Solid State
Micro Technology
for Music

SSM 2011

SSM 2011 AUDIO PREAMPLIFIER SYSTEM

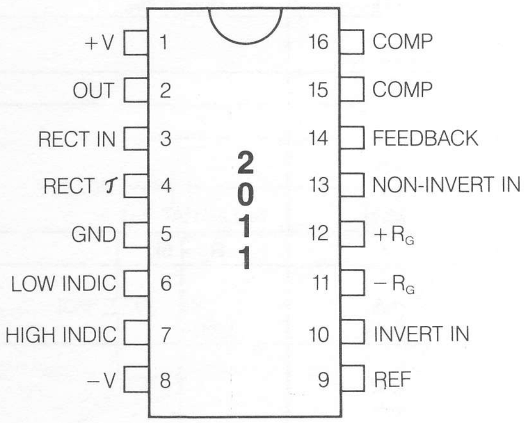
DESCRIPTION

The SSM 2011 is a low noise wide band preamplifier primarily intended for amplification of low level signals in audio applications. True differential inputs are featured with high common mode rejection and high input impedance, providing superior performance in applications where op-amps are inadequate. The balanced configuration eliminates the need for the usual large input coupling capacitors required for low noise operation. The 2011 is internally compensated for gains of one or greater while retaining a gain bandwidth product of 300MHz at a gain of 1000.

The 2011 includes level detector circuitry designed to illuminate an L.E.D. at 7dB below maximum output and another L.E.D. when the signal is greater than 20dB below the lower level. This feature can usually replace the need for individual V.U. or peak level meters in multi-channel mixer units.

FEATURES

- True Differential Inputs
- High Common Mode Rejection (120dB)
- High Supply Rejection (108dB)
- High Input Impedance (16MΩ)
- Low Input Noise (2.6 nV/$\sqrt{Hz}$)
- Very Wide Bandwidth (300kHz @ G = 1000)
- High Slew Rate (20V/μS)
- Full Internal Frequency Compensation
- Low Distortion (0.004% @ G = 100)
- No Crossover Distortion
- On Board Level Detector and High/Low L.E.D. Drivers
- Low Turn on Transient



PIN OUT (TOP VIEW)

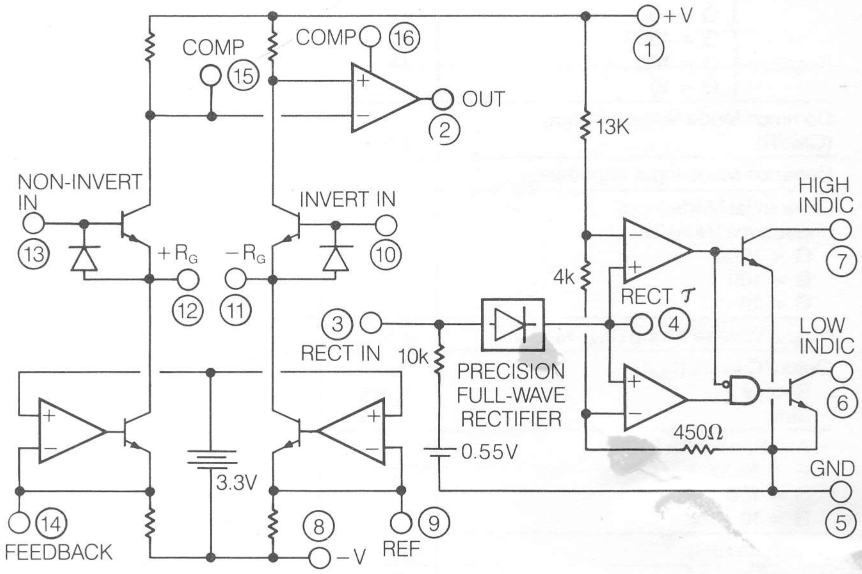


FIGURE 1. BLOCK DIAGRAM

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SPECIFICATIONS*
(PREAMPLIFIER SECTION)

OPERATING TEMPERATURE
−10°C to +55°C

STORAGE TEMPERATURE
−55°C to +125°C

The following specifications apply for $V_s = \pm 15V$, $T_A = 25°C$, configuration as shown in figure 2, unless otherwise noted.

