

# A Non-Complementary Audio Noise Reduction System

National Semiconductor  
Application Note 386  
Martin Giles



AN-386

## INTRODUCTION

The popularity of companding or complementary noise reduction systems is self-evident. Nearly all medium to high quality cassette tape decks include either Dolby® B or Dolby C type noise reduction. A scant few have different systems such as dbx or Hi-Com. The universal appeal of companders to n.r. system designers is the amount of noise reduction they can offer, yet one of the major reasons the Dolby B system gained dominance in the consumer marketplace is because it offered only a limited degree of noise reduction — just 10 dB. This was sufficient to push cassette tape noise down to the level where it became acceptable in good-quality applications, yet wasn't enough that undecoded playback on machines not equipped with a Dolby B system was unsatisfactory — quite the contrary, in fact. The h.f. boost on Dolby B encoded tapes when reproduced on systems with modest speakers was frequently preferred. Since companding systems are so popular, it is not unreasonable to ask, "why do we need another noise reduction system?"

For many of the available audio sources today, companders are not a solution for audio noise. When the source material is not encoded in any way and has perceptible noise, complementary noise reduction is not possible. This includes radio and television broadcasts, the majority of video tapes and of course, older audio tape recordings and discs. The DNR™ single-ended n.r. system has been developed specifically to reduce noise in such sources. A single-ended system, able to provide noise reduction where non previously existed, and which avoids compatibility restraints or the imposition of yet another recording standard for consumer equipment, is therefore attractive.

The DNR system can be implemented by either of two integrated circuits, the LM1894 or the LM832, both of which can offer between 10 and 14 dB noise reduction in stereo pro-

gram material. Although differing in some details (the LM832 is designed for low-signal, low-supply voltage applications) the operation of the integrated circuits is essentially the same. Two basic principles are involved; that the noise output is proportional to the system bandwidth, and that the desired program material is capable of "masking" the noise when the signal-to-noise ratio is sufficiently high. DNR automatically and continuously changes the system bandwidth in response to the amplitude and frequency content of the program. Restricting the signal bandwidth to less than 1 kHz reduces the audible noise and a special spectral weighting filter in the control path ensures that the audio bandwidth in the signal path is always increased sufficiently to pass any music that may be present. Because of this ability to dynamically analyze the auditory masking qualities of the program material, DNR does not require the source to be encoded in any special way for noise reduction to be obtained. This paper deals with the design and operating characteristics of the LM1894. For a more complete description of the principles behind the DNR system, refer to AN384.

## THE DNR SYSTEM FORMAT

A block diagram showing the basic format of the LM1894 is shown in *Figure 1*. This is a stereo system with the left and right channel audio signals each being processed by a controlled cut-off frequency ( $f_{-3\text{ dB}}$ ) low-pass filter. The filter cut-off frequency can be continuously and automatically adjusted between 800 Hz and 35 kHz by a signal developed in the control path. Both audio inputs contribute to the control path signal and are used to activate a peak detector which, in turn, changes the audio filters' cut-off frequency. The audio path filters are controlled by the same signal for equally matched bandwidths in order to maintain a stable stereo image.

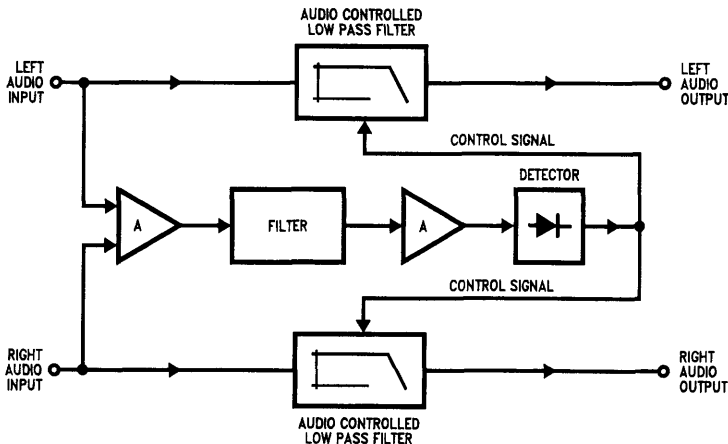
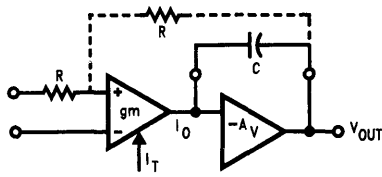


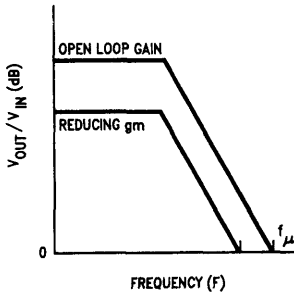
FIGURE 1. Stereo Noise Reduction System (DNR)

TL/H/8395-1



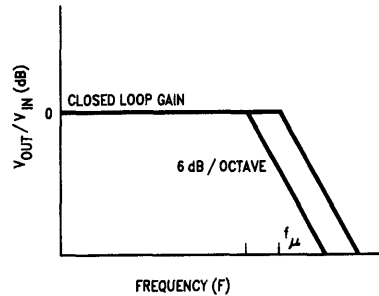
TL/H/8395-2

(a) Variable Lowpass Filter



TL/H/8395-13

(b) Open Loop Response



TL/H/8395-14

(c) Closed Loop Response

FIGURE 2

### VARIABLE CUT-OFF LOW DISTORTION FILTERS

By low distortion we mean a filter that has a flat response below the cut-off frequency, a smooth, constant attenuation slope above the cut-off frequency and does not peak at the cut-off frequency as this frequency is changed.

The circuit topology is shown in *Figure 2 (a)* and is, in fact, very similar to the pole-splitting frequency compensation technique used on many integrated circuit operational amplifiers (see pp. 24-26 of "Intuitive I/C Op Amps" by T. M. Fredericksen). A variable transconductance ( $g_m$ ) stage drives an amplifier configured as an integrator. The transconductance stage output current  $I_O$  is given by

$$I_O = g_m V_{in} \quad (1)$$

and if the second amplifier is considered ideal, then the voltage  $V_{out}$  is the result of  $I_O$  flowing through the capacitive reactance of  $C$ . Therefore we can write

$$V_{out} = \frac{I_O}{2\pi fC} \quad (2)$$

Combining (1) and (2) we have

$$\frac{V_{out}}{V_{in}} = \frac{g_m}{2\pi fC} \quad (3)$$

At some frequency, the open loop gain will fall to unity ( $f = f_u$ ) given by

$$f_u = \frac{g_m}{2\pi C} \quad (4)$$

For a fixed value of capacitance, when the transconductance changes, then the unity gain frequency will change correspondingly as shown in *Figure 2 (b)*.

If we put dc feedback around both stages for unity closed loop gain, the amplitude response will be flat (or unity gain) until  $f_u$  is reached, and then will follow the open loop gain curve which is falling at 6 dB/octave. Since we control  $g_m$ , we can make  $f_u$  any frequency we desire and therefore have a controlled cut-off frequency low pass filter.

