

Overview

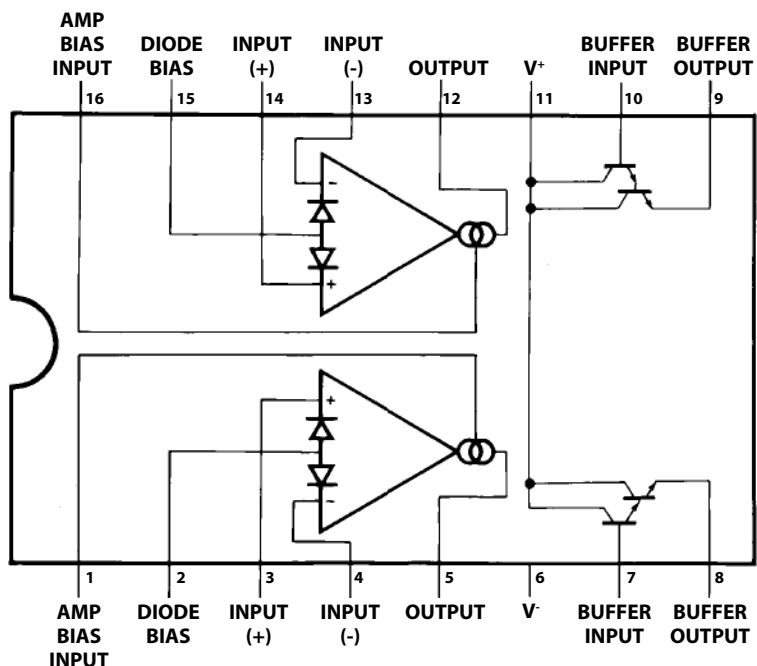
The V13700M/D series consists of two current controlled transconductance amplifiers, each with differential inputs and a push-pull output. The two amplifiers share common supplies but otherwise operate independently. Linearizing diodes are provided at the inputs to reduce distortion and allow higher input levels. The result is a 10 dB signal-to-noise improvement referenced to 0.5 percent THD. High impedance buffers are provided which are especially designed to complement the dynamic range of the amplifiers. The output buffers of the V13700M/D differ from those of the LM13600 in that their input bias currents (and hence their output DC levels) are independent of IABC. This may result in performance superior to that of the LM13600 in audio applications.

Its **features** are:

- gm adjustable over 6 decades
- Excellent gm linearity
- Excellent matching between amplifiers
- Linearizing diodes
- High impedance buffers
- High output signal-to-noise ratio

Block Diagram and Pin Description

Dual-In-Line and Small Outline Packages



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Electrical Characteristics**Absolute Maximum Ratings (Note 1)**

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

Supply Voltage V13700M/D	36 V _{DC} or ± 18 V
Differential Input	Voltage ± 5 V
Power Dissipation (Note 2) TA = 25°C V13700M/D	570 mW
Amplifier Bias Current (IABC)	2 mA
Output Short Circuit Duration	Continuous
Diode Bias Current (ID)	2 mA
Buffer Output Current (Note 3)	20 mA
Operating Temperature Range V13700M/D	0 °C to +70 °C
DC Input Voltage	+VS to -VS
Storage Temperature Range	-65 °C to +150 °C
Soldering Information	Soldering (10 sec.) 260 °C
Dual-In-Line Package	Vapor Phase (60 sec.) 215 °C
Small Outline Package	Infrared (15 sec.) 220 °C

Electrical Characteristics (Note 4)

Parameter	Conditions	V13700M/D			Units
		Min	Typ	Max	
Input Offset Voltage	Over Specified Temperature Range IABC = 5 μ A		0.4 0.3	4 4	mV
VOS Including Diodes	Diode Bias Current (ID) = 500 μ A		0.5	5	mV
Input Offset Change	5 μ A~ IABC ~ 500 μ A		0.1	3	mV
Input Offset Current			0.1	0.6	μ A
Input Bias Current	Over Specified Temperature Range		0.4 1	5 8	μ A
Forward Transconductance (g_m)		6700	9600	13000	μ mho
	Over Specified Temperature Range	5400			
g_m Tracking			0.3		dB
Peak Output Current	RL = 0, IABC = 5 μ A		5		μ A
	RL = 0, IABC = 500 μ A	350	500	650	
	RL = 0, Over Specified Temp Range	300			
Peak Output Voltage					V
Positive	RL = ∞ , 5 μ A $\leq$ IABC $\leq$ 500 μ A	12	14.2		
Negative	RL = ∞ , 5 μ A $\leq$ IABC $\leq$ 500 μ A	-12	-14.4		
Supply Current	IABC = 500 μ A, Both Channels		2.6		mA
VOS Sensitivity					μ V/V
Positive	~VOS/~V+		20	150	
Negative	~VOS/~V		20	150	
CMRR		80	110		dB
Common Mode Range		± 12	± 13.5		V
Crosstalk	Referred to Input (Note 5) 20 Hz < f < 20 kHz		100		dB
Differential Input Current	IABC = 0, Input = ± 4 V		0.02	100	nA
Leakage Current	IABC = 0 (Refer to Test Circuit)		0.2	100	nA
Input Resistance		10	26		k $\sim$
Open Loop Bandwidth			2		MHz
Slew Rate	Unity Gain Compensated		50		V/ μ s
Buffer Input Current	(Note 5)		0.5	2	μ A
Peak Buffer Output Voltage	(Note 5)	10			V

Note 1: "Absolute Maximum Ratings" indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is functional, but do not guarantee specific performance limits.

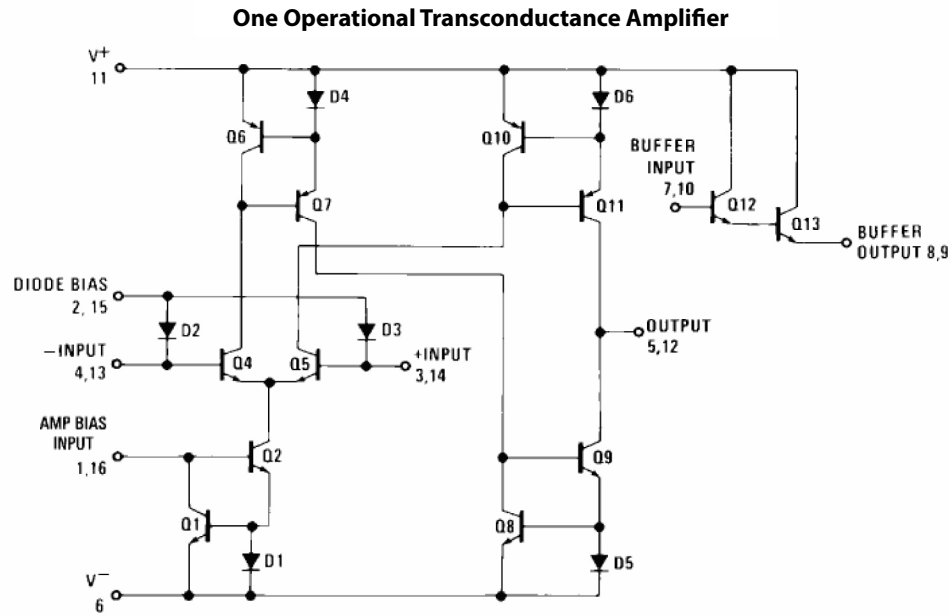
Note 2: For operation at ambient temperatures above 25 $^{\circ}$ C, the device must be derated based on a 150 $^{\circ}$ C maximum junction temperature and a thermal resistance, junction to ambient, as follows: V13700M/D, 90 $^{\circ}$ C/W; V13700M/D, 110 $^{\circ}$ C/W.

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

Note 4: These specifications apply for VS = ± 15 V, TA = 25 $^{\circ}$ C, amplifier bias current (IABC) = 500 μ A, pins 2 and 15 open unless otherwise specified. The inputs to the buffers are grounded and outputs are open.

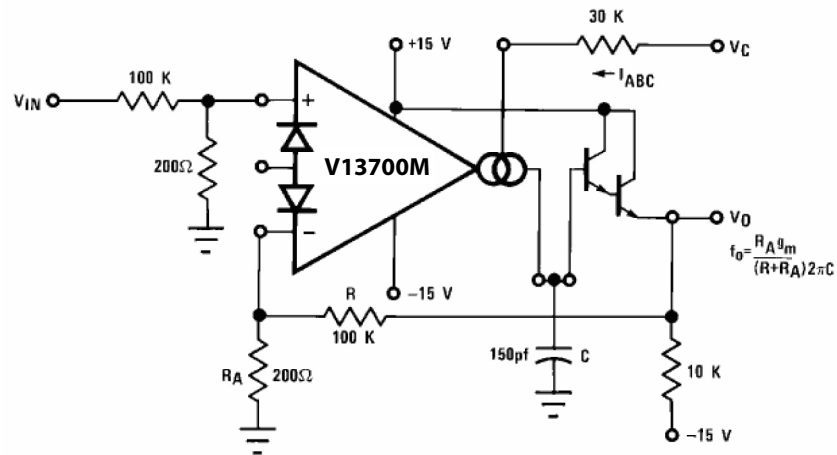
Note 5: These specifications apply for VS = ± 15 V, IABC = 500 μ A, ROUT = 5 k Ω connected from the buffer output to -VS and the input of the buffer is connected to the transconductance amplifier output.