PARAMETER (SYMBOL)	MIN	TYP	MAX	UNITS	CONDITIONS
Total Harmonic Distortion (THD) (Note 1)					
G = 1000					
@1kHz		0.035	0.05	%	V_{OUT} = 3.2V RMS
@ 5kHz		0.1	0.15	%	V_{OUT} = 3.2V RMS
@ 10kHz		0.18	0.25	%	V_{OUT} = 3.2V RMS
@ 20kHz		0.18	0.25	%	V_{OUT} = 3.2V RMS
G = 100					
@ 1kHz		0.004	0.006	%	V_{OUT} = 3.2V RMS
@ 10kHz		0.012	0.018	%	V_{OUT} = 3.2V RMS
G = 10					
@ 1kHz		0.002	0.004	%	V_{OUT} = 3.2V RMS
@ 10kHz		0.006	0.012	%	V_{OUT} = 3.2V RMS
Voltage Noise Referred to Input (E_N) (Note 1)					
G = 1000		0.45	0.6	μV RMS	30kHz Bandwidth
G = 100		1.12	1.5	μV RMS	30kHz Bandwidth
G = 10		6.4	8.5	μV RMS	30kHz Bandwidth
Input Current Noise (I_N) (Note 1)		150	230	pA RMS	30kHz Bandwidth
Gain Equation(G)		$G = \frac{2\,R_{FB}}{R_G}$			
Error From Gain Equation (ΔG)					
G = 1000		1	4	%	
G = 100		0.5	2	%	
G = 10		0.5	2	%	
Input Offset Voltage (V_{os})					
G = 1000		0.5	1.25	mV	$R_{REF} = R_{FB}$
G = 100		3	6	mV	$R_{REF} = R_{FB}$
G = 10		30	70	mV	$R_{REF} = R_{FB}$
Input Bias Current (I_b)		1.4	2.5	μA	$V_{CM} = 0_V$
Input Offset Current (I_{os})		0.1	0.3	μA	$V_{CM} = 0_V$
Common Mode Rejection Ratio (CMRR)					
G = 1000	110	120		dB	
G = 100	90	100		dB	$V_{CM} = \pm 5V$ DC to 60Hz
G = 10	70	80		dB	
Power Supply Rejection Ratio (PSRR)					
Positive G = 1000	110	120		dB	12V ≤ V+ ≤ 17V
Positive G = 100	90	100		dB	V− = −15V
Positive G = 10	70	80		dB	
Negative G = 1000	98	108		dB	−12V ≥ V− ≥ −17V
Negative G = 100	78	88		dB	+V = 15V
Negative G = 10	58	68		dB	$R_{REF} = R_{FB}$
Common Mode Voltage Range (CMVR)	±5	±7		V	
Common Mode Input Impedance		80		MΩ	
Differential Mode Input Impedance (Note 2)					
G = 1000		160		kΩ	
G = 100		1.6		MΩ	
G = 10		16		MΩ	
Output Voltage Swing (V_{OUT} MAX.)	±5	±6		V	R_L = 5k
Output Current (I_{OUT})					
Source		20		mA	
Sink		2.5		mA	
−3dB Bandwidth (GBW)					
G = 1000		300		kHz	
G = 100		600		kHz	
G = 10		1000		kHz	
Slew Rate (SR)		20		V/μS	
Supply Current (I_{SY})					
Positive		7.5	10	mA	Excluding L.E.D. current
Negative		5.5	7	mA	

Notes
1). Due to production test limitations, maximum limits for noted parameters are sample tested. 95% of units are guaranteed to meet specification, 99% of units will meet specification +50%.
2). See section on differential input impedance.

Principle of Operation

The 2011 operates on the principle of internal current feedback, provided by R_{FB} in figure 2. In conjunction with the gain setting resistor, R_G, this produces a precise differential gain of $\frac{2R_{FB}}{R_G}$.

Resistor R_{REF} provides a reference point for the output, which is usually connected to ground. The current feedback frees both inputs from feedback components, allowing the 2011 to operate in inverting, noninverting or true differential mode.

The detector circuit is AC coupled from the preamplifier output. This removes the effect of an internal offset in the precision rectifier and any offset voltage of the preamplifier. The detector recovery time constant is set for R_1 and C_2. The values shown give about 10mS.

Noise

The 2011 is capable of extremely good noise performance but this can easily be degraded by careless system design.