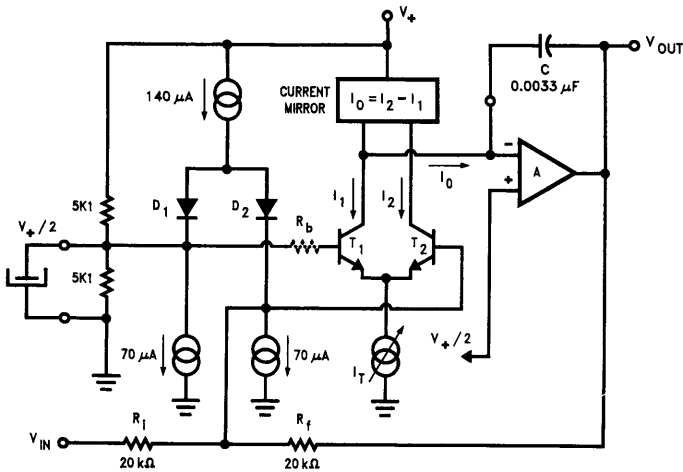
A more detailed schematic is given in *Figure 3* and shows the resistors  $R_f$  and  $R_i$  which provide dc feedback around the circuit for unity closed-loop gain (i.e. at frequencies below  $f_u$ ). The transconductance stage consists of a differential pair  $T_1$  and  $T_2$  with current mirrors replacing the more conventional load resistors. The output current  $I_O$  to the integrator stage is the difference between  $T_1$  and  $T_2$  collector currents.

For a differential pair, as long as the input differential voltage is small — a few millivolts — the  $g_m$  is dependent on the tail current  $I_T$  and can be written

$$g_m = \frac{q}{kT} \times \frac{I_T}{2} \quad (5)$$

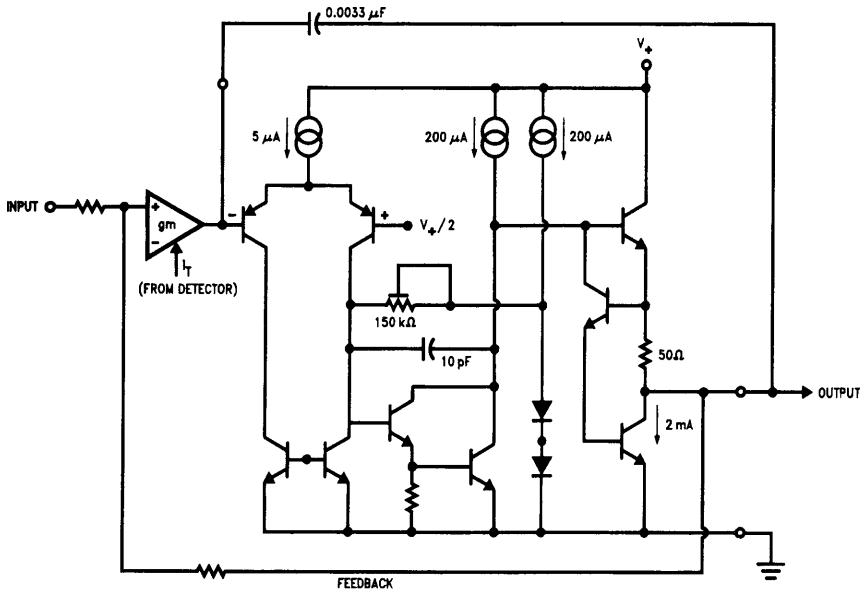
$$\text{where } \frac{q}{kT} = \frac{1}{26 \text{ mV}} @25^\circ\text{C}$$

For frequencies below the cut-off frequency, the amplifier is operating closed loop, and the dc feedback via  $R_f$  will keep the input differential voltage very small. However, as the input signal frequency approaches cut-off, the loop gain decreases and larger differential voltages will start to appear across the bases of  $T_1$  and  $T_2$ . When this happens, the  $g_m$  is no longer linearly dependent on the tail current  $I_T$  and signal distortion will occur. To prevent this, two diodes  $D_1$  and  $D_2$  biased by current sources are added to the input stage. Now the signal current is converted to a logarithmically related voltage at the input to the differential pair  $T_1$  and  $T_2$ . Since the diodes and the transistors have identical geometries and temperature excursions, this conversion will exactly compensate for the exponential relationship between the input voltage to  $T_1$  and  $T_2$  and the output collector currents. As long as the signal current is less than the current available to the diodes, the transconductance amplifier will have a linear characteristic with very low distortion.



TL/H/8395-3

FIGURE 3. Variable Lowpass Filter with Distortion Correcting Diodes and Control Voltage Offset Compensation



TL/H/8395-4

FIGURE 4. The OP AMP Output Stage of the LM1894

For the entire circuit, if  $R_1 = R_f = R$  and the diode dynamic resistance is  $r_d$ , we can write the transfer characteristic as

$$\frac{V_{out}}{V_{in}} = \frac{-1}{\left(1 + \frac{4\pi fCK 26 \times 10^{-3}}{I_T}\right)} \quad (6)$$

$$\text{where } K = \left(2 + \frac{R}{2r_d}\right)$$

Therefore the pole frequency for  $C = 0.0033 \mu F$  is

$$f_u = I_T / 4 \pi 26 \times 10^{-3} CK = I_T \times 33.2 \times 10^6$$

$$\text{for } f_u = 1 \text{ kHz, } I_T = 33.2 \mu A$$

$$\text{for } f_u = 35 \text{ kHz, } I_T = 1.1 \text{ mA}$$

In operation, the transconductance stage current  $I_T$  for the LM1894 will vary between the levels given above in response to the control path detected voltage. Notice that with the circuit values given in Figure 3 the maximum output voltage swing at the cut-off frequency is about  $1V_{rms}$  (use equation 2 and put  $I_O = I_T = 33 \mu A$ ) and this is specified in the LM1894 data sheet as the input voltage for 3% THD. This is, of course, the condition for minimum bandwidth when noise only is normally present at the input. When signals are simultaneously present causing the audio bandwidth to increase out to 35 kHz, the transconductance stage current is over 1 mA, allowing signal swings at 1 kHz (theoretically) of over  $34 V_{rms}$ . Practically, at maximum bandwidth the output swing is determined by the output stage saturation voltages which are dependent on the supply volt-

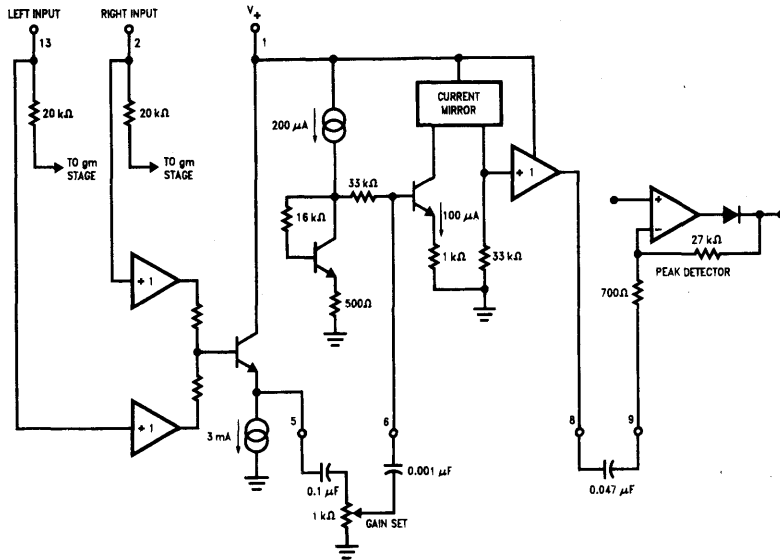


FIGURE 5. Control Path Amplifiers and Filters

TL/H/8395-5

age (see Figure 4). With a 15 V<sub>DC</sub> supply, the LM1894 can handle well over 4 V<sub>rms</sub>.