Schematic Diagram



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Typical Application

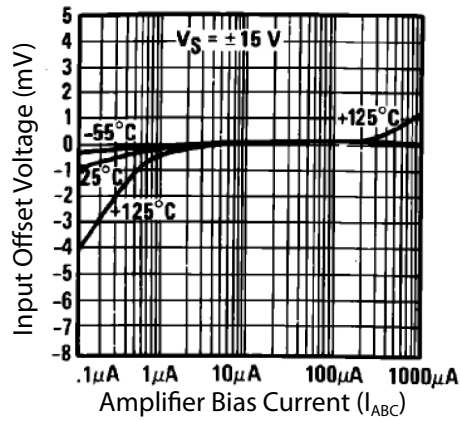


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Voltage Controlled Low-Pass Filter

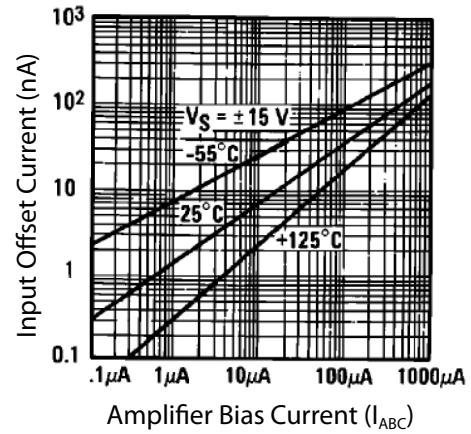
Typical Performance Characteristics

Input Offset Voltage



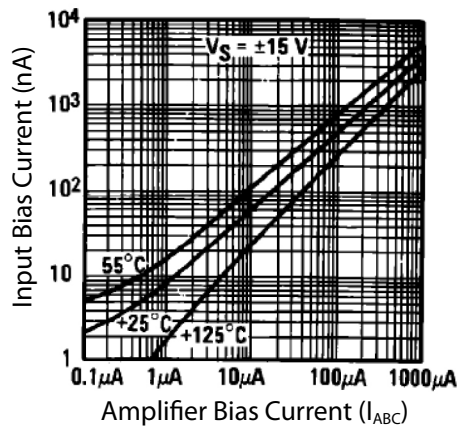
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Input Offset Current



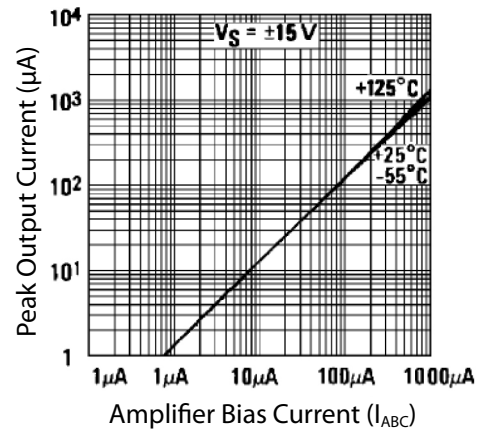
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Input Bias Current



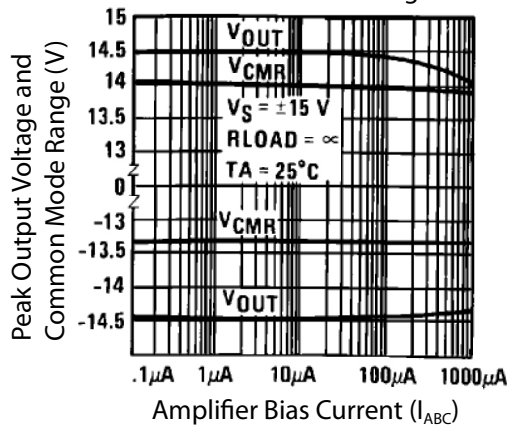
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Peak Output Current



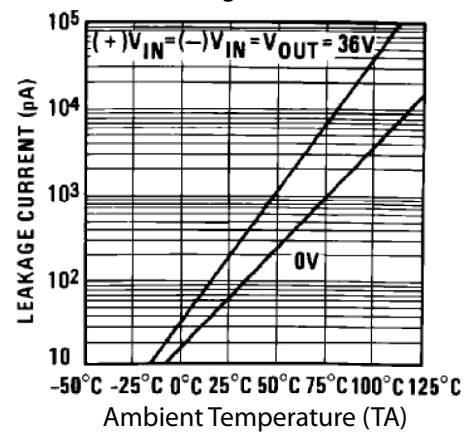
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Peak Output Voltage and Common Mode Range



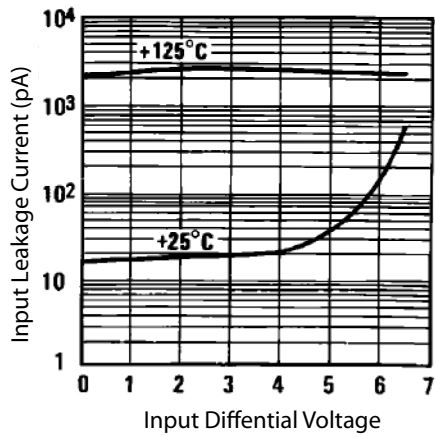
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Leakage Current



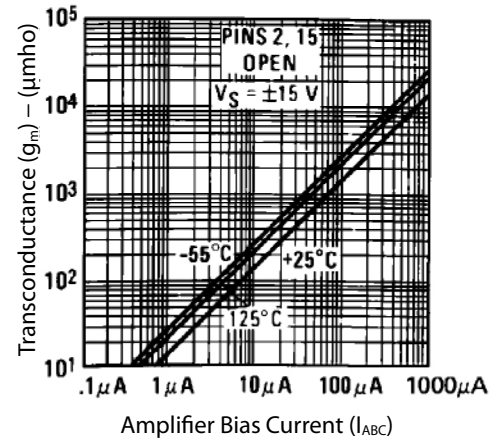
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Input Leakage



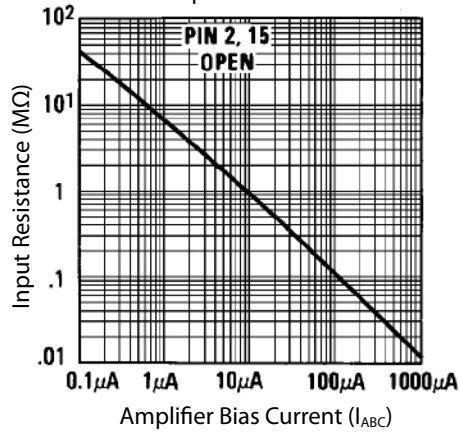
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Transconductance



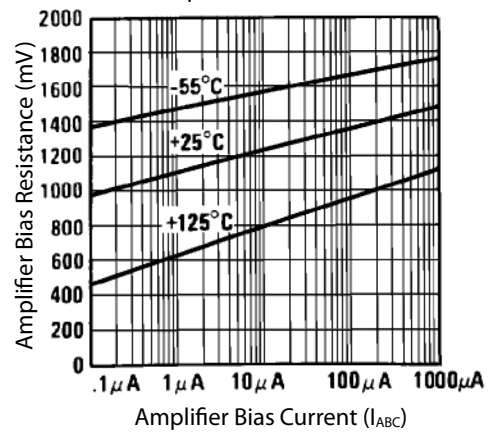
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Input Resistance



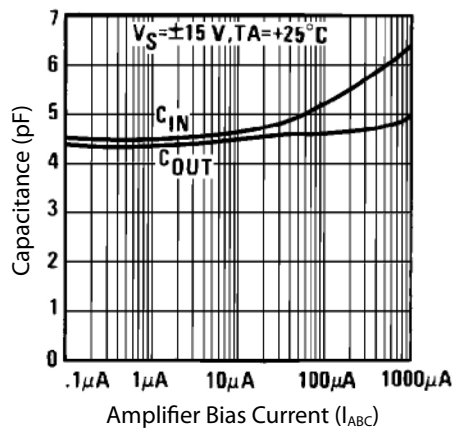
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Amplifier Bias Voltage vs Amplifier Bias Current



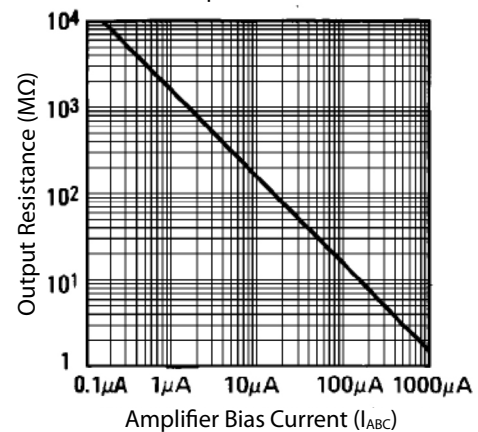
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Input and Output Capacitance



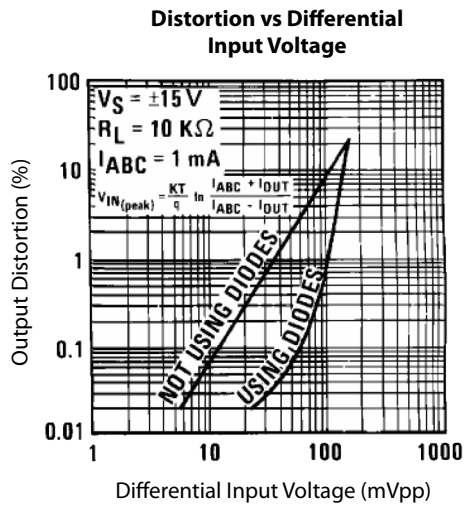
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Output Resistance