The first rule is to keep source resistances as small as possible since all resistors generate thermal noise which can only be reduced by refrigeration. The RMS value of thermal noise can be calculated from:

$$E_{RMS} = \sqrt{4kTR\,\Delta f}$$

where k = Boltzmann's constant (1.38×10^{-23} J/K)
T = absolute temperature (K)
R = resistor value (Ω)
f = bandwidth of interest

Added to this is the (small) effect of input noise current, which produces an RMS voltage noise equal to:

$$E_{RMS} = I_N R$$

where I_N = current noise of the device

Since the noise sources are uncorrelated, an RMS summation technique must be used and it is more convenient to refer to a general equation when performing noise calculations.

In a 30kHz bandwidth, the input referred noise of the 2011 in μV RMS reduces to:

$$E_{RMS} \simeq \sqrt{E_N^2 + \frac{I_N^2 R^2}{5 \times 10^5} + 0.48R}$$

where E_N = input voltage noise at operating gain (μV RMS)
I_N = input current noise (pA RMS) ($\simeq$ 150 pA RMS for the 2011)
R = effective resistance between the input terminals in kΩ

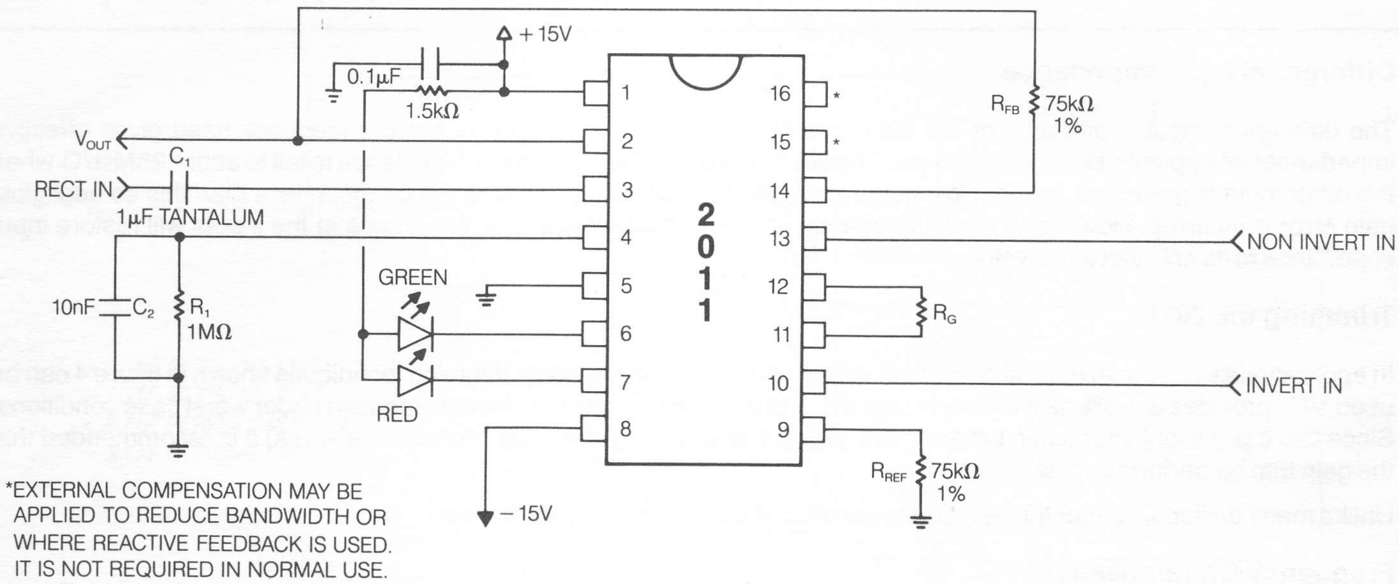


FIGURE 2. STANDARD CONFIGURATION. $R_G = \frac{150k\Omega}{\text{REQUIRED GAIN}}$

Additional noise can be generated by the use of low quality resistors in the 2011 gain setting network. Some resistors (notably carbon composition) generate excess noise proportional to the current flowing through them. Since both feedback and reference resistors are subject to standing currents, this represents a potential noise source. Experiments show that the excess noise is lower than commonly believed, but for applications requiring lowest possible noise it is recommended that metal film, cermet or wirewound resistors be used; though carbon film also gives a closely comparable noise performance.

