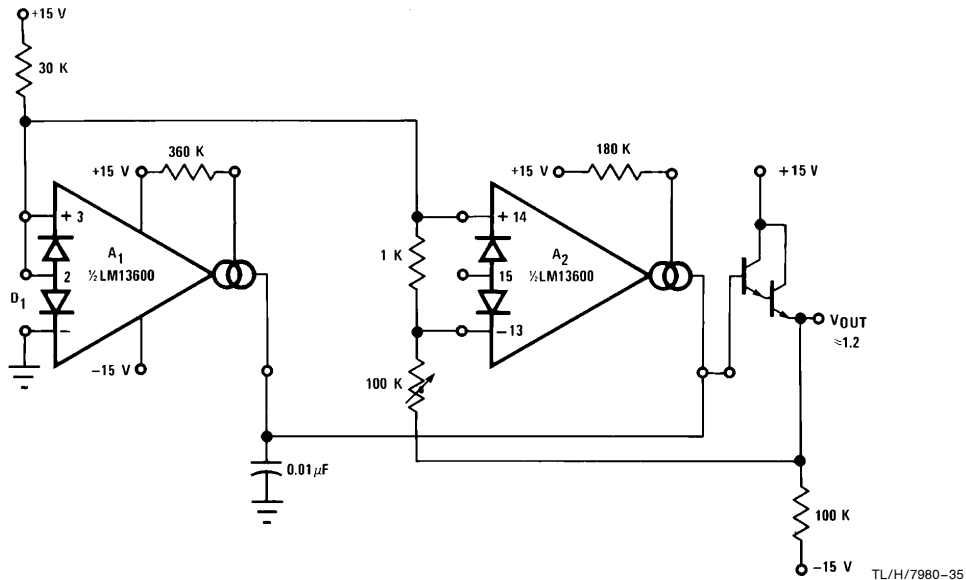


FIGURE 28. Delta VBE Reference

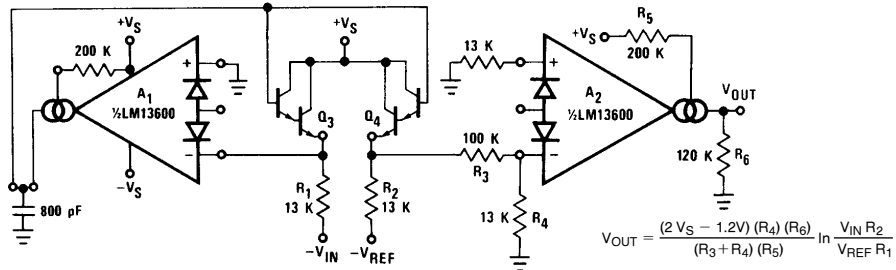


FIGURE 29. Log Amplifier

$$V_{OUT} = \frac{(2V_S - 1.2V)(R_4)(R_6)}{(R_3 + R_4)(R_5)} \ln \frac{V_{IN} R_2}{V_{REF} R_1}$$