While there are other circuit topologies that can be used to obtain a variable cut-off low pass filter, this design has certain advantages, especially when it comes to avoiding control feedthrough. Control feedthrough is the name given to voltage offsets that can occur in the audio path as the transconductance stage current changes. The audible effect is a low level "bacon frying" noise or pops as the bandwidth changes. To prevent such voltage offsets occurring, the differential stage T<sub>1</sub> and T<sub>2</sub>, the current mirrors and the diodes are arranged to provide good tracking over the entire range of the bandwidth control current I<sub>T</sub>. Because the transconductance stage is driving the inverting input to an operational amplifier — a virtual ground — there will be no voltage swing at this node. This eliminates possible offset voltages from output impedance changes in the current mirror and T<sub>1</sub> collector caused by different operating currents. Last, but not least, a source of offset voltages are the base currents of T<sub>1</sub> and T<sub>2</sub>. Because the transistors have a finite current gain, when the tail current I<sub>T</sub> is increased, these base currents must increase slightly. T<sub>1</sub> base current is provided by the reference voltage (V<sub>+</sub>/2), but T<sub>2</sub> base current must come via the feedback resistor R<sub>f</sub>. This current is not normally available from D<sub>2</sub> because the feedback loop is holding T<sub>1</sub> and T<sub>2</sub> base voltages equal. By adding the resistor R<sub>b</sub> in series with T<sub>1</sub> base, a compensating offset voltage is produced across the input diodes. This reduces the current in D<sub>1</sub> slightly and increases the current in D<sub>2</sub> correspondingly, allowing it to supply the increased base current requirement of T<sub>2</sub>.

#### THE CONTROL PATH

The purpose of the control path is to ensure that the audio bandwidth is always sufficiently wide to pass the desired signal, yet in the absence of this signal will decrease rapidly enough that the noise also present does not become audible. In order to do this, the control path must recognize the masking qualities of the signal source and the detector stage must be able to take advantage of the characteristics

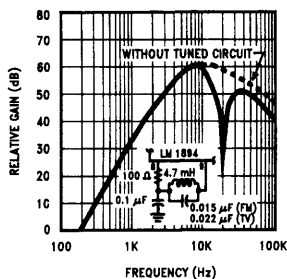
of the human ear so that audible signal distortion or unmasking does not occur.

Figure 5 shows a block diagram of the control path including the external components. A straight-forward summing amplifier combines the left and right channel inputs and acts as a buffer amplifier for the gain control. Because the noise level for signal sources can be different — cassette tapes are between -50 dB and -65 dB (depending on whether Dolby B encoding is employed) and FM broadcast noise is around -45 dB to over -75 dB (depending on signal strength) — the control path gain is adjusted such that a noise input is capable of just increasing the audio bandwidth from its minimum value. This ensures that any program material above the noise level increases the audio bandwidth so that the material is passed without distortion. Setting the potentiometer (or an equivalent pair of resistors) will be described in more detail later.

The gain control potentiometer is also part of the DNR filter characteristic derived from auditory masking considerations — see AN384. Combined with a 0.1 μF coupling capacitor, the total resistance of the potentiometer will cause a signal attenuation below 1.6 kHz.

$$\text{i.e. } f_1 = \frac{1}{2\pi RC} = \frac{1}{2\pi \times 10^3 \times 0.1 \times 10^{-6}} = 1.6 \text{ kHz}$$

This helps to prevent signals with a high amplitude but no high frequency content above 1 kHz — such as a bass drum — from activating the control path detector and unnecessarily opening the audio bandwidth. For signals that do have a significant high frequency content (predominantly harmonics), the control path sensitivity is increased at a 12 dB/octave rate. This rapid gain in sensitivity is important since the harmonic content of program material typically falls off quickly with increasing frequency. The 12 dB/octave slope is provided by cascading two RC high pass filters composed of the coupling capacitors to the control path gain stage and detector stage and the internal input resistors to these stages. Individual corner frequencies of 5.3 kHz and 4.8 kHz respectively are used, with a combined corner frequency around 6 kHz. Above 6 kHz the gain can be allowed to



TL/H/8395-6

FIGURE 6. Control Path Frequency Response

decrease again since the signal energy content between 1 kHz and 6 kHz (the critical masking frequency range) will have already caused the audio bandwidth to extend beyond 30 kHz, allowing passage of any high frequency components in the audio path.

Under some circumstances, not normal to music or speech, the source can contain relatively high level, high frequency components which are not necessarily accompanied by large levels of low frequency signal energy providing noise masking. These are spurious components such as the line scan frequency in a television receiver (15.734 kHz) or sub-carrier signals such as the 19 kHz pilot tone in FM stereo broadcasting. Although both these components should be low enough to be inaudible in the audio path, their presence in the control path could cause a change in the minimum bandwidth and hence the amount of available noise reduction. Since these unwanted components are at frequencies higher than the desired control path frequency range, they are easily accommodated by including a notch filter in the control path at the specified frequency. A resonant L-C circuit with a Q of 30 will attenuate 19 kHz by over 28 dB. If a 10% tolerance 0.015  $\mu$ F capacitor is used, the coil can be a fixed 4.7 mH inductance. For 15.734 kHz a 0.022  $\mu$ F capacitor is needed. When those frequency components are not present (i.e. in cassette tapes) the L-C circuit is eliminated and the gain amplifier and detector stage are coupled together with a single 0.047  $\mu$ F capacitor.

Apart from providing the proper frequency response the control path gain must be enough to ensure that the detector threshold can be reached by very low noise input levels. The summing amplifier has unity gain to the sum of the left and right channel inputs and the necessary signal gain of 60 dB is split between the following gain amplifier and the detector stage. For the gain amplifier

$$A_V = 33 \times 10^3 / (r_e + 10^3) = 26.2 \\ = 28.4 \text{ dB}$$

For the detector stage, the gain to negative signal swings is

$$A_V = 27 \times 10^3 / 700 = 38.6 = 31.7 \text{ dB}$$

With over 60 dB gain and typical source input noise levels, the gain potentiometer will normally be set with the wiper arm close to the ground terminal.