2011 Inputs

Although the 2011 inputs are fully floating, care must be exercised to ensure that both inputs have a DC bias connection capable of maintaining them within the input common mode range. The usual method of achieving this is to ground one side of a transducer figure 3(a), but an alternative way is to float the transducer and use two resistors to set the bias point figure 3(b). The value of these resistors can be up to 10kΩ or so, but they should be kept as small as possible to limit common mode noise. Noise generated in the resistors themselves is negligible since it is attenuated by the transducer impedance. Balanced transducers give the best noise immunity, and interface directly as in figure 3(c).

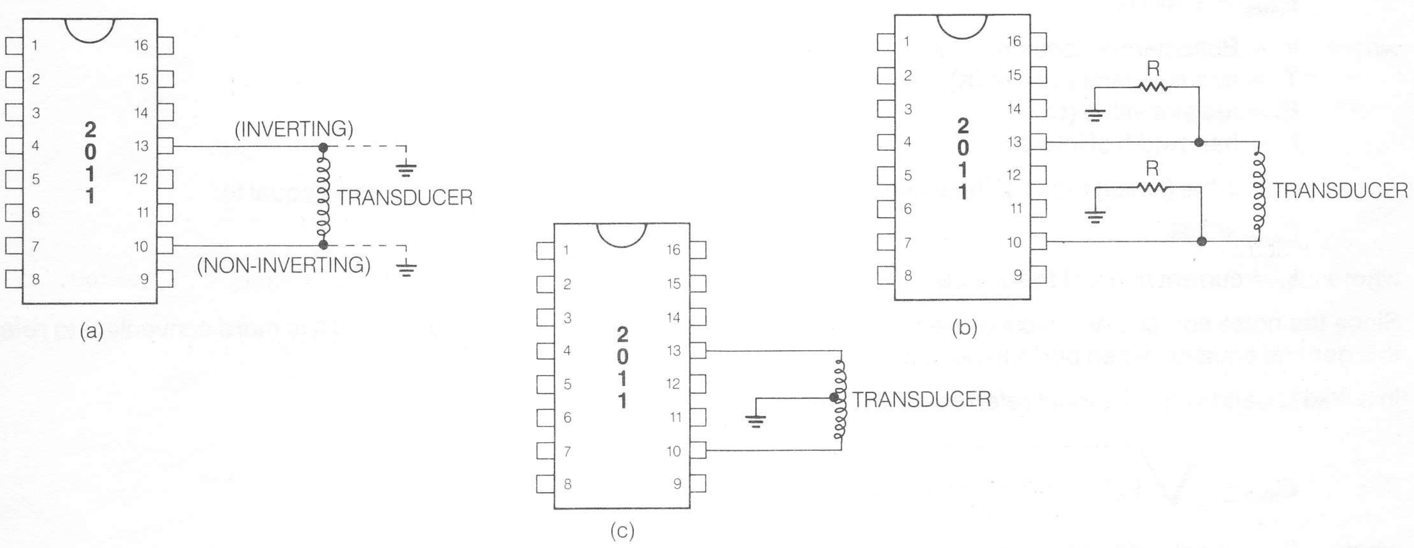


FIGURE 3. THREE WAYS OF INTERFACING TRANSDUCERS FOR HIGH NOISE IMMUNITY.
(a) Single ended. (b) Pseudo differential. (c) True differential.

Differential Input Impedance

The differential input impedance of the 2011 is quite high, however the type of current feedback used gives effective impedances of opposite signs at each input. The effect of this is to cause the input impedance to fall to about 25MΩ/G when the other input is grounded. In most low noise applications, source impedances will be much less than this so negligible gain error is incurred. However, if input impedance is critical, balancing source resistance at the inputs will restore input impedance to its full differential value.