TL/H/7980-36

## Additional Applications (Continued)

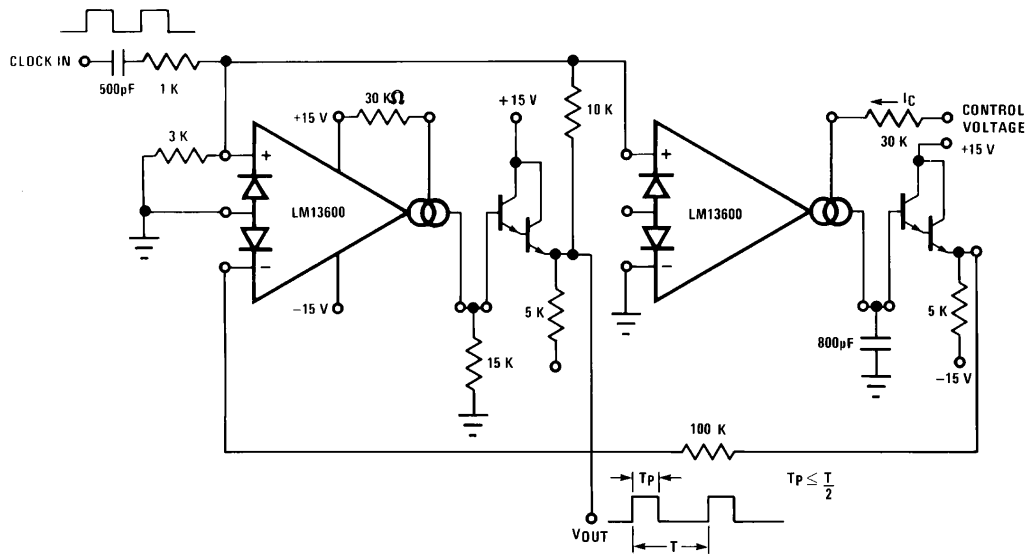


FIGURE 30. Pulse Width Modulator

TL/H/7980-37

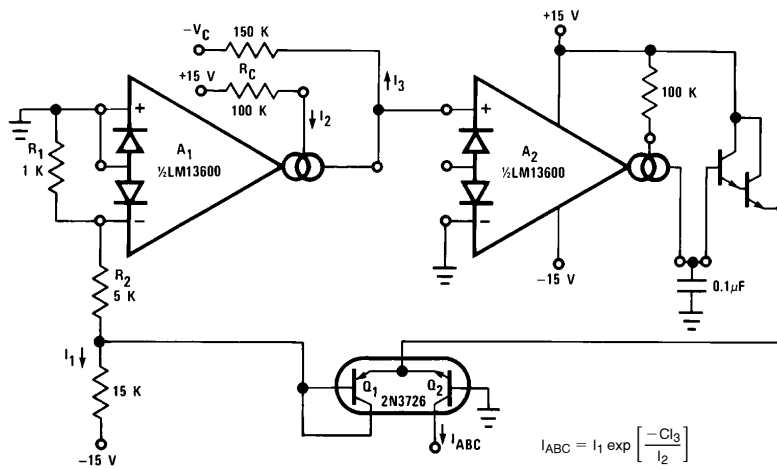
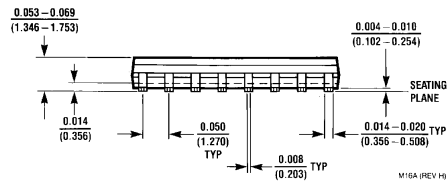
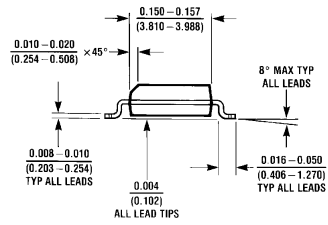
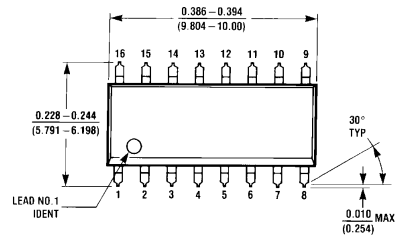


FIGURE 31. Logarithmic Current Source

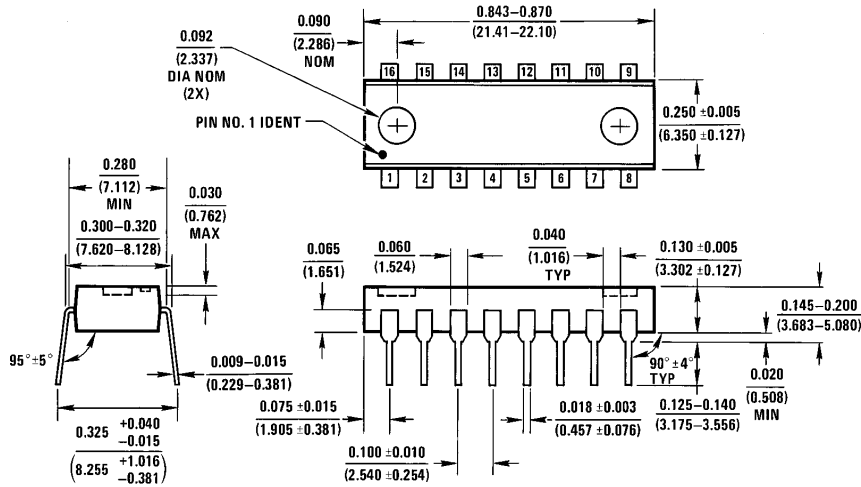
TL/H/7980-38

**Physical Dimensions** inches (millimeters)



**S.O. Package (M)**  
**Order Number LM13600M**  
**NS Package Number M16A**

**Physical Dimensions** inches (millimeters) (Continued)



**Molded Dual-In-Line Package (N)**  
**Order Number LM13600N or LM13600AN**  
**NS Package Number N16A**

N16A (REV E)

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