

LM1897 Low Noise Preamplifier for Tape Playback Systems

General Description

The LM1897 is a dual high gain preamplifier for applications requiring optimum noise performance. It is an ideal choice for a tape playback amplifier when a combination of low noise, high gain, good power supply rejection, and no power up transients are desired. The application also provides transient-free muting with a single pole grounding switch.

Features

- Programmable turn-on delay
- Transient-free power up—no pops
- Transient-free muting
- Low noise— $0.6 \mu\text{V}$ CCIR/ARM in a DIN circuit referenced to gain at 1 kHz

- Low Voltage Battery Operation 4V
- Wide gain bandwidth due to broadband two amplifier approach 76 dB @ 20 kHz
- High power supply rejection 105 dB
- Low distortion 0.03%
- Fast slew rate $6\text{V}/\mu\text{s}$
- Short circuit protection
- Internal diodes for diode switching applications
- Low cost external parts
- Excellent low frequency response
- Prevents "click" from being recorded onto the tape during power supply cycling in tape playback applications

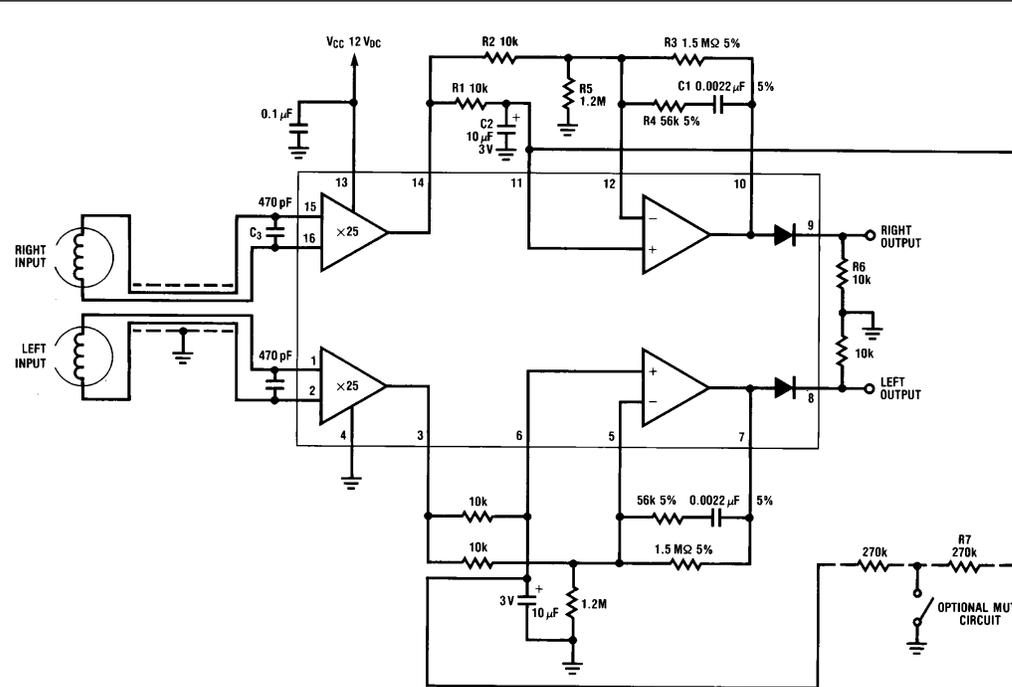


FIGURE 1. Typical Tape Playback Preamplifier Application

TL/H/7094-1

Order Number LM1897N
See NS Package Number N16E

LM1897 Low Noise Preamplifier for Tape Playback Systems

Absolute Maximum Ratings

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

Supply Voltage	18V
Voltage on Pins 8 and 9	18V
Package Dissipation (Note 1)	715 mW

Storage Temperature	-65°C to +150°C
Operating Temperature	0°C to +70°C
Minimum Voltage On Any Pin	-0.1 V _{DC}
Lead Temperature (soldering, 10 sec.)	260°C

Electrical Characteristics (T_A = 25°C, V_{CC} = 12V, See Circuit—Figure 2)