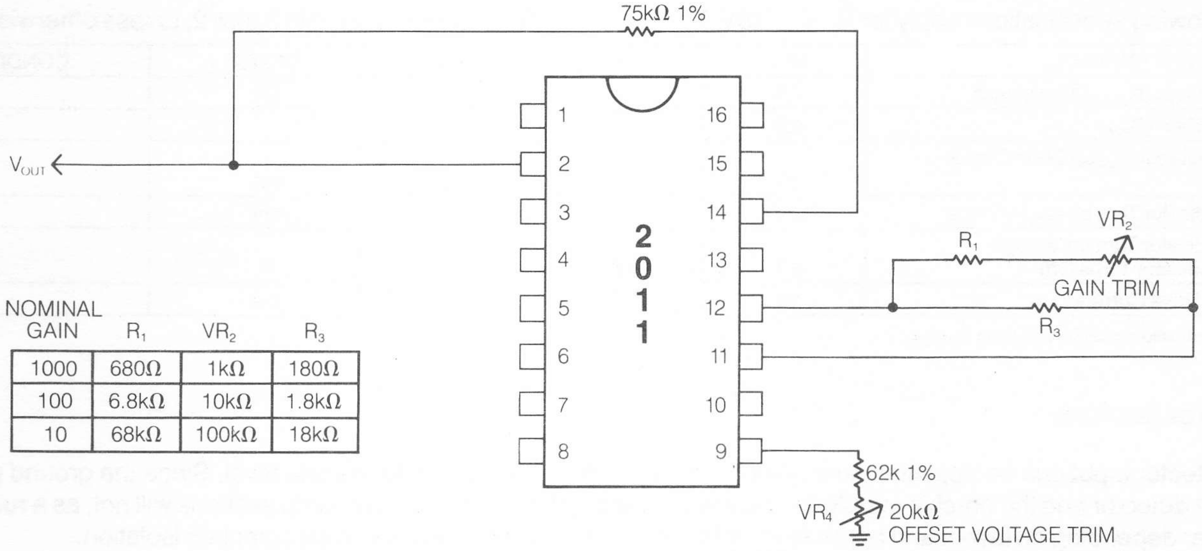
Trimming the 2011

In applications where gain error and/or offset voltage are critical, the optional trimming techniques shown in figure 4 can be used. VR_2 provides a ±5% gain error trim and VR_4 has enough range to trim the offset to zero under worst case conditions. Since there is a slight interaction between the gain trim and the offset voltage (but not vice versa) it is recommended that the gain trim be performed first.

Unlike many designs, neither trim affects the common mode rejection performance.

Frequency Compensation

The 2011 is stable at all gains, but at $G<5$ there is some peaking in the response characteristic. Where this is undesirable, a capacitor of around 5pF should be connected between pins 15 and 16. This yields a bandwidth of about 500kHz. These pins can also be used to provide overcompensation in applications where excess bandwidth is a disadvantage.



NOMINAL GAIN	R_1	VR_2	R_3
1000	680Ω	1kΩ	180Ω
100	6.8kΩ	10kΩ	1.8kΩ
10	68kΩ	100kΩ	18kΩ

FIGURE 4. TRIMMING THE 2011

Tape Playback Preamplifier

Since the output of a tape head rises at 6dB/octave, it is usual to perform preamplification and equalization simultaneously. Care must be taken to ensure that noise performance is not degraded, and the circuit of figure 5 achieves this by boosting gain at low frequencies and cutting at high frequencies. The overall gain obeys the 2011 gain equation at 450Hz with C_1 = 6.8nF and 600Hz with C_1 = 2.7nF. Maximum boost of about 20dB occurs at 30Hz and maximum cut of about −10dB occurs at greater than 3.4kHz with C_1 = 6.8nF and greater than 5.5kHz with C_1 = 2.7nF. These correspond to the following asymptotic corner frequencies:

3¾, 1⅞ I.P.S.	f_c = 1.8kHz
7½, 15 I.P.S.	f_c = 3.2kHz
Ferric Cassette	f_c = 1.3kHz
Chrome Cassette	f_c = 2.3kHz

This closely approximates the N.A.B./D.I.N. standards for tape equalization. C_2 effectively rolls off gain at low frequencies, producing a DC gain as given by the 2011 gain equation.

Note that unlike op-amp equalization circuits gain can be varied (or adjusted) over a wide range by a choice of R_G, without affecting the frequency response characteristic.