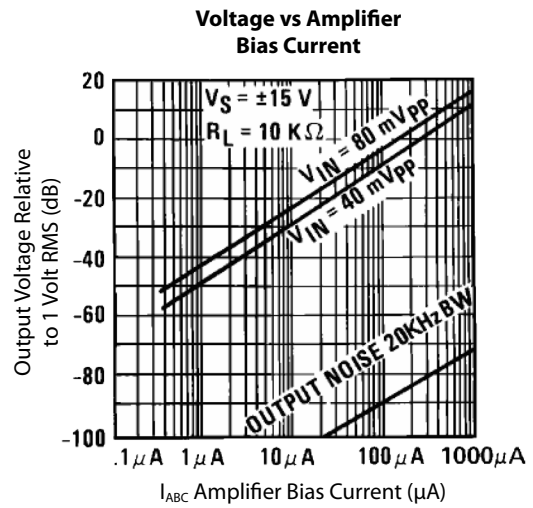


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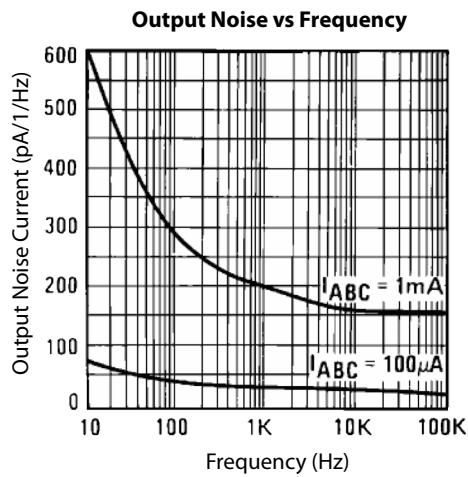
Typical Performance Characteristics (Continued)



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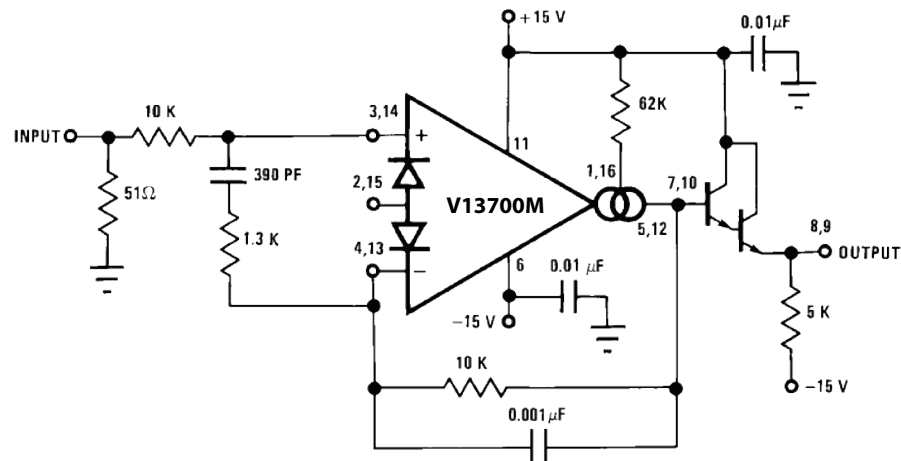


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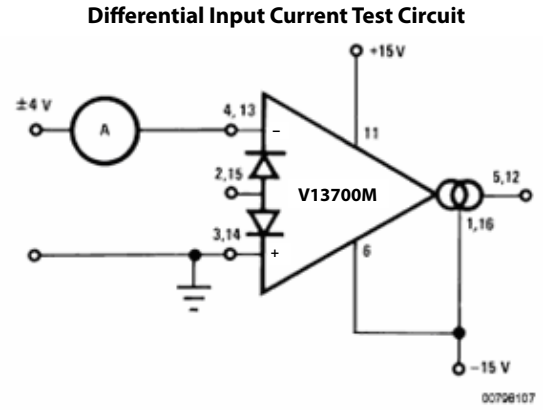
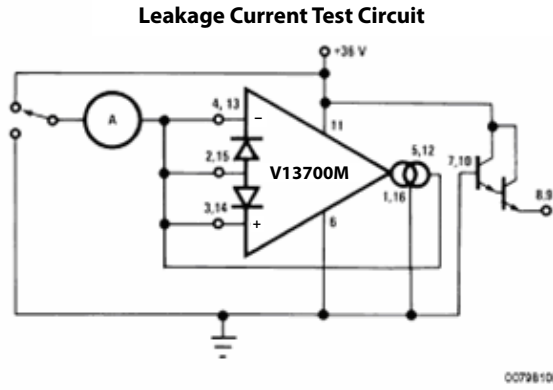


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Unity Gain Follower



Typical Performance Characteristics (Continued)



Circuit Description

The differential transistor pair Q4 and Q5 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 Q5 and Q4 respectively. With the exception of Q12 and Q13, all transistors and diodes are identical in size. Transistors Q1 and Q2 with Diode D1 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:

$$\begin{aligned} \frac{kT}{q} \ln \frac{I_5}{I_4} &\approx \frac{kT}{q} \frac{I_5 - I_4}{I_4} \\ I_5 &\approx I_4 \approx \frac{I_{ABC}}{2} \end{aligned} \quad (3)$$

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

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:

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

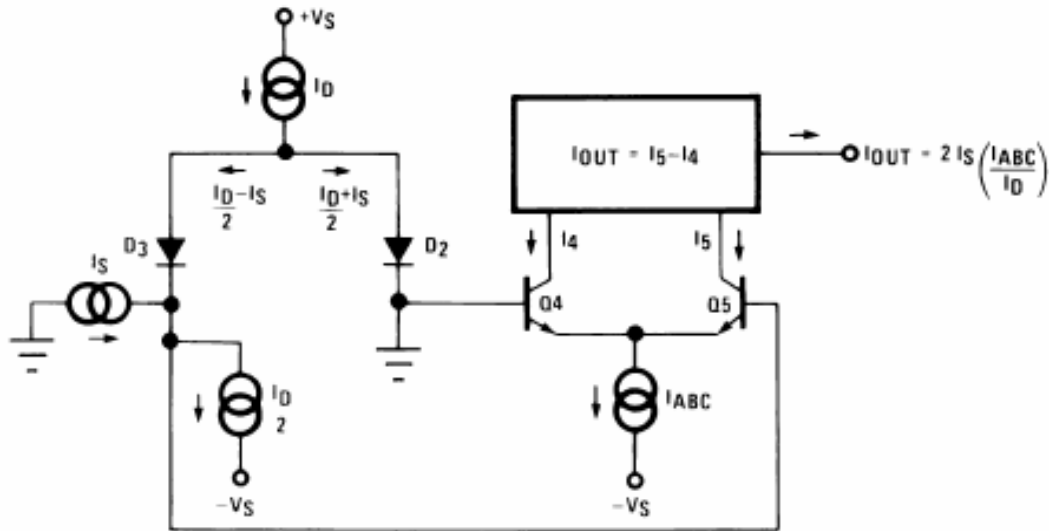
Since the diodes and the input transistors have identical geometries and are subject to similar voltages and temperatures, the following is true:

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

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.

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 Ω 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.

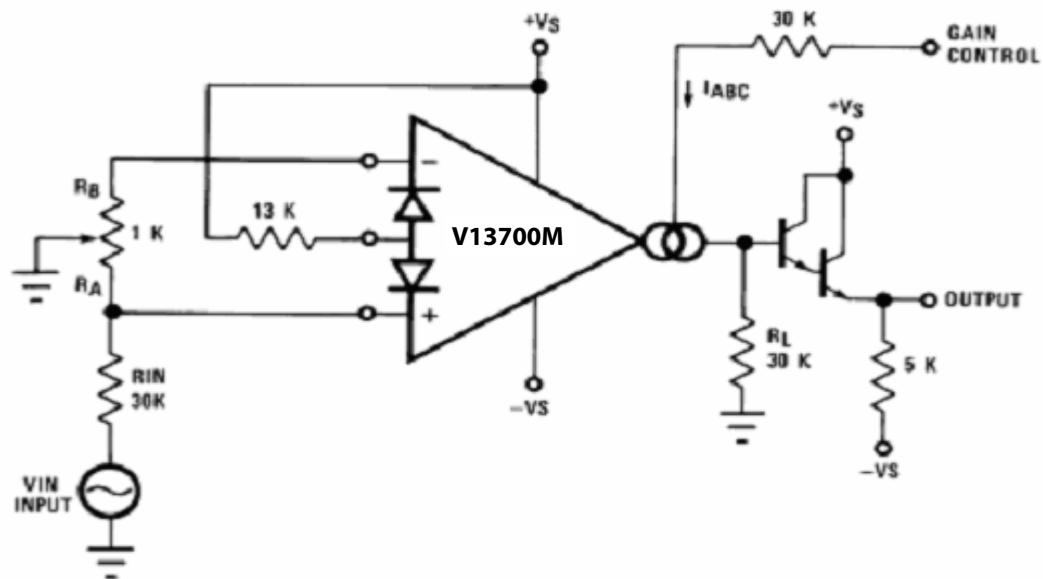


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Figure 1. Linearizing Diodes

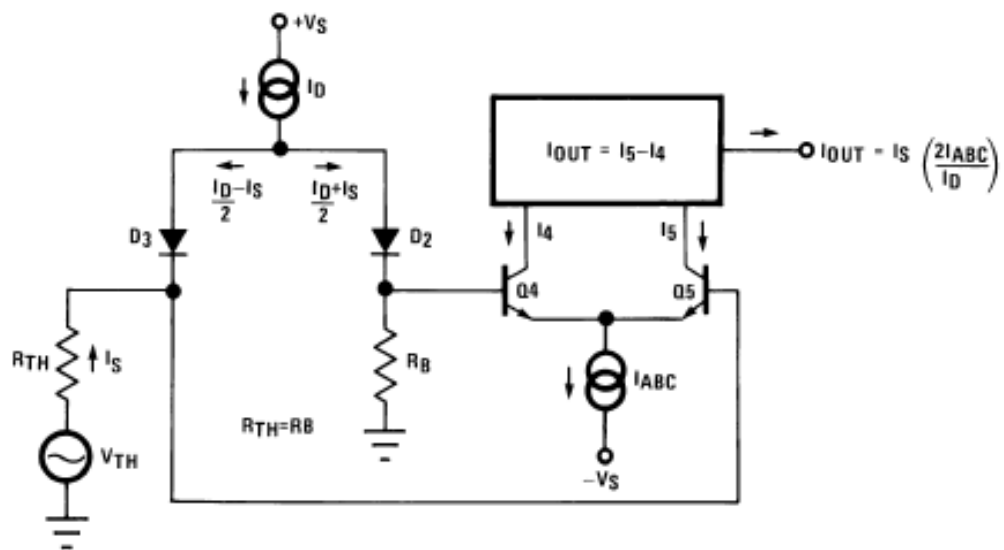
For optimum signal-to-noise performance, I_{ABC} should be as large as possible as shown by the Output Voltage vs. Amplifier Bias Current graph. Larger amplitudes of input signal also improve the S/N ratio. The linearizing diodes help here by allowing larger input signals for the same output distortion 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_e) 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.