Parameter	Conditions	Min	Typ	Max	Units			
Operating Supply Voltage Range	R ₅ removed from circuit	4		18	V			
Supply Current	V _{CC} = 12V		6	12	mA			
Total Harmonic Distortion	f = 1 kHz, V _{IN} = 0.3 mV, Pins 7 & 10, Figure 2		0.03		%			
THD + Noise (Note 2)	f = 1 kHz, V _{OUT} = 1V, Pins 7 & 10, Figure 2		0.10	0.25	%			
Power Supply Rejection	Input Ref. f = 1 kHz, 1 V _{RMS}	85	105		dB			
Channel Separation	f = 1 kHz, Output = 1 V _{RMS} , Output to Output	40	60		dB			
Signal to Noise (Note 3)	Unweighted 32 Hz–12.74 kHz (Note 2) CCIR/ARM (Note 4) A Weighted CCIR, Peak (Note 5)		58		dB			
			62		dB			
			64		dB			
			52		dB			
Noise	Output Voltage CCIR/ARM (Note 4)		120	200	μV			
Input Amplifiers	f = 1 kHz		Input Bias Current	0.5	2.0	μA		
			Input Impedance	50		kΩ		
			A.C. Gain	27	28	29	dB	
			A.C. Gain Imbalance		±0.15	±0.5	dB	
			D.C. Output Voltage	1.8	2.2	2.6	V	
			D.C. Output Voltage Mismatch	Pins 3 and 14	-200	±30	+200	mV
			Output Source Current	Pins 3 and 14	2	10		mA
			Output Sink Current	Pins 3 and 14	300	600		μA
Output Amplifiers	Stable Operation D.C.		Closed Loop Gain	5		V/V		
			Open Loop Voltage Gain		110		dB	
			Gain Bandwidth Product		5		MHz	
			Slew Rate		6		V/μs	
			Input Offset Voltage		2	5	mV	
			Input Offset Current		20	100	nA	
			Input Bias Current		250	500	nA	
			Output Source Current	Pin 7 or 10	2	10		mA
			Output Sink Current	Pin 7 or 10	400	900		μA
			Output Voltage Swing	Pin 7 or 10		11		V _{PP}
			Output Diode Leakage	Voltage on Pins 8 and 9 = 18V		0	10	μA

Note 1: For operation in ambient temperatures above 25°C, the device must be derated based on a 150°C maximum junction temperature and a thermal resistance of 175°C/Watt junction to ambient.

Note 2: Measured with an average responding voltmeter using the filter circuit in Figure 4. This simple filter is approximately equivalent to a "brick wall" filter with a passband of 20 Hz to 20 kHz (see "Application Hints" section). For 1 kHz THD the 400 Hz high pass filter on the distortion analyzer is used.

Note 3: The numbers are referred to an output level of 160 mV at Pins 7 and 10 using the circuit of Figure 2. This corresponds to an input level of 0.3 mV RMS at 333 Hz.

Note 4: Measured with an average responding voltmeter using the Dolby lab's standard CCIR filter having a unity gain reference at 2 kHz.

Note 5: Measured using the Rhode-Schwarz psophometer, model UPGR.

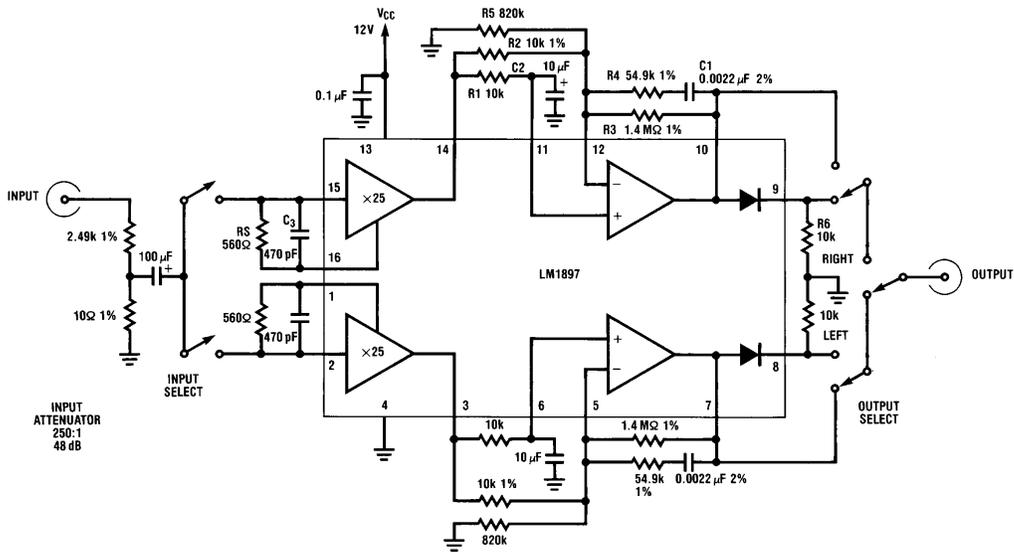


FIGURE 2. General Test Circuit

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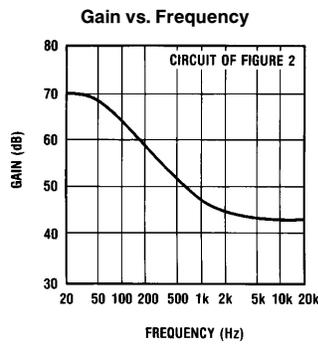