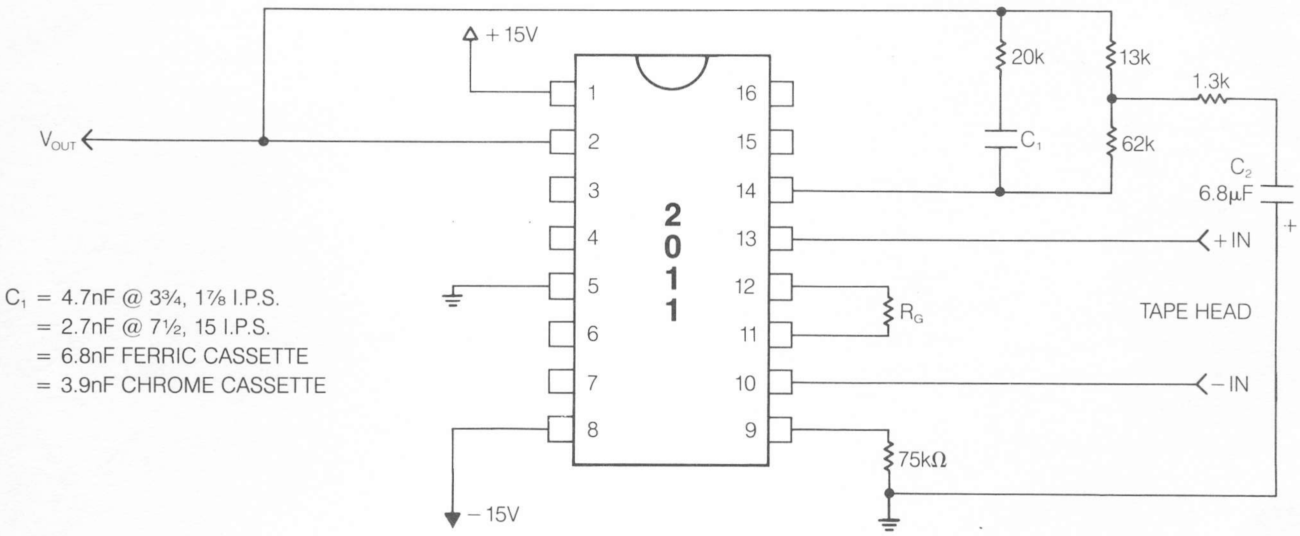


FIGURE 5. TAPE PLAYBACK PREAMPLIFIER

SPECIFICATIONS* (DETECTOR SECTION) **OPERATING TEMPERATURE −10°C to +55°C** **STORAGE TEMPERATURE −55°C to +125°C**

The following specifications apply for $V_s = \pm 15V$, $T_A = 25°C$, configuration as shown in figure 2, unless otherwise noted.

PARAMETER (SYMBOL)	MIN	TYP	MAX	UNITS	CONDITIONS
Rectifier Input ($R_{REC\ IN}$) Resistance	7.5	10	12.5	kΩ	
Rectifier Gain (G_{REC})	0.9	1	1.1		
Rectifier Output ($I_{REC\ SS}$) Short Circuit Current	10	25	40	mA	
Lower Indicator Turn on (V_{TL}) Voltage	340	385	450	mV	
Upper Indicator Turn on Voltage (Lower Indicator Turns Off)	3.7	3.85	4.0	V	
Indicator Drive Current	10	25		mA	

*Final specifications may be subject to change.

Detector Section

The detector input can be applied to any point in the signal processing path to indicate level. Since the ground pin serves only the detector and the on-chip supply lines are well isolated, the detector and preamp sections will not, as a rule interact. However depending on layout and application, additional steps can be taken to ensure complete isolation.

One possibility is to power the L.E.D. s from a separate positive supply. Another way is to use a voltage clamp such as the one in figure 6, so that the current drawn from the positive supply by the L.E.D. circuit is almost constant regardless of input signal condition.

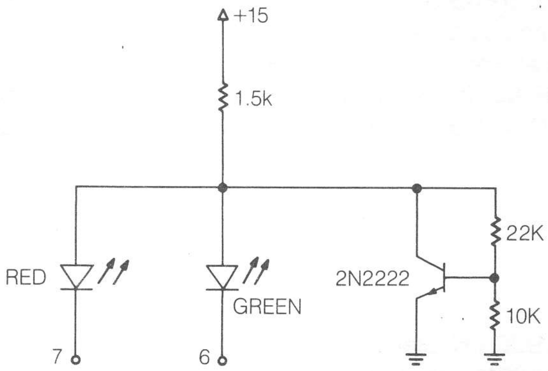


FIGURE 6. L.E.D. VOLTAGE CLAMP