#### THE DETECTOR STAGE

The last part of the LM1894 to be described is the detector stage which includes a negative peak detector and a voltage to current converter. As noted earlier, the input resistance of the detector, together with the input coupling capacitor, forms part of the control path filter. Similarly the output resistance from the detector and the gain setting feedback resistor help to determine the detector time constants. With

a pulse or transient input signal, the rise time is 200  $\mu$ s to 90% of the final detected voltage level. Actual rise-times will normally be longer with the detector tracking the envelope of the combined left and right channel signals after they have passed through the control path filter.

An interesting difference to compandor performance can be demonstrated with a 10 kHz tone burst. Since the LM1894 detector responds only to negative signal peaks, it will take about four input cycles to reach 90% of the final voltage on the detector capacitor (this is the 500  $\mu$ s time constant called out in the data sheet). After the first two cycles the audio bandwidth will have already increased past 10 kHz and a comparison of the input and output tone bursts will show only a *slight* loss in amplitude in these initial cycles. A compandor, however, usually cannot afford a fast detector time constant since the rapid changes in system gain that occur when a transient signal is processed can easily cause modulation products to be developed which may not be treated complementarily on playback. Therefore there is a time lag before the system can change gain, which may be to the maximum signal compression (as much as 30 dB depending on the compandor type). Failure to compress immediately at the start of the tone burst means that an *overshoot* is present in the signal which can be up to 30 dB higher than the final amplitude. To prevent this overshoot from causing subsequent amplifier overload (which can last for several times the period of the overshoot), clippers are required in the signal path, limiting the dynamic range of the system. Obviously, the LM1894 does not need clippers since no signal overshoots in the audio path are possible.

When the input signal transient decays, the diode in the detector stage is back biased and the capacitor discharges primarily through the feedback resistor and takes about 60 ms to reach 90% of the final value.

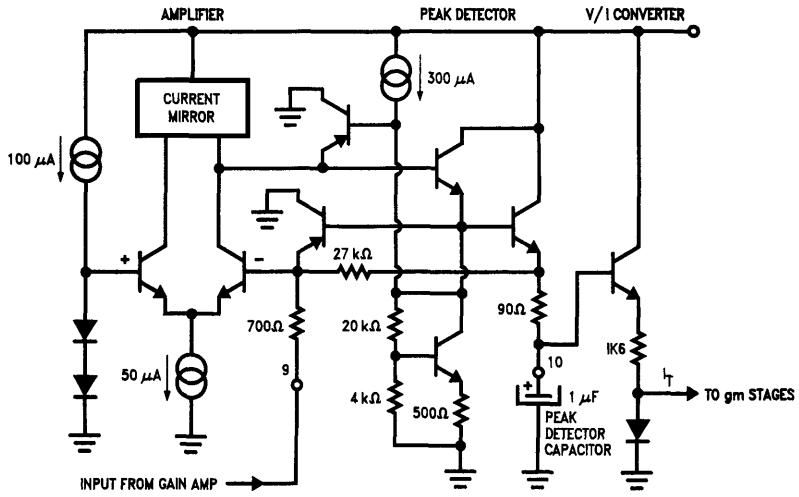
$$T = RC \times 2.3 = 27 \times 10^3 \times 1 \times 10^{-6} \times 2.3 \\ = 62.1 \text{ ms}$$

The decay time constant is required to protect the reverberatory or "ambience" qualities of the music. For material with a limited high frequency content or a particularly poor S/N ratio, some benefit can be obtained with a faster decay time—a resistor shunted across the detector capacitor will do this. Resistors less than 27 k $\Omega$  should not be used since very fast decay times will permit the detector to start tracking the signal frequency. For signal amplitudes that are not producing the full audio bandwidth, this will cause a rapid and audible modulation of the audio bandwidth.

#### BYPASSING THE SYSTEM

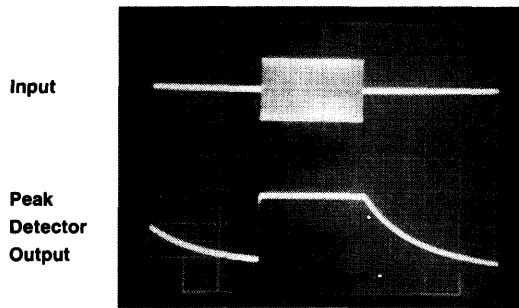
Sometimes it is necessary or desirable to bypass the n.r. system. This will allow a direct and instantaneous comparison of the effect that the system is having on the program material and will assist in arriving at the correct setting for the control path gain potentiometer. This facility is not practical with compandors unless unencoded passages occur in the program material. Also, should the action of the compandor become more objectionable than the noise in the original material, there is no way of switching the n.r. system off.

One way of bypassing is to simply use a double pole switch to route the signals around the LM1894. This physically ensures complete bypassing but does present a couple of problems. First, there may be a level change caused by the different impedances presented to the following audio stages when switching occurs. Second, the signal now has to be routed to the front panel where the switch is located, perhaps calling for shielded cable.



TL/H/8395-7

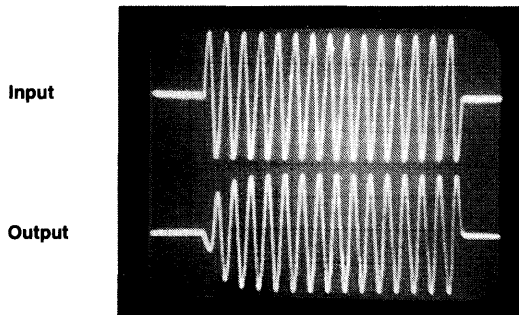
FIGURE 7. Peak Detector and Voltage to Current Converter



TL/H/8395-8

TIME: 20 ms/DIV

Peak Detector Response, 500 mV/Div



TL/H/8395-12

TIME: 0.5 ms/DIV

Audio Output Response, 10 kHz Tone Burst

FIGURE 8

A different technique, which avoids these problems, is to switch the LM1894 permanently into the full audio bandwidth mode. Since this provides a high S/N ratio path and low distortion the impact on the signal is minimal. Two methods can be used to switch the LM1894 audio bandwidth fully open, both with a single pole switch that is not in the audio path. Simply grounding the input of the peak detector amplifier will generate the maximum bandwidth control current and simultaneously prevent any control signals reaching the detector. Usually this is more than adequate since the maximum audio bandwidth is 34 kHz, but in some cases the 1 dB loss at 17 kHz produced by the single pole audio filters may not be desired. *Figure 9* shows a way to increase the audio bandwidth to 50 kHz (-1 dB at 25 kHz) by pulling up the detector capacitor to the reference voltage level ( $V_{+}/2$ ) through a 1 k $\Omega$  resistor. This method is useful only for higher supply voltage applications. To increase the bandwidth significantly the detector capacitor must be pulled up to around 5V ( $V_{+} > 10V$ ). Although a separate voltage source other than the reference pin could be used when  $V_{+}$  is less than 10V, this can cause an internal circuit latch-up if the voltage on the detector increases faster than the reference voltage at initial turn-on.