FIGURE 3. Frequency Response of Test Circuit

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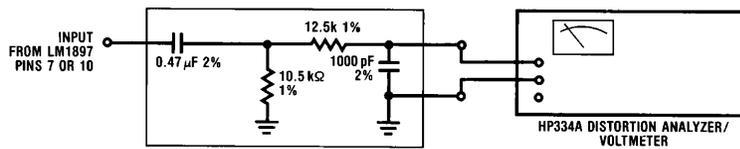
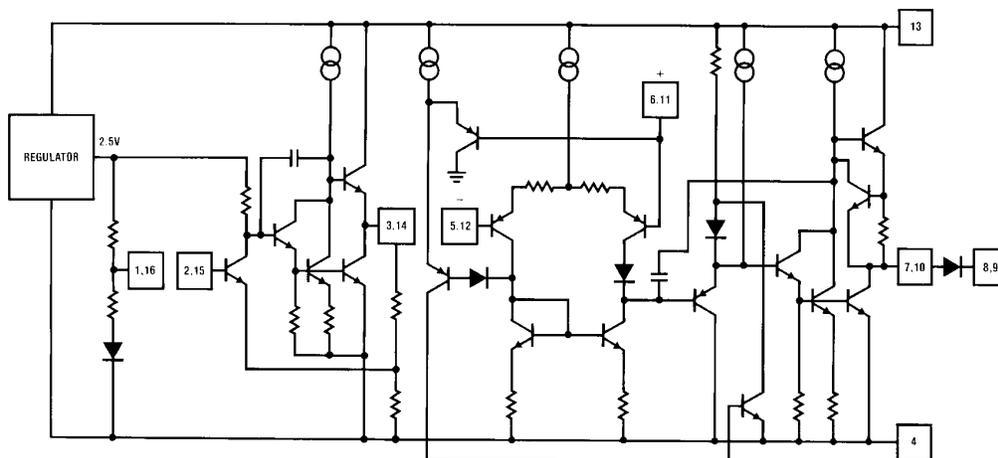


FIGURE 4. Simple 32 Hz-12740 Hz Filter and Meter

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FIGURE 5. Schematic Diagram

External Component	Component (Refer to Figure 1) External Component Function	Normal Range of Value	External Component	Component (Refer to Figure 1) External Component Function	Normal Range of Value
R ₁	Set turn-on delay and second amplifier's low frequency pole.	2 kΩ–40 kΩ	R ₆	Biases the output diode when it is used in DC switching applications. This resistor can be excluded if diode switching is not desired.	2 kΩ–47 kΩ
C ₂	Leakage current in C ₂ results in DC offset between the amplifier's inputs and therefore this current should be kept low. R ₁ is set equal to R ₂ such that any input offset voltage due to bias current is effectively cancelled. An input offset voltage is generated by the input offset current multiplied by the value of these resistors.	0.1 μF–10 μF (Low Leakage)	C ₃	Often used to resonate with tape head in order to compensate for tape playback losses including tape head gap and eddy current. For a typical cassette tape head, the resonant frequency selected is usually between 13 and 17 kHz.	100 pF–1000 pF
R ₂	Set the DC and low frequency gain of the output amplifier. The total input offset voltage will also be multiplied by the DC gain of this amplifier. It is therefore essential to keep the input offset voltage specification in mind when employing high DC gain in the output amplifier; i.e. 5 mV × 400 = 2V offset at the output.	2 kΩ–40 kΩ	R ₅	Increases the output DC bias voltage from the nominal 2.2V value (See the Application Hints section).	100 kΩ–10 MΩ
R ₃	Set tape playback equalization characteristics in conjunction with R ₃ (calculations for the component values are included in the Applications Hints section).	500 kΩ–10 MΩ	R ₇	Optionally used for tape muting. The use of this resistor can also provide "No Pop" turn-off if desired.	
R ₄		10 kΩ–200 kΩ			
C ₁		0.00047 μF–0.01 μF			

Application Hints

DISTORTION MEASUREMENT METHOD

In order to clearly interpret and compare specifications and measurements for low noise preamplifiers, it is necessary to understand several basic concepts of noise. An obvious example is the measurement of total harmonic distortion at very low input signal levels. Distortion analyzers provide outputs which allow viewing of the distortion products on an oscilloscope. The oscilloscope often reveals that the "distortion" being measured contains 1) distortion, 2) noise, and 3) 50 or 60 cycle AC line hum.