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Figure 2. Voltage Controlled Amplifier



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Figure 3. Equivalent VCA Input Circuit

Stereo Volume Control

The circuit of Figure 4 uses the excellent matching of the two V13700M/D 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 Ω 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:

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

The constant term in the above equation may be cancelled by feeding $I_S \times I_D R_C / 2(V^- + 1.4 V)$ 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 $V_O = 0 V$ for $V_{IN2} = 0 V$. R_M also serves as the load resistor for I_O .

Stereo Volume Control (Continued)

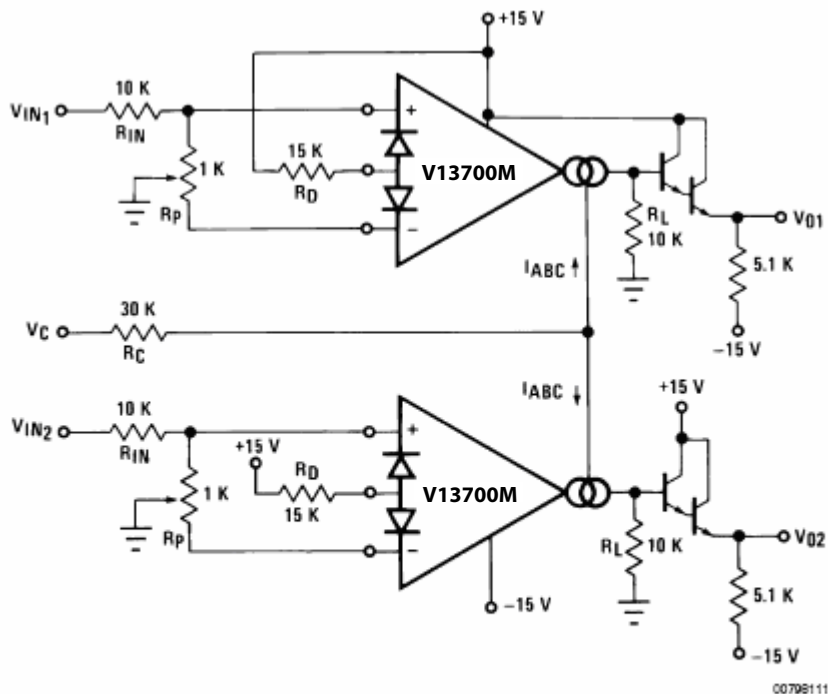


Figure 4. Stereo Volume Control

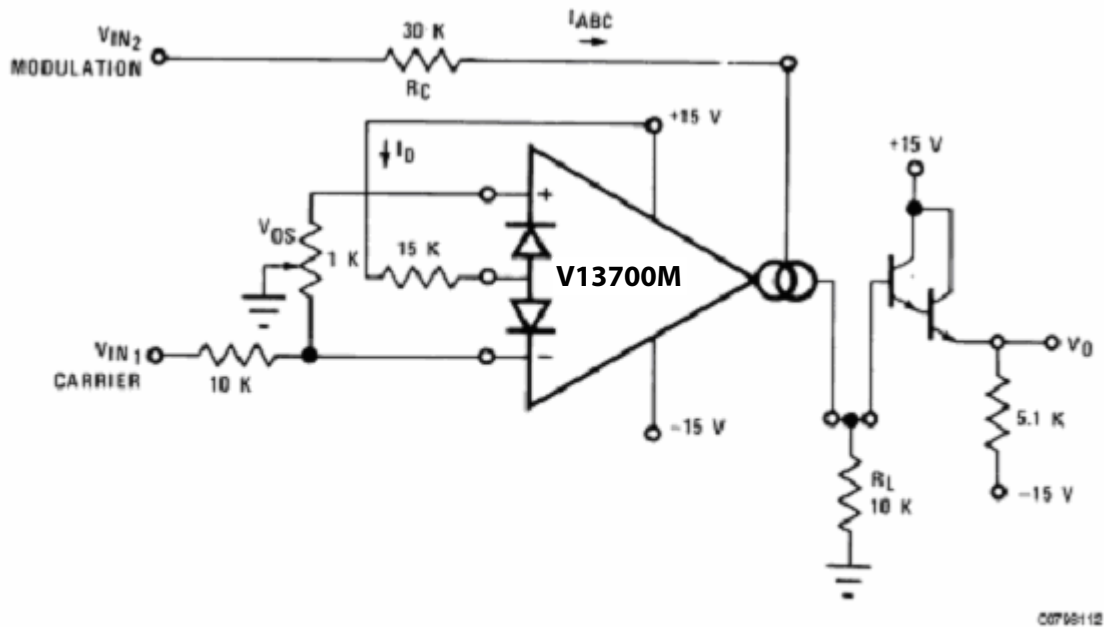


Figure 5. Amplitude Modulator

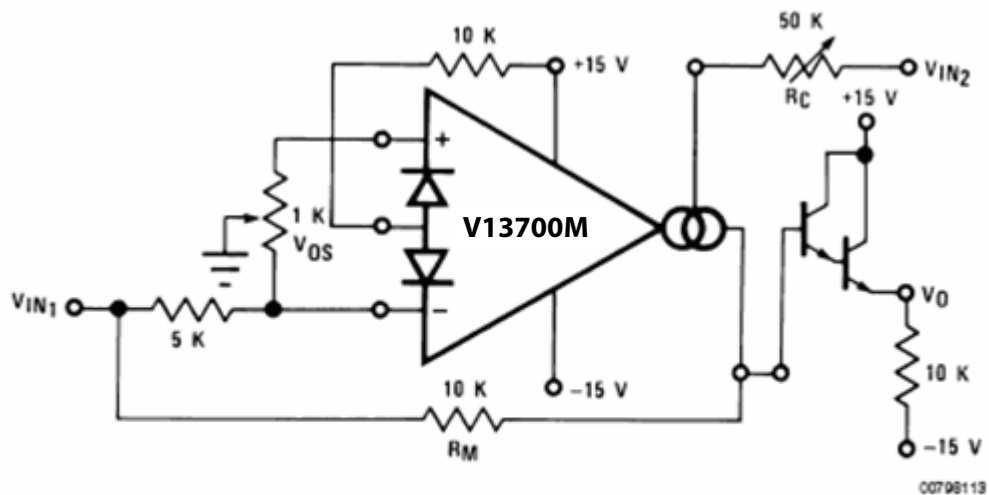
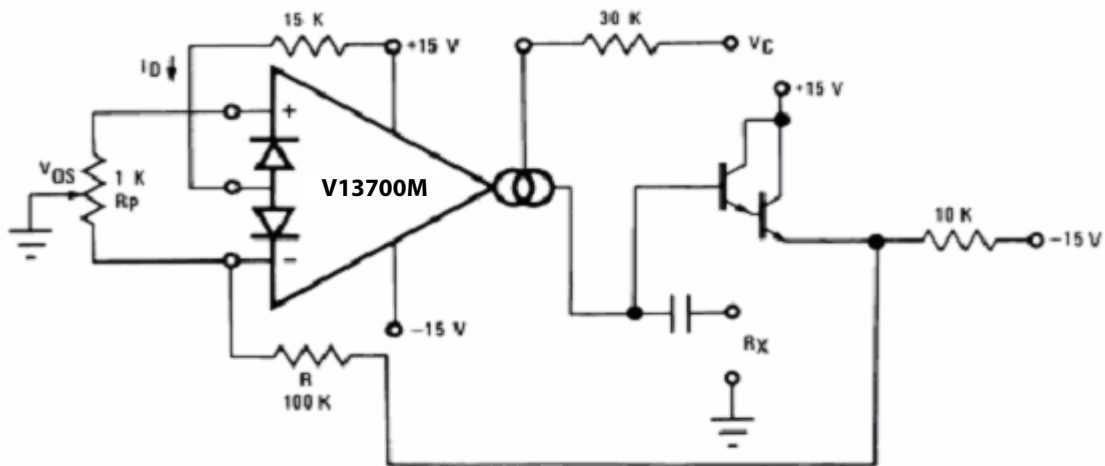


Figure 6. Four-Quadrant Multiplier

Noting that the gain of the V13700M/D 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 ($3V_{BE}$) 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.