#### GENERAL SYSTEM MEASUREMENTS AND PRECAUTIONS

For most applications the external components shown in *Figure 9* will be required. In fact, the only recommended deviation from these values is the substitution of an equivalent pair of fixed resistors for the gain setting potentiometer. Location of the LM1894 in the audio path is important and should be prior to any tone or volume controls. In tape systems, right after the playback head pre-amplifier is the best place, or at the stereo decoder output (after de-emphasis and the multiplex filter) in an FM broadcast receiver. The LM1894 is designed for a nominal input level of 300 mVrms and sources with a much lower pre-amplifier output level will either require an additional gain block or substitution of the LM832 which is designed for 30 mVrms input levels.

The same circuit as *Figure 9* can be used for measurements on the I/C performance but, as with any other n.r. system, care in interpretation of the results may be necessary. For example, while the decay time constant for a tone burst signal is pretty constant, the attack time will depend on the tone frequency.

Sometimes separation of the audio path input and the control path is required, particularly when the frequency response or the THD with low input signal levels is being measured. If the audio and control paths are not separated then a typical audio system measurement of the frequency response will not appear as expected. This is because the control path frequency response is non-linear, exhibiting low sensitivity at low frequencies. When a low level input signal is swept through the audio frequency range, at low frequencies the audio -3 dB bandwidth will be held at 1 kHz, and the audio path signal will fall in amplitude as the signal goes above 1 kHz. As the signal frequency gets yet higher, the increasing sensitivity of the control path will allow the detector to be activated and the audio path -3 dB frequency starts to overtake the signal frequency. This causes the output signal amplitude to increase again giving the appearance that there is a dip in the audio frequency response around 1-2 kHz. It is worth remembering at this point that the audio path frequency response is always flat below some corner frequency and rolls off at 6 dB/octave above this frequency. In normal operation this corner frequency is the result of the aggregate control path signals in the 1 kHz to 6 kHz region and not the result of a single input frequency. To properly measure the frequency response of the audio path at a particular signal input frequency and amplitude, the control path input is separated by disconnecting  $C_5$  from Pin 5 and injecting the signal through  $C_5$  only. Then, a separate swept frequency response measurement can be made in the audio path. Similarly measurements of THD should include separation of the audio and control path inputs.

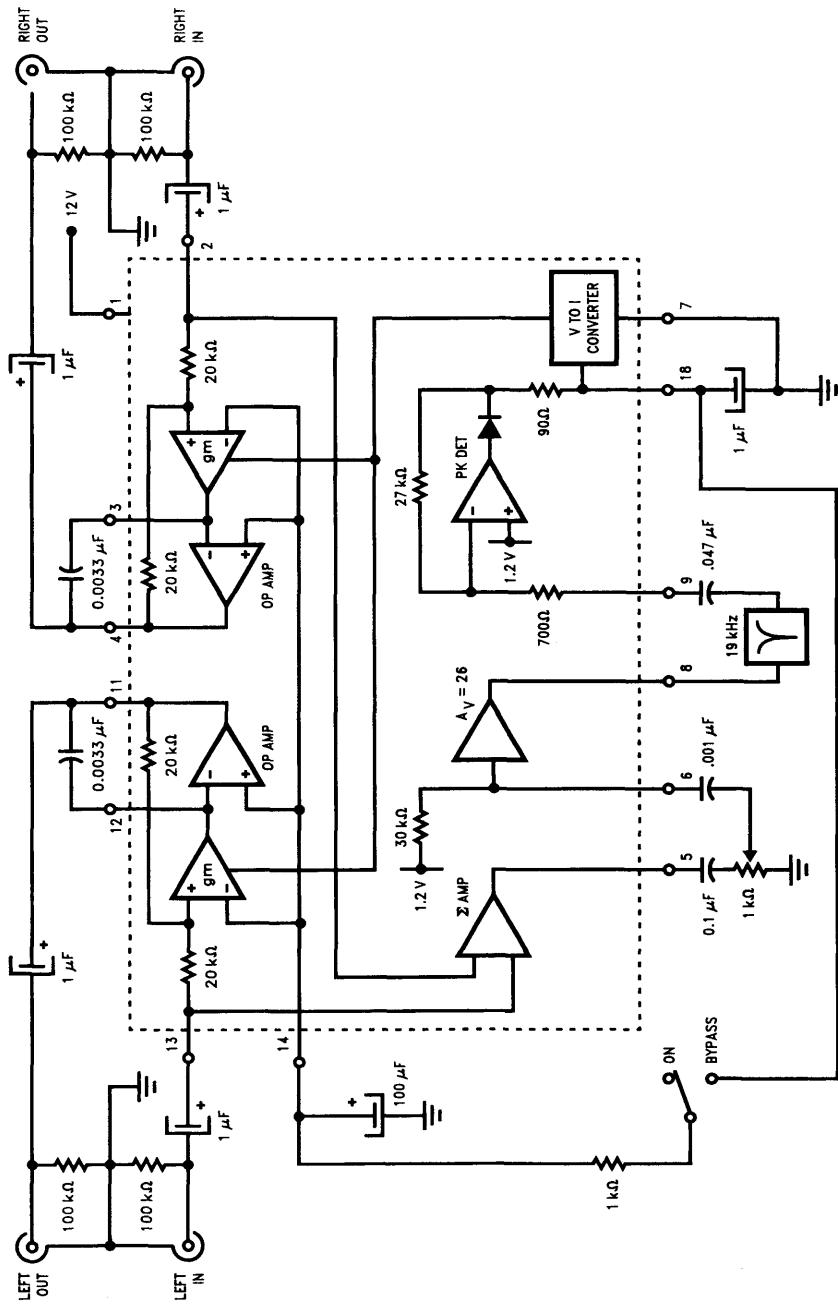


FIGURE 9. Complete Stereo Noise Reduction System



## PITFALLS - OR WHAT TO LISTEN FOR

Many people are understandably wary of non-complementary n.r. systems since there is no perfect means for distinguishing between the desired signal and noise. A thorough understanding of the psycho-acoustic basis for noise masking will go a long way to allaying these fears, but a much simpler method is to listen to a variety of source material with a DNR system being switched in and out. Even so, improper implementation of the LM1894—wrong location in the audio path changing either the level or frequency response of the source—or incorrect external component values, or the wrong sensitivity setting, can all strongly affect the audio in an undesired way. Sometimes, unhappily, the source is really beyond repair and some compromise must be made. Phonograph discs with bad scratches may require special treatment (a click and pop remover) and some older tape recordings may show some or all of the following problems.

### 1) Pumping:

Incorrect selection of the control path bandwidth external components can result in an audible increase in noise as the input level changes. This is most likely to be heard on solo instruments or on speech. Sometimes the S/N rate is too poor and masking will not be completely effective - i.e., when the bandwidth is wide enough to pass the program material, the increase in noise is audible. Cutting down on the pumping will also affect the program material to some extent and judgement as to which is preferable is required. Sometimes a shorter decay time constant in the detector circuit will help, especially for a source which always shows these characteristics, but for better program material a return to the recommended detector characteristics is imperative.