Application Hints (Continued)

Line hum can be detected by using the "line sync" on the oscilloscope (horizontal sync selector). The triggering of a constant wave form indicates that AC line pickup is present. This is usually the result of electro-magnetic coupling into the preamplifier's input or improper test equipment grounding, which simply must be eliminated before making further measurements!

Input coupling problems can usually be corrected by any one of the following solutions: 1) shielding the source of the magnetic field (using mu metal or steel), 2) magnetically shielding the preamplifier, 3) physically moving the preamplifier far enough away from the magnetic field, or 4) using a high pass filter ($f_0 = 200 \text{ Hz} - 1 \text{ kHz}$) at the output of the preamplifier to prevent any line signal from entering the distortion analyzer. Ground loop problems can be solved by rearranging ground connections of the circuit and test equipment.

Separating noise from distortion products is necessary when it is desired to find the actual distortion and not the signal-to-noise ratio of an amplifier. The distortion produced by the LM1897 is predominately a second harmonic. It is for this reason that the third and higher order harmonics can be filtered without resulting in any appreciable error in the measurement. The filter also reduces the amount of noise in the measured data. Another more tedious technique for measuring THD is to use a wave analyzer. Each harmonic is measured and then summed in an RMS calculation. A typical curve is plotted for distortion vs. frequency using this method. A typical curve is also included using a 20 Hz to 20 kHz 4th order filter.

To specify the distortion of the LM1897 accurately and also not require unusual or tedious measurements the following method is used. The output level is set to one volt RMS at 1 kHz (approximately 5 millivolts at the input). The output is filtered with the circuit of *Figure 4* to limit the bandwidth of the noise and measured with a standard distortion analyzer. The analyzer has a filter that is switched in to remove line hum and ground loop pick-up as well as unrelated low frequency noise. The resulting measurement is fast and accurate.

SIGNAL-TO-NOISE RATIO

In the measurement of the signal-to-noise ratio, misinterpretations of the numbers actually measured are common. One amplifier may sound much quieter than another, but due to improper testing techniques, they appear equal in measurements. This is often the case when comparing integrated circuit to discrete preamplifier designs. Discrete transistor preamps often "run out of gain" at high frequencies and therefore have small bandwidths to noise as indicated below.

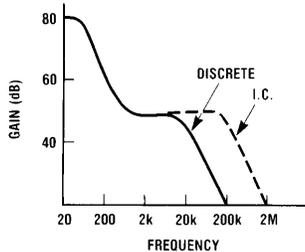


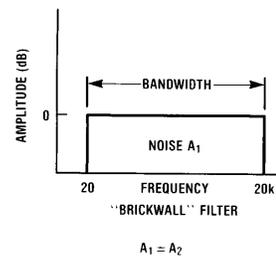
FIGURE 6

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Integrated circuits have additional open loop gain allowing additional feedback loop gain in order to lower harmonic distortion and improve frequency response. It is this additional bandwidth that can lead to erroneous signal to noise measurements if not considered during the measurement process. In the typical example above, the difference in bandwidth appears small on a log scale but the factor of 10 in bandwidth, (200 kHz to 2 MHz) can result in a 10 dB theoretical difference in the signal-to-noise ratio (white noise is proportional to the square root of the bandwidth in a system).

In comparing audio amplifiers it is necessary to measure the magnitude of noise in the audible bandwidth by using a "weighting" filter.¹ A "weighting" filter alters the frequency response in order to compensate for the average human ear's sensitivity to certain undesirable frequency spectra. The weighting filters at the same time provide the bandwidth limiting as discussed in the previous paragraph.

The 32 Hz to 12740 Hz filter shown in *Figure 4* is a simple two pole, one zero filter, approximately equivalent to a "brick wall" filter of 20 Hz to 20 kHz. This approximation is absolutely valid if the noise has a flat energy spectrum over the frequencies involved. In other words a measurement of a noise source with constant spectral density through either of the two filters would result in the same reading. The output frequency response of the two filters is shown in *Figure 7*.



$$A_1 = A_2$$

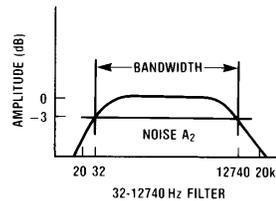


FIGURE 7

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Typical signal-to-noise figures are listed for several weighting filters which are commonly used in the measurement of noise. The shape of all weighting filters is similar, with the peak of the curve usually occurring in the 3 kHz-7 kHz region as shown below.