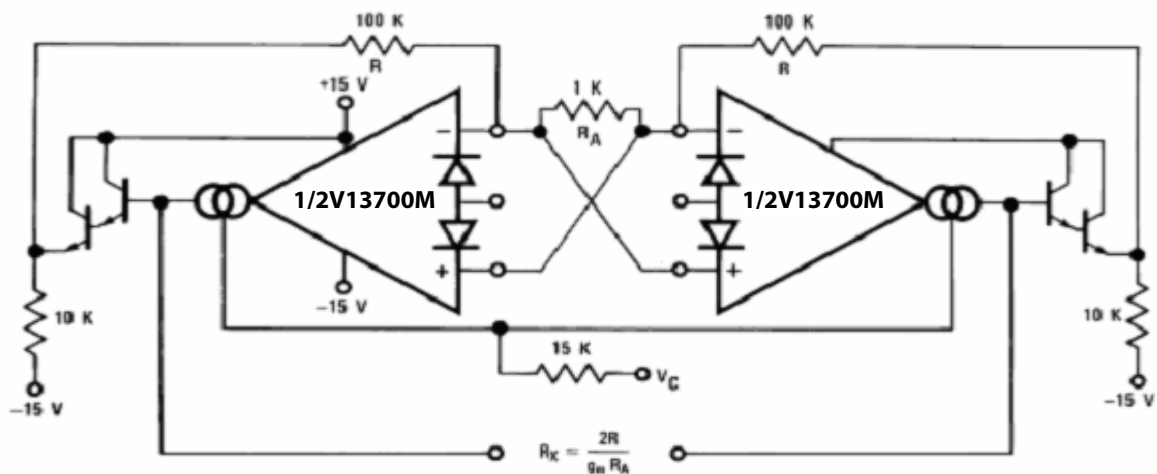


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Figure 9. Voltage Controlled Resistor with Linearizing Diodes

Voltage Controlled Filters

OTA's are extremely useful for implementing voltage controlled filters, with the V13700M/D 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, these filters perform well over several decades of frequency.



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Figure 10. Floating Voltage Controlled Resistor

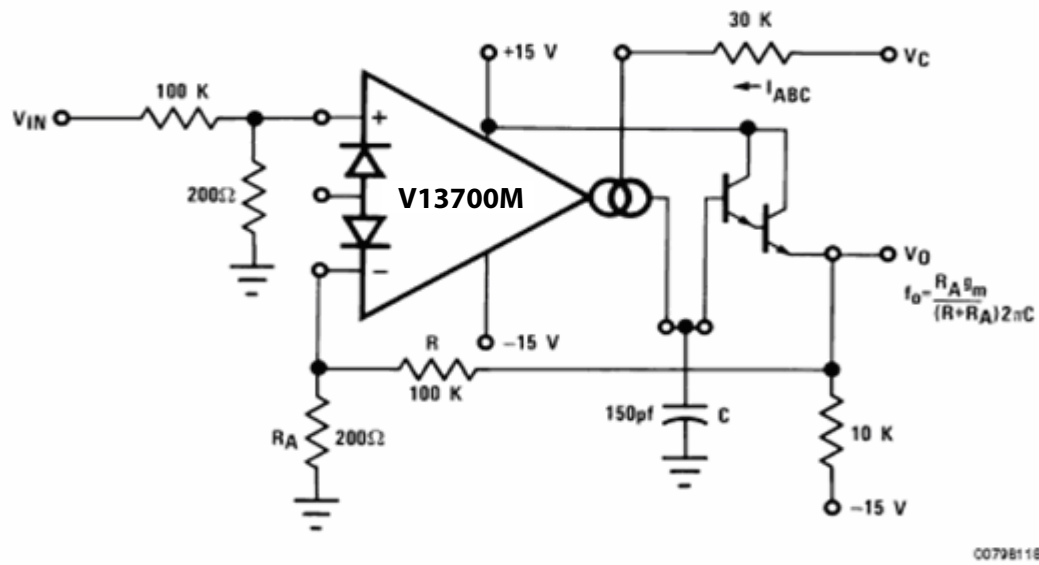


Figure 11. Voltage Controlled Low-Pass Filter

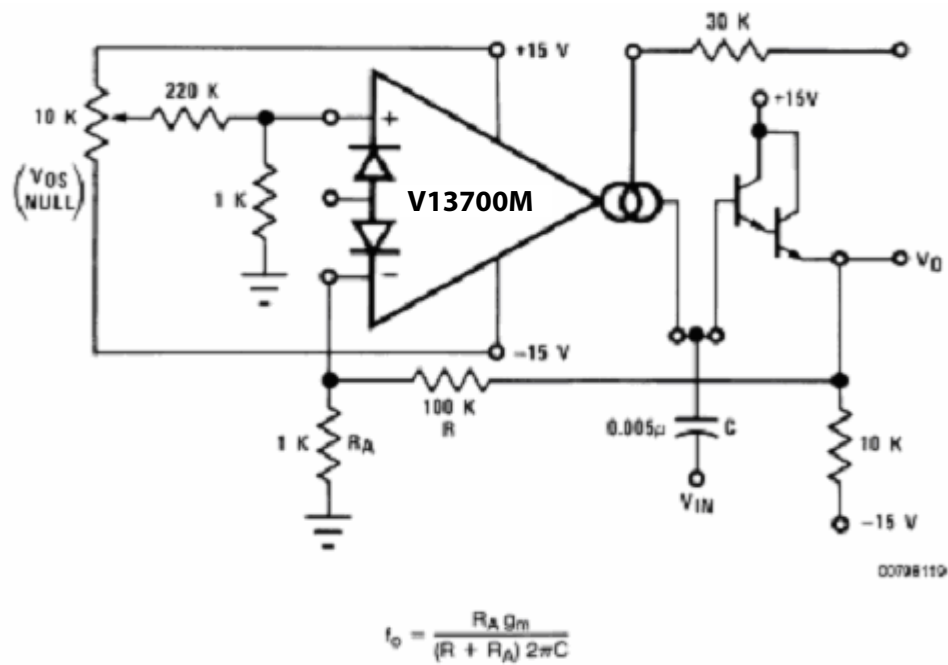
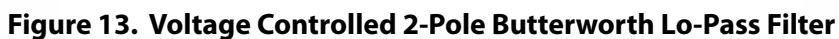


Figure 12. Voltage Controlled Hi-Pass Filter



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 V13700M/D. 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 5 V 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 V13700M/D 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° or 180° for the inverter and 60° per filter stage. This VCO operates from 5 Hz to 50 kHz with less than 1% THD.

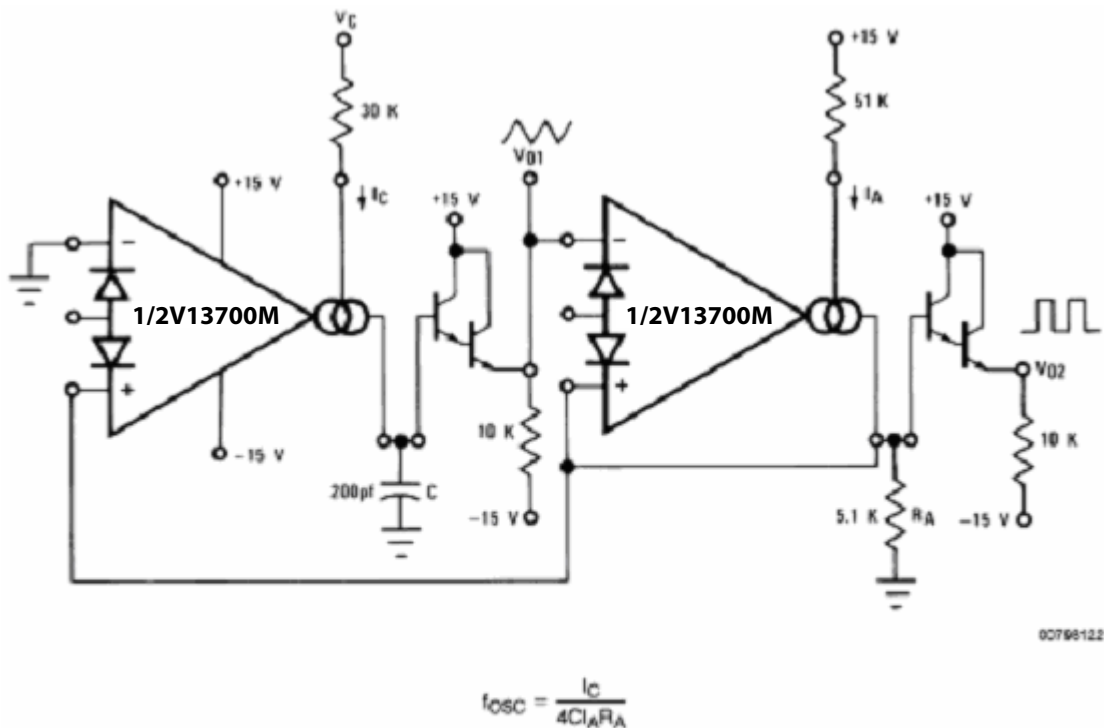
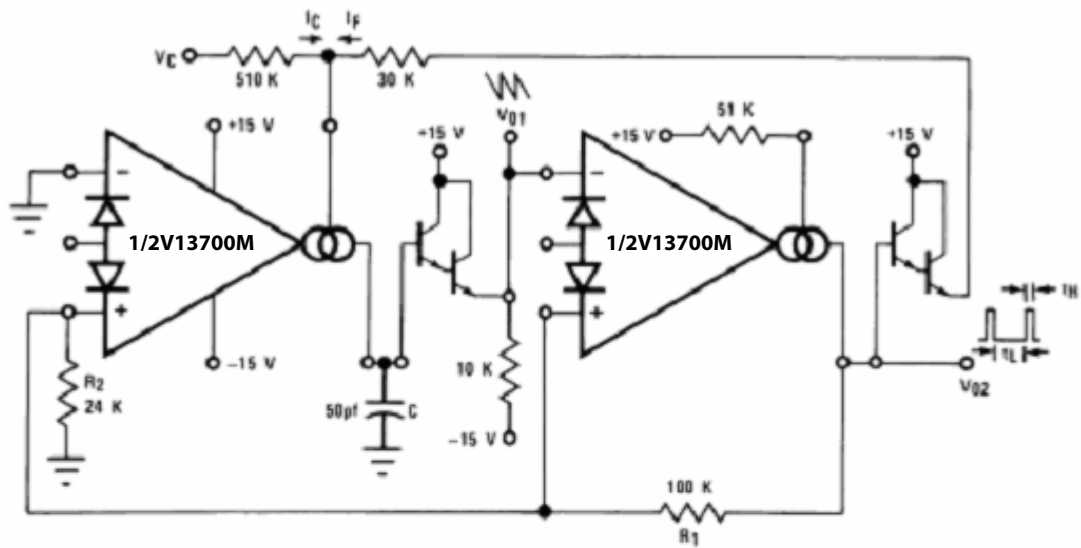