### 2) High Frequency Loss:

This can be caused by an improper control path gain setting—perhaps deliberate because of the source S/N ratio as described above—or incorrect values for the audio path filter capacitors. Capacitors larger than the recommended values will scale the operating bandwidth lower, causing lower  $-3$  dB corner frequencies for a given control path signal. Return to the correct capacitor values and the appropriate control path gain setting will *always* ensure that the h.f. content of the signal source is preserved.

### 3) Apparent High Frequency Loss:

The ability to instantaneously A/B the source with and without noise reduction can sometimes exhibit an apparent loss of h.f. signal content as the DNR system operates. This is most likely to happen with sources having an S/N ratio of less than 45 dB and is a subjective effect in that the program material probably does not have any significant h.f. components. It has been reported several times elsewhere that adding high frequency noise (hiss) to a music signal with a limited frequency range will seem to add to the h.f. content of the music. Trying sources with a higher S/N ratio that do not demonstrate this effect can re-assure the listener that the DNR system is operating properly. Alternatively a control path sensitivity can be used that leaves the audio bandwidth slightly wider, preserving the "h.f. content" at the expense of less noise reduction in the absence of music.

### 4) Sensitivity Setting:

Since this is the only adjustment in the system, it is the one most likely to cause problems. Improper settings can cause any of the previously described problems. Factory pre-sets can (and are) used, but only when the source is well defined with known noise level. For the user who intends to noise reduce a variety of sources, the control path gain potentiometer is required and should be adjusted for each application. A bypass switch is helpful in this respect since it allows rapid A/B comparison. Another useful aid is a bandwidth indicator, shown in *Figure 10*. This is simply an LED display driver, the LM3915, operating from the voltage on the detector filter capacitor at Pin 10 of the LM1894. The LM3915 will light successive LEDs for each 3 dB increase in voltage. The resistor values are chosen such that the capacitor voltage when the LM1894 is at minimum audio bandwidth, is just able to light the first LED, and a full audio bandwidth control signal will light the upper LED. Experience will show that adjusting the sensitivity so that the noise in the source (no signal is present) is just able to light the second LED, will produce good results. This display also provides constant reassurance that the system audio bandwidth really is adequate to process the music. A simpler detector, using a dual comparator and a couple of LEDs can be constructed instead, with threshold levels selected to show the correct sensitivity setting, minimum bandwidth, maximum bandwidth or some intermediate bandwidth as desired.

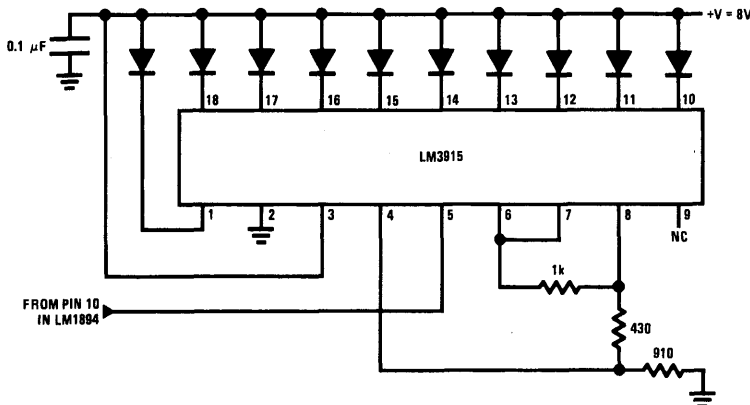


FIGURE 10. Bar Graph Display of Peak Detector Voltage

TL/H/8395-10

DNR Component Diagram

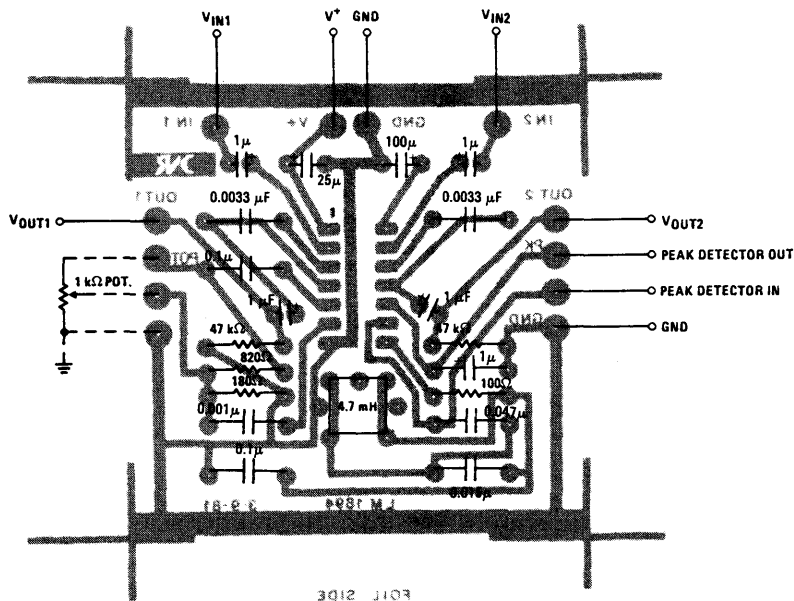


FIGURE 11. Printed Circuit Layout

TL/H/8395-11

# DNR™ Applications of the LM1894

National Semiconductor  
Application Note 390  
Martin Giles  
Kery Lacanette



## INTRODUCTION

The operating principles of a single-ended or non-complementary audio noise reduction system, DNR, have been covered extensively in a previous application note AN384, Audio Noise Reduction and Masking. Although the system was originally implemented with transconductance amplifiers (LM13600) and audio op-amps (LM387), dedicated I/Cs have since been developed to perform the DNR function. The LM1894 is designed to accommodate and noise reduce the line level signals encountered in video recorders, audio tape recorders, radio and television broadcast receivers, and automobile radio/cassette receivers. A companion device, the LM832, is designed to handle the lower signal levels available in low voltage portable audio equipment. This note deals chiefly with the practical aspects of using the LM1894, but the information given can also be applied to the LM832.

## THE BASIC DNR APPLICATION CIRCUIT

At the time of writing, the LM1894 has already found use in a large variety of applications. These include:

- AUTOMOTIVE RADIOS
- TELEVISION RECEIVERS
- HOME MUSIC CENTERS
- PORTABLE STEREOS (BOOM BOXES)
- SATELLITE RECEIVERS
- AUDIO CASSETTE PLAYERS
- AVIONIC ENTERTAINMENT SYSTEMS
- HI-FI AUDIO ACCESSORIES
- BACKGROUND MUSIC SYSTEMS
- ETC.

In the majority of these applications the circuit used is identical to that shown in *Figure 1*, and this is the basic stereo Dynamic Noise Reduction System. Although a split power supply can be used, a single positive supply voltage is shown, with ac coupled inputs and outputs common in many consumer applications. This supply voltage can be between 4.5 V<sub>DC</sub> and 18 V<sub>DC</sub> but operation at the higher end of the range (above 8 V<sub>DC</sub>) is preferred, since this will ensure adequate signal handling capability. The LM1894 is optimized for a nominal input signal level of 300 mVrms but with an 8 V<sub>DC</sub> supply it can handle over 2.5 Vrms at full audio bandwidth. Smaller nominal signal levels can be processed but below 100 mVrms there may not be sufficient gain in the control path to activate the detector with the source noise. In this instance, and where battery powered operation is desired, the LM832 is a better choice. The LM832 has identical operating principles and a similar (but not identical) pin-out. It is optimized for input levels around 30 mVrms and a supply voltage range from 1.5 V<sub>DC</sub> to 9.0 V<sub>DC</sub>.