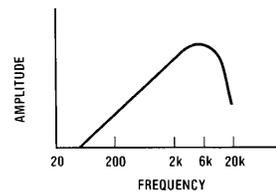


FIGURE 8

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Application Hints (Continued)

In addition to noise filtering, differing meter types give different noise readings. Meter responses include: 1) RMS reading, 2) average responding, 3) peak reading, and 4) quasi peak reading. Although theoretical noise analysis is derived using true RMS (root mean square) based calculations, most actual measurement is taken with ARM (Average Responding Meter) test equipment. Unless otherwise noted an average responding meter is used for all AC measurements in this data sheet.

BASIC CIRCUIT APPROACH

The LM1897 IC incorporates a two stage broadband design which minimizes noise, attains overall DC stability and prevents audible transients during turn-on.

The first stage is a direct coupled amplifier with an internal gain of 25 V/V (28 dB). Direct coupling to the tape head reduces input source impedance and external component cost by removing the input coupling capacitor. A typical input coupling capacitor of 1 μ F has a reactance of 1.5 k Ω at 100 Hz. The resulting noise due to the amplifier's input noise current can dominate the noise voltage at the output of the playback system. The input of the amplifier is biased from a reference voltage that is temperature compensated to produce a quiescent DC voltage of 2.2V at the output of the first stage. The input stage bias current that flows through the tape head is kept below 2 μ A in order to prevent any erasure of tape moving past the head. An added advantage of DC biasing is the prevention of large current transients during the charging of coupling capacitors at turn-on and turn-off.

The second stage provides additional gain and proper equalization while preventing audible turn-on transients or "pops". The output (Pin 10) is kept low until C2 charges through R1. When the voltage on C2 gets close to the DC voltage on Pin 14, the output rises exponentially to its final DC value. The result is a transient-free turn-on characteristic.

Internal diodes are provided to facilitate electronic diode switching popular in automotive applications.

The general test circuit illustrates the topography of the system. The components determining the overall frequency response are external due to the extreme sensitivity when matching a DIN equalization curve.

MUTE CIRCUIT

The LM1897 can be muted with the addition of two resistors and a grounding switch, as shown in *Figure 1*. When the circuit is not muted the additional resistors have no effect on the AC performance. They *do* have an effect on the DC Q point however.

The difference in the DC output voltages of the input amplifiers is applied across the mute resistors (R7) and the positive input resistors (R1). This results in an additional offset at the input of the output amplifiers. To keep this offset to a minimum R7 should be as large as possible to achieve effective muting. In all cases R7 should be at least ten times R1. A typical value of R7 is 25 to 50 times R1.

CAPACITOR-COUPLED INPUT

The LM1897 is intended to be coupled directly to the signal source. Direct coupling permits faster turn-on and less low-frequency noise than would be possible with a capacitor-coupled input. However, there are some applications which require that the signal source be referred to ground and coupled to the input through a capacitor. *Figure 9* is an example of an LM1897 with a capacitor-coupled input. As shown, the circuit has a flat frequency response and is suitable for use as a microphone preamp.

R₈ provides a DC path for input bias current. The value of R₈ should be as low as possible without loading the source. A very large value of R₈ can cause excessive DC offset at the amplifier output. In order to avoid turn-on pops, the inverting input of the second amplifier must be at a higher voltage than the non-inverting input when V_{CC} is applied. R₁₀, R₁₁, R₁₂, and D₁ ensure that this condition will be met. If later stages in the playback system employ turn-on muting circuitry, these extra components may not be needed. The value of R₁₀ depends on V_{CC} as defined by the following relationship:

$$R_{10} = (V_{CC} - 1) \times 1k$$

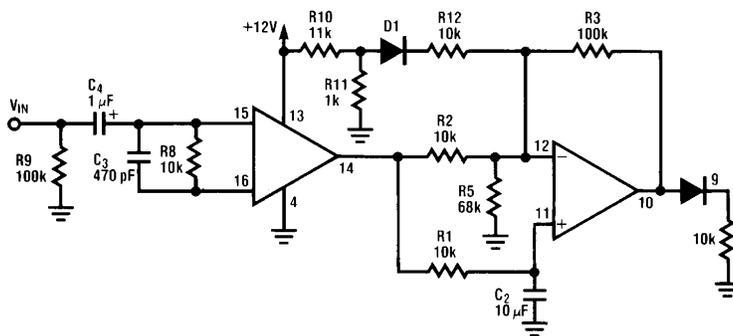


FIGURE 9. Microphone Preamplifier with Capacitor Coupled Input

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Application Hints (Continued)