Figure 15. Triangular/Square-Wave VCO



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$$V_{PK} = \frac{(V^+ \pm 0.8V) R_2}{R_1 + R_2}$$

$$t_H \approx \frac{2V_{PK}C}{I_F}$$

$$t_L \approx \frac{2V_{PK}C}{I_C}$$

$$f_0 \approx \frac{I_C}{2V_{PK}C} \text{ for } I_C < I_F$$

Figure 16. Ramp/Pulse VCO

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 2 V amplitude turns on the amplifier through R_B 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 speeded up by shorting the diode bias pin to the inverting input so that an additional discharge current flows through D_1 when the amplifier output switches low. A special feature of this timer is that the other amplifier, when biased from V_O , can perform another function and draw zero stand-by power as well.

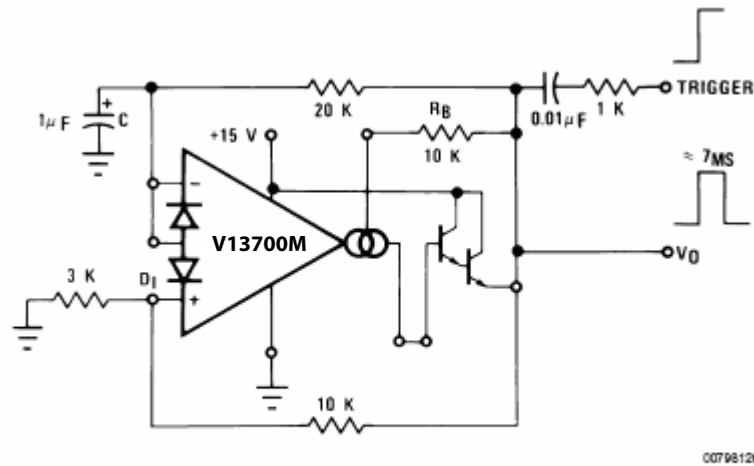


Figure 19. Zero Stand-By Power Time

The operation of the multiplexer of Figure 20 is very straightforward. When A1 is turned on it holds V_O equal to V_{IN1} and when A2 is supplied with bias current then it controls V_O . C_C and R_C serve to stabilize the unity-gain configuration of amplifiers A1 and A2. The maximum clock rate is limited to about 200 kHz by the V13700M/D slew rate into 150 pF when the $(V_{IN1}-V_{IN2})$ differential is at its maximum allowable value of 5 V. The Phase-Locked Loop of Figure 21 uses the four-quadrant multiplier of Figure 6 and the VCO of Figure 18 to produce a PLL with a $\pm 5\%$ hold-in range and an input sensitivity of about 300 mV.

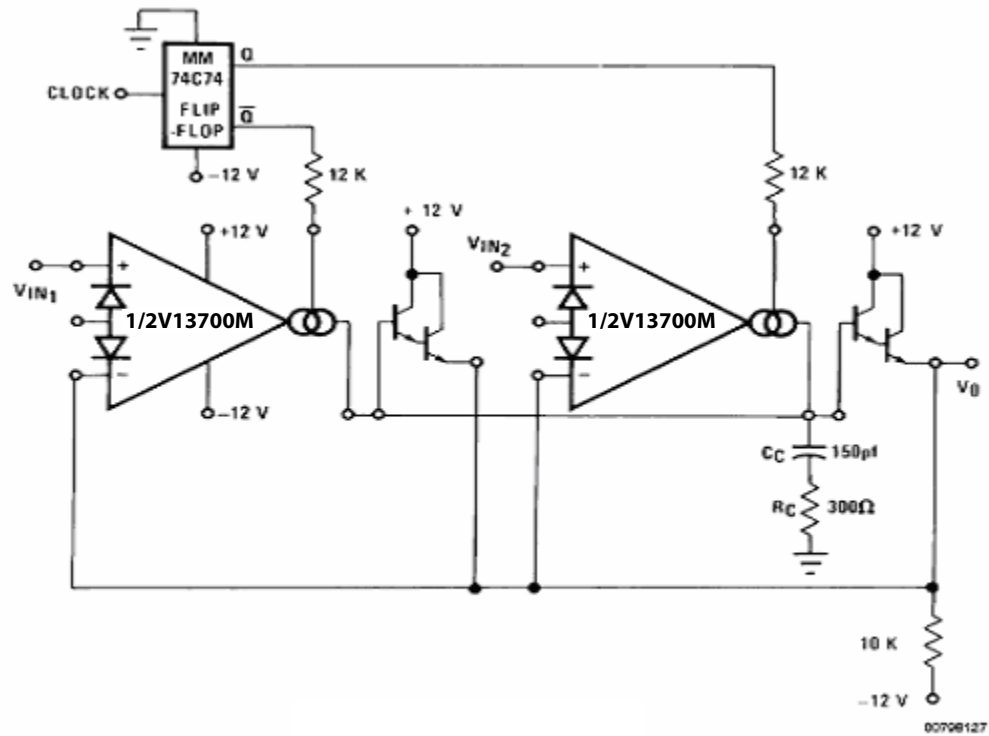


Figure 20. Multiplexer

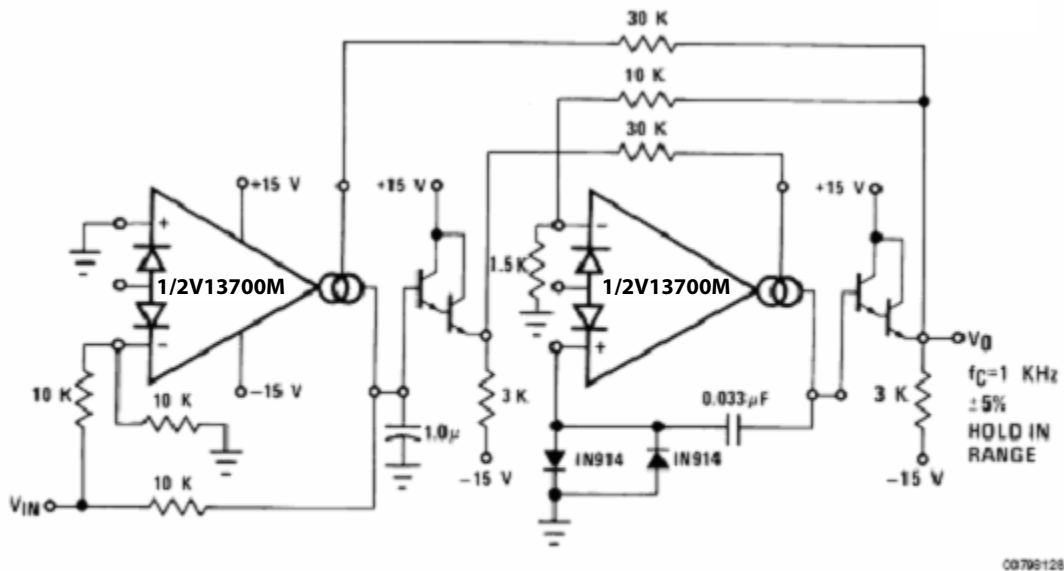


Figure 21. Phase Lock Loop

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 22. Schmitt Trigger

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 $(V_H - V_L) C_t$ is sourced into C_f and R_t . This once per cycle charge is then balanced by the current of V_O/R_t . The maximum F_{IN} is limited by the amount of time required to charge C_t from V_L to V_H with a current of I_B , where V_L and V_H represent the maximum low and maximum high output voltage swing of the V13700M/D. D1 is added to provide a discharge path for C_t 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 $V_{IN} PK$. Pulling the output of A2 low through D1 serves to turn off A1 so that V_O remains constant.