The capacitors connected at Pins 12 and 3 determine the range of -3 dB cut off frequencies for the audio path filters. Increasing the capacitor value scales the range downward - the minimum frequency becomes lower and the maximum or full bandwidth frequency will decrease proportionally. Similarly, smaller capacitors will raise the range.

$$f_{-3\text{ dB}} = I_T / 9.1C \quad (I_T = 33 \mu\text{A MIN}) \quad (1)$$

$$(\quad = 1.05 \text{ mA MAX})$$

For normal audio applications the recommended value of 0.0033 μF should be adhered to, producing a frequency range from 1 kHz to 35 kHz.

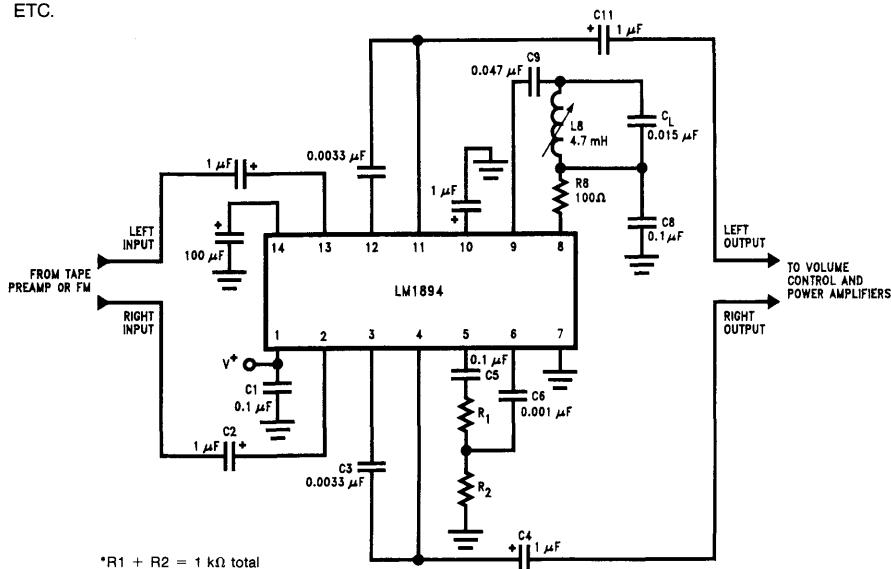
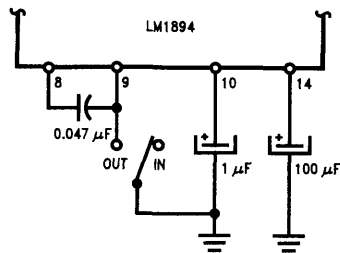


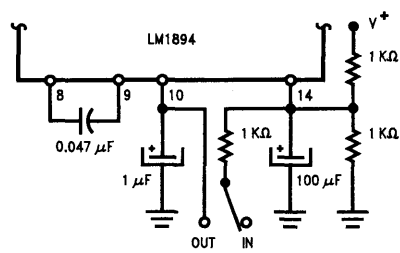
FIGURE 1. Complete DNR Application Circuit

TL/H/8420-1



TL/H/8420-2

(a)



TL/H/8420-18

(b)

FIGURE 2. Two Methods of DNR IN/OUT Switching

The two resistors connected at Pin 5 set the overall control path gain, and hence the system sensitivity. A lower tap point will decrease the sensitivity for high signal level sources, and a higher tap point will accommodate lower level sources. For purposes of initial calibration it is best to replace the resistors with a 1 k $\Omega$  potentiometer (the wiper arm connecting through C<sub>6</sub> to Pin 6), and follow the procedures outlined below. Once the correct adjustment point has been found, the position of the wiper arm is measured and an equivalent pair of resistors are used to replace the potentiometer. This, of course, can be done only if the source has a relatively fixed noise floor—the output from an audio cassette tape for example. For an add-on audio accessory the potentiometer should be retained as a front panel control to allow adjustment for individual sources. Use of DNR with multiple sources is described later.

### SYSTEM CALIBRATION

System calibration can be performed in a number of ways. With the source connected play a blank but biased section of the cassette tape. Set the potentiometer so that the wiper arm is at ground and then steadily rotate it until a slight increase in the output noise level is heard. Alternatively, with source program material present, set the potentiometer with the wiper arm connected to the Pin 5 end of the slider and again rotate until the high frequency content of the program material appears to begin to be attenuated. Then return the potentiometer wiper slightly towards Pin 5 so that the music is unaffected.

A third method of adjustment can be done with an oscilloscope monitoring the voltage on the control path detector filter capacitor, Pin 10. This will show a steady dc voltage around 1V while the wiper arm of the potentiometer is at ground. As the wiper arm is rotated, this voltage will start to increase. About 200 mV above the quiescent value will usually be the right point. Note that this will not be a steady dc voltage but a random peak, low amplitude sawtooth waveform caused by peak detection of the source noise in the control path.

Whatever method is used to determine the potentiometer setting, this setting should be confirmed by listening to a variety of programs and comparing the audio quality while switching DNR in and out of the circuit. This is easily accomplished by grounding Pin 9 which will disable the control path and force the audio filters to maximum bandwidth, Figure 2(a). Also shown is a second method of ON/OFF switching that gives an increased maximum bandwidth over that obtained in normal operation. Although the switch is not a required front panel control it can be an important feature. Unlike compander systems, DNR can be switched out leaving the source completely unprocessed in any way. With a switch, the user can always be assured that the noise reduction is not affecting the program material.

Apart from the basic circuit shown in Figure 1, all applications of the DNR system have another feature in common—the location of the LM1894 in the signal chain. As Figure 3 shows, the LM1894 is *always* placed right after the signal source pre-amplifier and *before* any circuit that includes user adjustable controls for volume or frequency response. The reasons for this are obvious. If the gain of the signal amplifier preceding DNR is changed arbitrarily, the noise input level to the LM1894 will not be at the correct point to begin activation of the audio path filters. Either reduced noise reduction will be obtained, or the high frequency content of the program material will be affected. A change in system gain prior to the LM1894 requires a corresponding change in the control path threshold sensitivity. Similarly modifying the frequency response, by heavy boost or cut of the mid to high frequencies, will have the same effect of changing the required threshold setting—apart from modifying the masking qualities of the program material.