Design Equation

The overall gain of the circuit is given by:

$$A_V = 25 \left[\frac{-R_4 R_3}{R_2(R_3 + R_4)} \right] \left(s + \frac{1}{R_4 C_1} \right) \left(s + \frac{1}{(R_3 + R_4) C_1} \right) \quad (1)$$

Standard cassette tapes require equalization of 3180 μ s (50 Hz) and 120 μ s (1.3kHz). These time constants result in an AC gain at 1 kHz given by:

$$A_V(1 \text{ kHz}) = 25 \left(\frac{-R_4 R_3}{R_2(R_3 + R_4)} \right) 1.663 \left\{ \begin{array}{l} 3180 \mu\text{s or } 50 \text{ Hz} \\ \text{and} \\ 120 \mu\text{s or } 1326 \text{ Hz} \end{array} \right\} \quad (2)$$

Using the pole and zero locations of the transfer function, the two other equations needed to solve for the component values are:

$$R_4 = \frac{1}{2\pi C_1(1326 \text{ Hz})} \quad (3)$$

$$R_3 = \frac{1}{2\pi C_1(50 \text{ Hz})} - \frac{1}{2\pi C_1(1326 \text{ Hz})} = \frac{1}{2\pi C_1(51.96)} \quad (4)$$

We can now solve for C_1 as a function of R_2 , or:

$$A_V(1 \text{ kHz}) = -25 \left\{ \frac{\left[\frac{1}{2\pi C_1(1326)} \right] \left[\frac{1}{2\pi C_1(51.96)} \right]}{\left[\frac{1}{R_2 2\pi C_1(50)} \right]} \right\} (1.663) \quad (5)$$

$$C_1 = \frac{-4.80 \times 10^{-3}}{R_2 [A_V(1 \text{ kHz})]} \quad (6)$$

When chromium dioxide tape is used, the defined time constants are 3180 μ s and 70 μ s. This changes equation (3) to:

$$R_4 = \frac{1}{2\pi C_1(2274 \text{ Hz})} \quad (7)$$

The value of R_3 is normally not changed. This results in an error of less than 0.2 dB in the low frequency response.

The output voltage of the LM1897 is set by the input amplifier DC voltage at pin 3 or 14, and by R_3 and R_5 .

$$\text{Nominal } V_{OUT} (\text{pin 7 or } 10) = 2.2 \left(1 + \frac{R_3}{R_5} \right) \quad (8)$$

Pins 8 and 9 are biased 0.7 volts less than V_{OUT} (pin 7 or 10). When these diodes are used the output (pin 7 or 10) should be biased at one half the minimum operating supply voltage. Equation (8) can be rewritten to solve for R_5 .

$$R_5 = \frac{2.2 R_3}{V_O - 2.2} \quad (9)$$

The output voltage of the LM1897 will vary from that given in equation (8) due to variations in the input amplifier DC voltage as well as the output amplifier input bias current, input offset current and input offset voltage. The following equation gives the worst case variation in the output voltage.

$$\Delta V_{OUT} = \pm \left[\Delta V_{PIN 3} \left(1 + \frac{R_3}{R_5} \right) + \frac{R_3}{R_5} \left(\Delta I_{BIAS} (R_1 - R_2) \right) + \frac{I_{OS}}{2} (R_1 + R_2) + V_{OS} \right] \quad (10)$$

Using the worst case values in the electrical characteristics reduces this to

$$\Delta V_{OUT} = \pm \left[0.4 \left(1 + \frac{R_3}{R_5} \right) + \frac{R_3}{R_2} \left(200 \text{ nA} (R_1 - R_2) + 50 \text{ nA} (R_1 + R_2) + 5 \text{ mV} \right) \right] \quad (11)$$

The turn-on delay is set by R_1 and C_2 ; delay can be approximated by:

$$\text{Delay Time } t = R_1 C_2 \ln \left(\frac{2.2}{V_{ODC}} \right) \left(\frac{R_3}{R_2} \right) \quad (12)$$

Example

If we desire a tape preamp with 100 mV output signal from a tape head with a nominal output of 0.5 mV at 1 kHz for standard ferric cassette tape, the external components are determined as follows. The value of R_2 is arbitrarily set to 10 k Ω .

$$R_1 = R_2 = 10 \text{ k}$$

This minimizes errors due to the output amplifier bias currents.