Figure 23. Tachometer

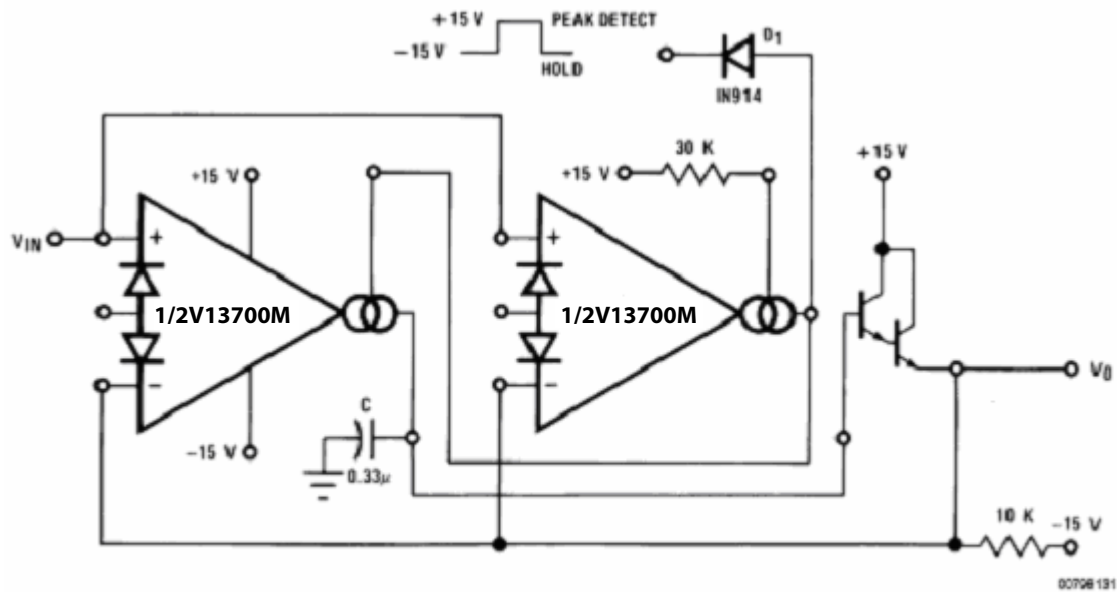
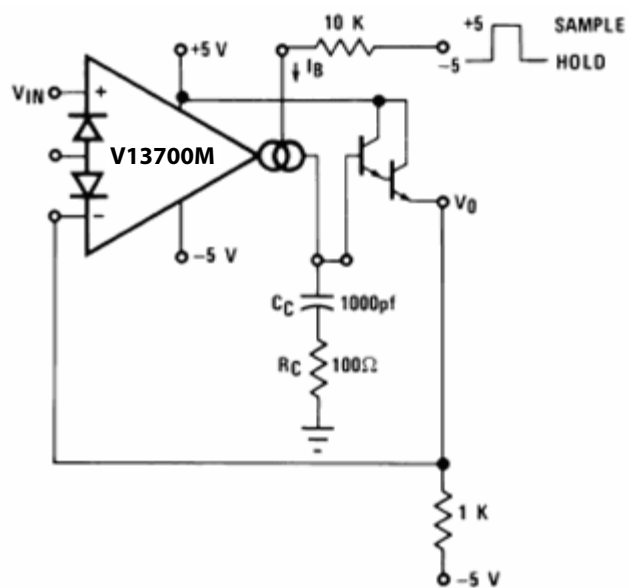


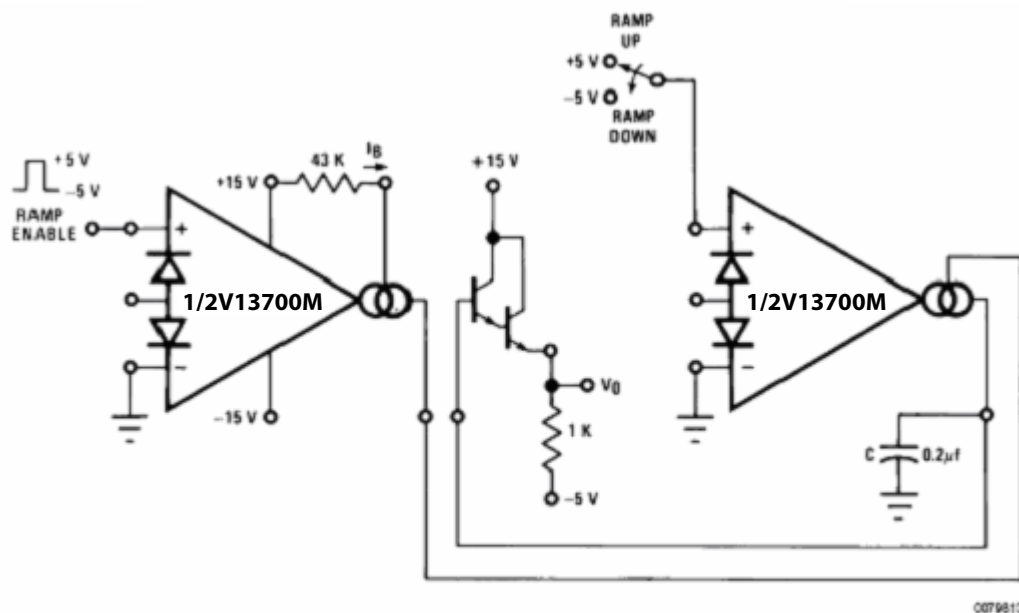
Figure 24. Peak Detector and Hold Circuit

The Ramp-and-Hold of Figure 26 sources I_B into capacitor C whenever the input to A_1 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 A_1 is constant. The output power of amplifier A_1 is monitored by squaring amplifier A_2 and the average compared to a reference voltage with amplifier A_3 . The output of A_3 provides bias current to the diodes of A_1 to attenuate the input signal. Because the output power of A_1 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 A_4 adjusts the ratio of currents through the diodes to be equal and therefore the voltage at the output of A_4 is proportional to the RMS value of the input voltage. The calibration potentiometer is set such that V_O reads directly in RMS volts.



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Figure 25. Sample-Hold Circuit



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Figure 26. Ramp and Hold

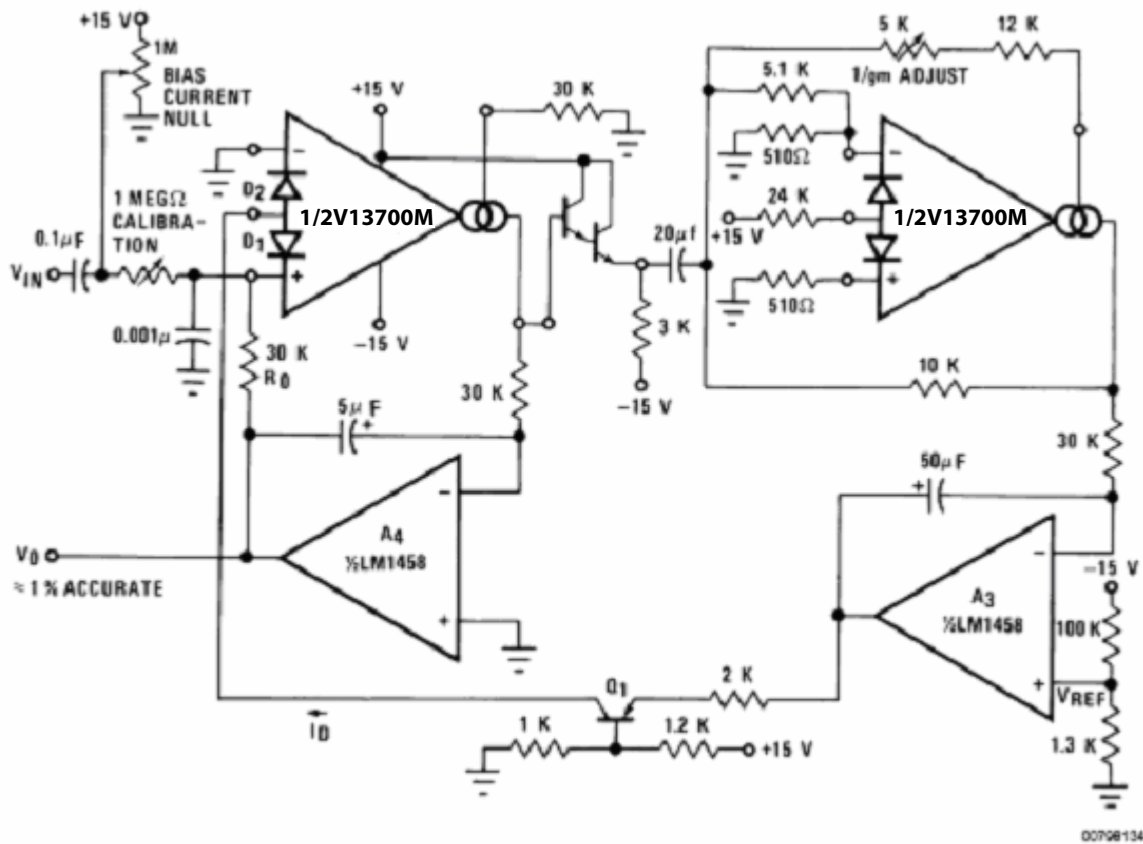


Figure 27. True RMS Converter

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 1.2 V, zero TC at about 1.2 V, and negative TC below 1.2 V. This is accomplished by balancing the TC of the A2 transfer function against the complementary TC of D1. The wide dynamic range of the V13700M/D allows easy control of the output pulse width in the Pulse Width Modulator of Figure 29. For generating IABC over a range of 4 to 6 decades of current, the system of Figure 30 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 $I_3 = -VC/R_C$. 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:

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

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

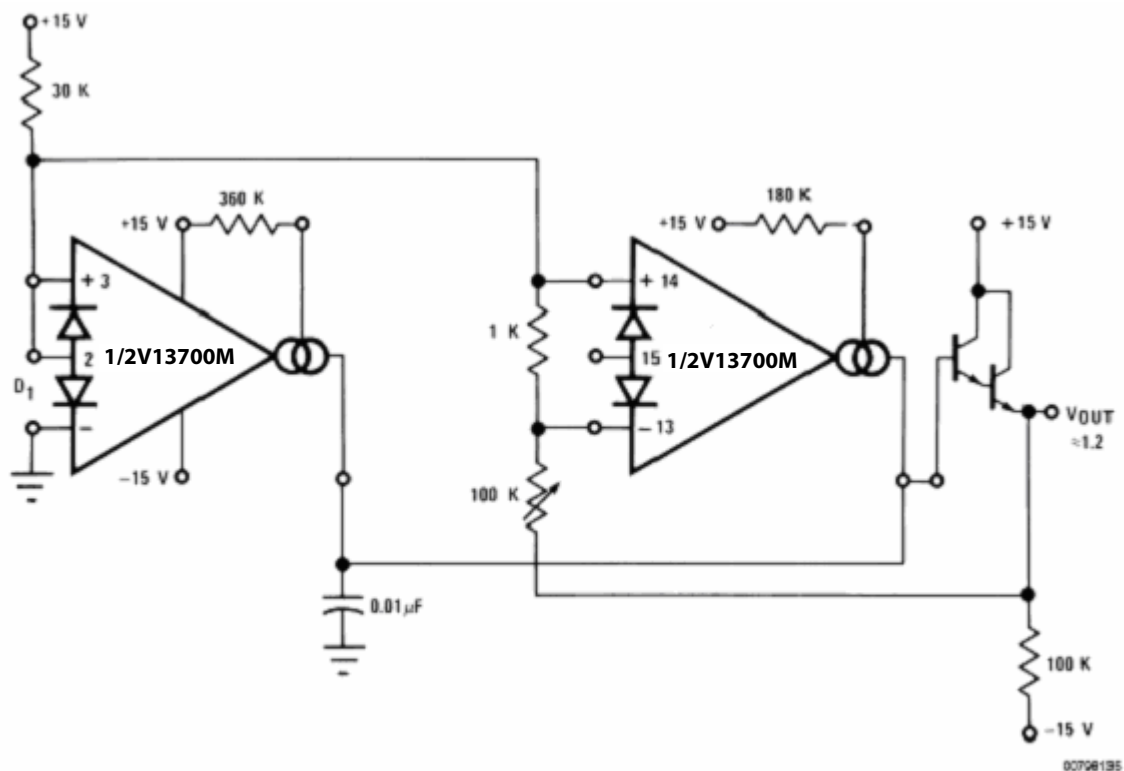


Figure 28. Delta VBE Reference

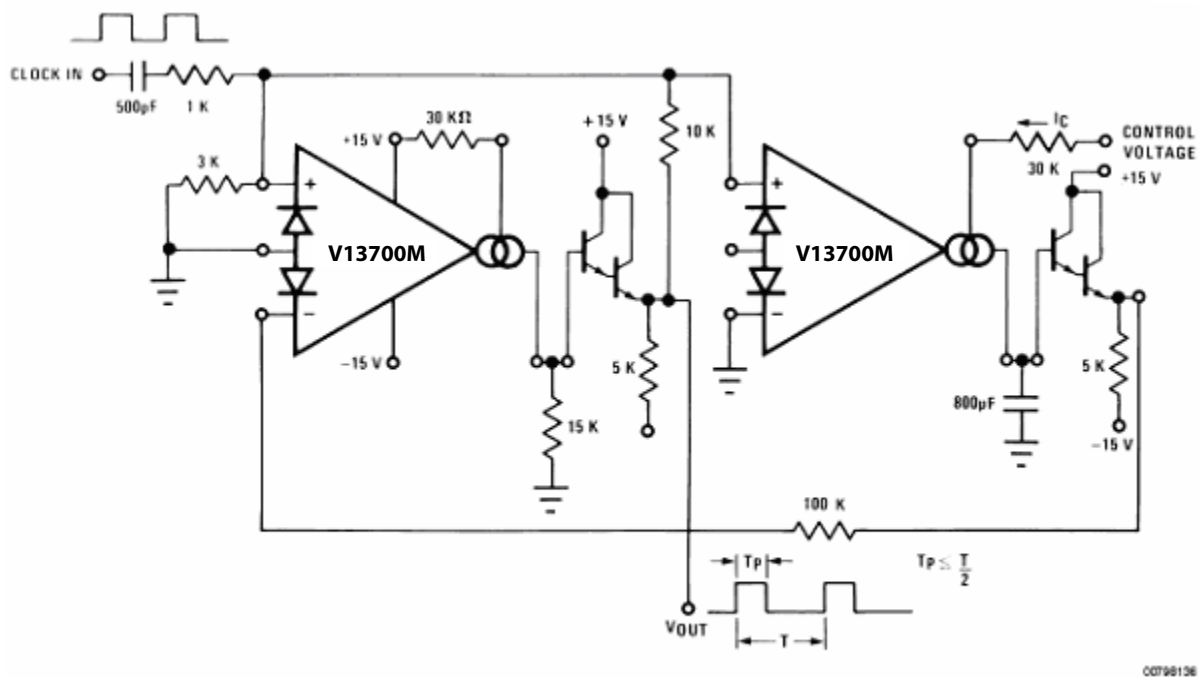


Figure 29. Pulse Width Modulator

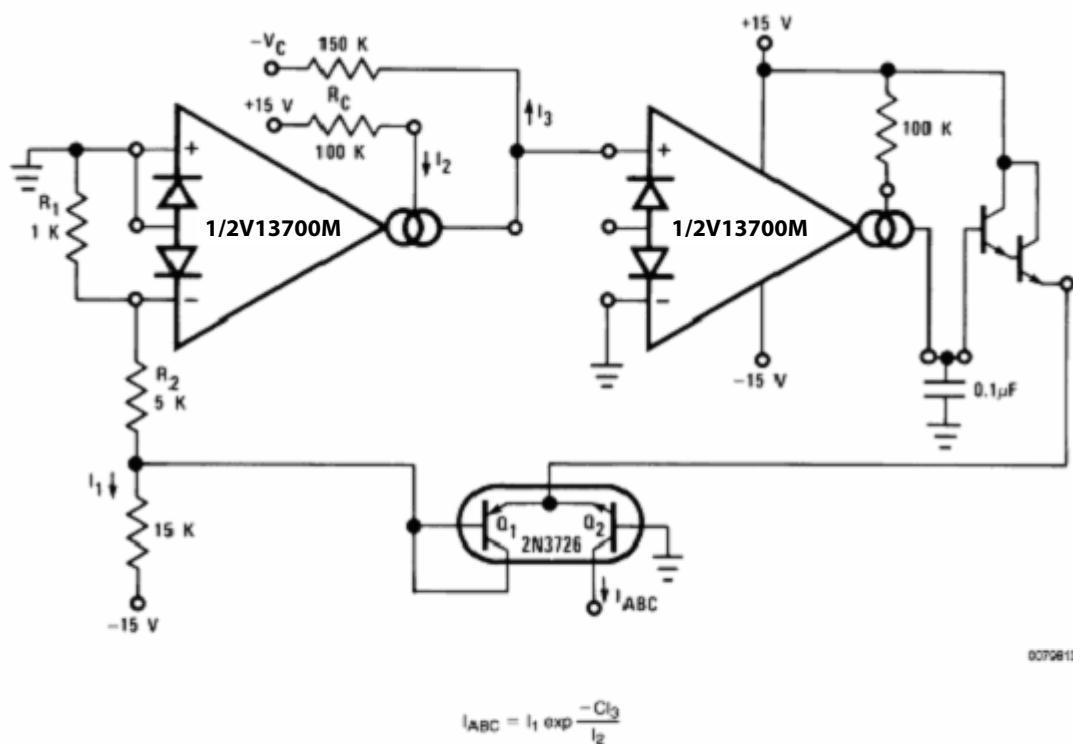
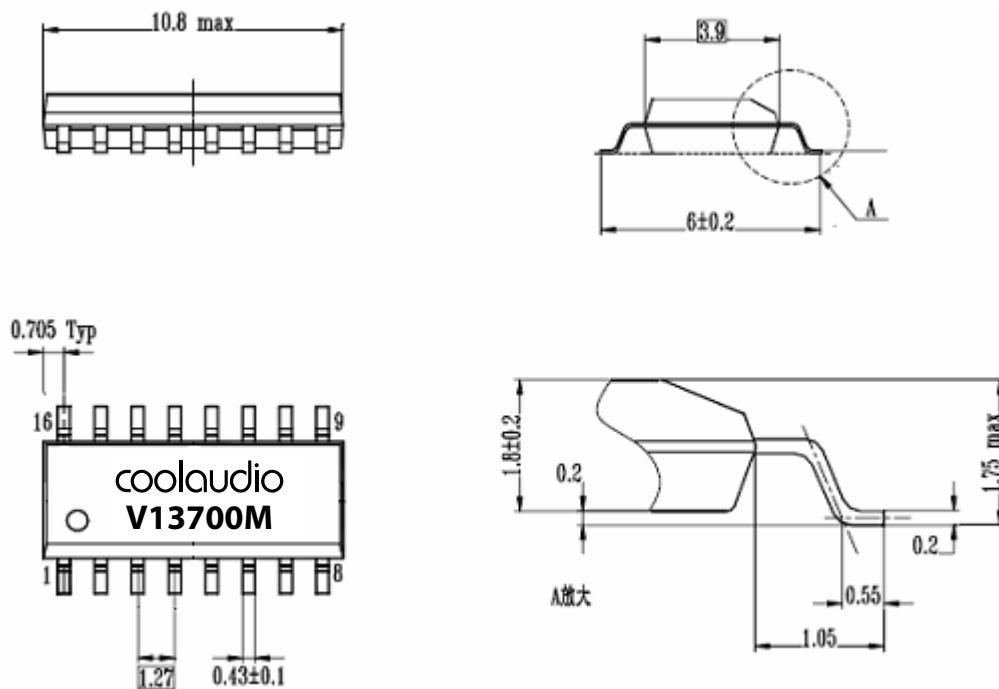


Figure 30. Logarithmic Current Source

Package Dimensions

SOP16



DIP16