$$C_1 = \frac{-4.80 \times 10^{-3}}{10 \text{ k}\Omega \left[\frac{-100 \text{ mV}}{0.5 \text{ mV}} \right]} = 2400 \text{ pF} \rightarrow 0.0022 \mu\text{F}$$

Use 0.0022 μ F and determine:

$$R_4 = \frac{1}{2\pi C_1(1326)} = 54.6 \text{ k}\Omega \rightarrow 54.9 \text{ k}\Omega \text{ } 1\%$$

$$R_3 = \frac{1}{2\pi C_1(51.96)} = 1.39 \text{ M}\Omega \rightarrow 1.4 \text{ M}\Omega \text{ } 1\%$$

To bias the output amplifier output voltage at 6 volts (half supply):

$$R_5 = \frac{2.2(1.4 \text{ M}\Omega)}{6 - 2.2} = 811 \text{ k}\Omega \rightarrow 820 \text{ k}\Omega$$

The maximum variation in the output voltage is found using equation (11):

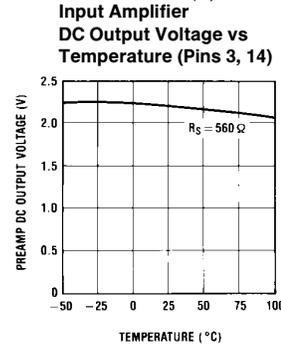
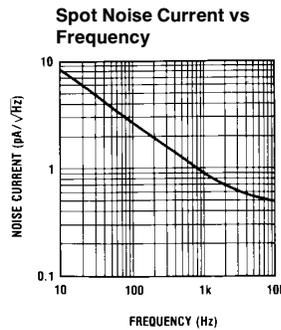
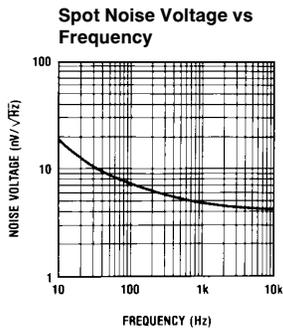
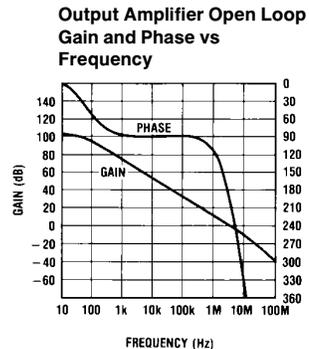
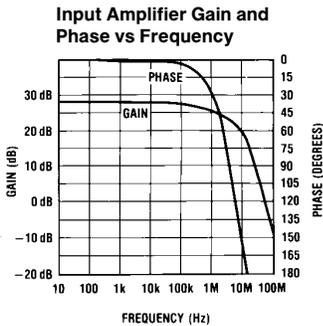
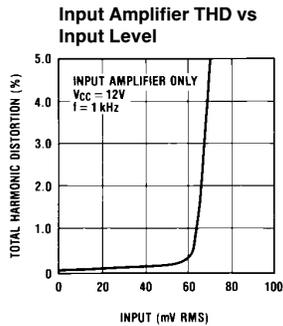
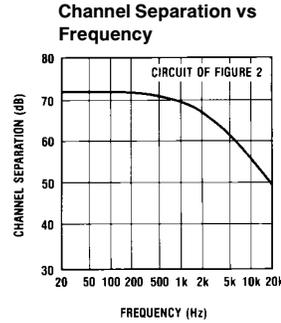
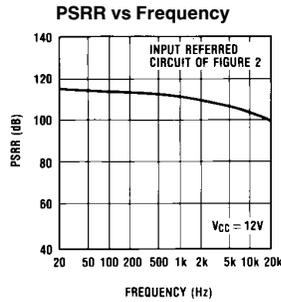
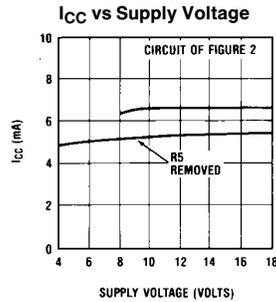
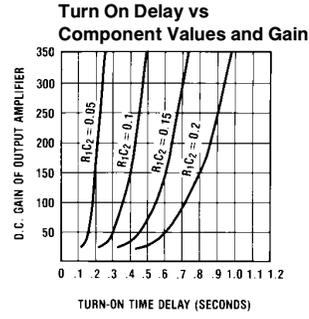
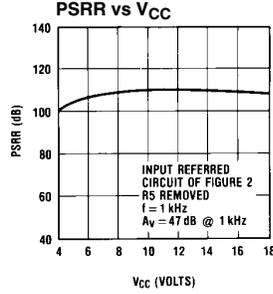
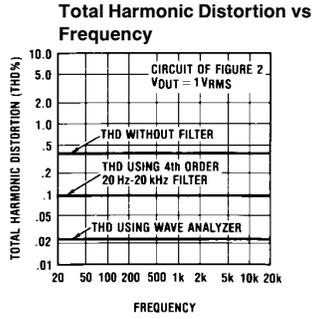
$$\Delta V_{OUT} = \pm 1.9 \text{ V}$$

The low frequency response and turn-on delay determine the value of C_2 . For $R_1 = 10 \text{ k}$ and $C_2 = 10 \mu\text{F}$ the low frequency 3 dB point is 1.6 Hz and the turn-on delay is 0.4 seconds, from equation (12).

The complete circuit is shown in Figure 2. A circuit with 5% components and biased for a minimum supply of 10 volts is shown in Figure 1. If additional gain is needed R_1 and R_2 can be reduced without changing the frequency response of the circuit.

Reference 1: CCIR/ARM: *A Practical Noise Measurement Method*; by Ray Dolby, David Robinson and Kenneth Gundry, AES Preprint No. 1353 (F-3).

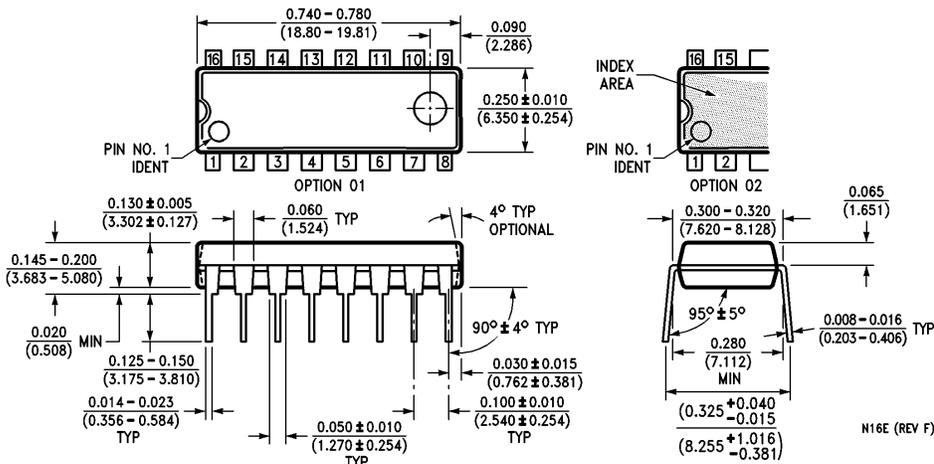
Typical Performance Characteristics



TL/H/7094-10



Physical Dimensions inches (millimeters)



Molded Dual-In-Line Package (N)
Order Number LM1897N
NS Package Number N16E

LIFE SUPPORT POLICY

NATIONAL'S PRODUCTS ARE NOT AUTHORIZED FOR USE AS CRITICAL COMPONENTS IN LIFE SUPPORT DEVICES OR SYSTEMS WITHOUT THE EXPRESS WRITTEN APPROVAL OF THE PRESIDENT OF NATIONAL SEMICONDUCTOR CORPORATION. As used herein:

1. Life support devices or systems are devices or systems which, (a) are intended for surgical implant into the body, or (b) support or sustain life, and whose failure to perform, when properly used in accordance with instructions for use provided in the labeling, can be reasonably expected to result in a significant injury to the user.
2. A critical component is any component of a life support device or system whose failure to perform can be reasonably expected to cause the failure of the life support device or system, or to affect its safety or effectiveness.



National Semiconductor Corporation
 1111 West Bardin Road
 Arlington, TX 76017
 Tel: 1(800) 272-9959
 Fax: 1(800) 737-7018

National Semiconductor Europe
 Fax: (+49) 0-180-530 85 86
 Email: cnjwge@tevm2.nsc.com
 Deutsch Tel: (+49) 0-180-530 85 85
 English Tel: (+49) 0-180-532 78 32
 Français Tel: (+49) 0-180-532 93 58
 Italiano Tel: (+49) 0-180-534 16 80

National Semiconductor Hong Kong Ltd.
 19th Floor, Straight Block,
 Ocean Centre, 5 Canton Rd.
 Tsimshatsui, Kowloon
 Hong Kong
 Tel: (852) 2737-1600
 Fax: (852) 2736-9960

National Semiconductor Japan Ltd.
 Tel: 81-043-299-2309
 Fax: 81-043-299-2408

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