### HOW MUCH NOISE REDUCTION?

The actual sensitivity setting that is finally used, and the amount of noise reduction that is obtained, will depend on a number of factors. As the data sheet for the LM1894 and other application notes have explained in some detail, the noise reduction effect is obtained by audio bandwidth restriction with a pair of matched low-pass filters. A CCIR/ARM\* weighted noise measurement is used so that the measured improvement obtained with DNR correlates well to the subjective impression of reduced noise. This is another way of stating that the source noise spectrum level versus frequency characteristic can have a large impact on how "noisy" we judge a source to be—and concomitantly how much of the "noisiness" can be reduced by decreasing the audio bandwidth. Fortunately most of the audio noise sources we deal with are smooth although not necessarily flat, resembling white noise. The weighting characteristic referred to above generally gives excellent correlation. For example, if the source -3 dB upper frequency limit is only 2 kHz (an AM radio), reducing the audio path bandwidth down to 800 Hz will improve the S/N ratio by only 5 to 7 dB. On the other hand, if the source bandwidth exceeds at least 8 kHz then from 10 dB to 14 dB noise reduction can be obtained. Of course, it is always worth remembering that this is the reduction in the source noise—any noise added in circuits *after* the LM1894 may contribute to the audible output and prevent the full noise reduction effect. To see how easily this can happen, we will consider the noise levels at various points in a typical automotive radio using an I/C tone and volume control, and an I/C power amplifier, both with and without noise reduction of the cassette player.

\*See pp. 2-9 to 2-10, Audio Handbook, National Semiconductor 1980.

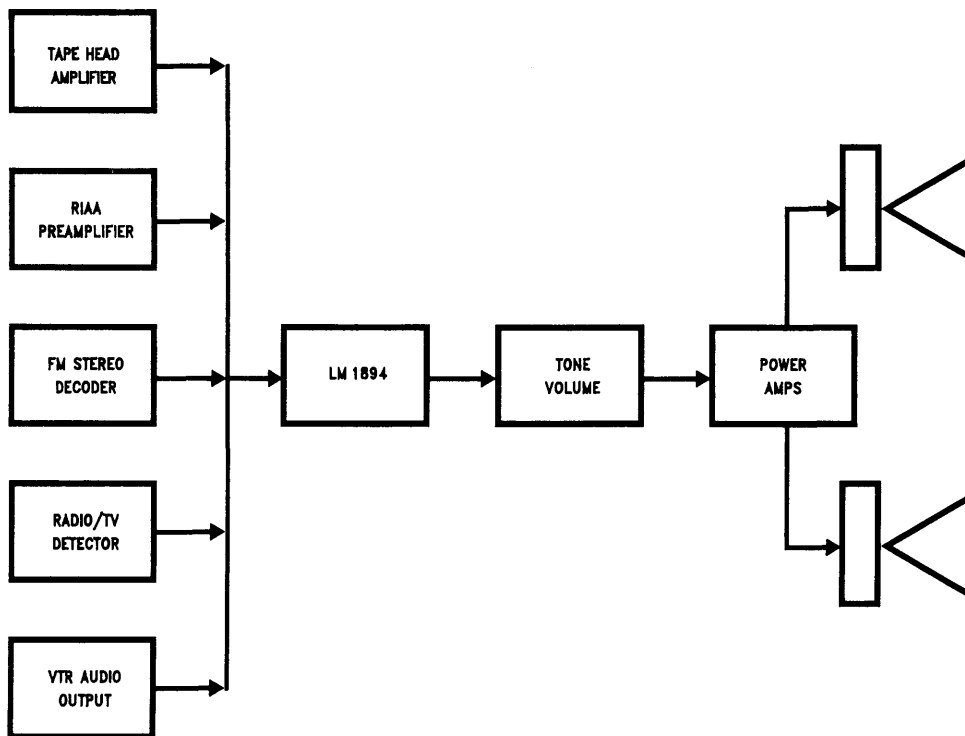
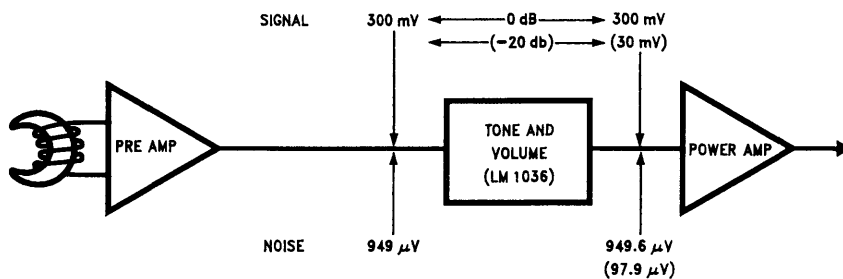


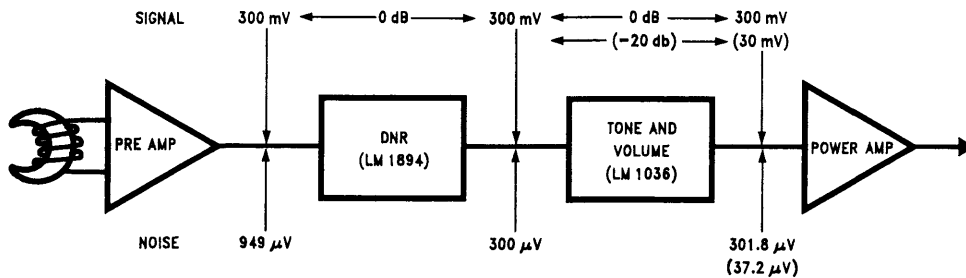
FIGURE 3. Location of DNR in Audio Systems

TL/H/8420-3



(a) Without Noise Reduction

TL/H/8420-19



(b) With Noise Reduction

TL/H/8420-4

FIGURE 4. Signal and Noise Levels in the Audio Path

If we assume that the tape head pre-amplifier gain is such that the nominal output level (corresponding to  $O''VU''$ ) is 300 mVrms, then for a typical cassette tape the noise will be 50 dB lower, or 949  $\mu$ V. The gain of the tone and volume control (an LM1036) is unity or 0 dB at maximum volume setting, with an output noise level of 33  $\mu$ V with no signal applied. With the tape pre-amplifier connected, the output noise from the LM1036 will be  $V_n$  where

$$V_n = 10^{-6} \sqrt{(33)^2 + (949)^2} = 949.6 \mu\text{V} \quad (2)$$

Clearly, the LM1036 has caused an insignificant increase in the background noise level (0.006 dB). Even when the volume control is set at  $-20$  dB overall gain, the LM1036 intrinsic noise level is 22  $\mu$ V. The tape noise level is now 94.9  $\mu$ V ( $-20$  dB) and the output noise  $V_n$  is

$$V_n = 10^{-6} \sqrt{(22)^2 + (94.9)^2} = 97.4 \mu\text{V} \quad (3)$$

Once more an insignificant contribution on the part of the LM1036 (0.23 dB).

Now we add noise reduction between the tape head amplifier and the LM1036. Usually this will mean over 10 dB reduction in the tape noise so that the input of the LM1036 sees 300  $\mu$ V noise. At 0 dB gain we have

$$V_n = 10^{-6} \sqrt{(33)^2 + (300)^2} = 301.8 \mu\text{V} \quad (4)$$

But at  $-20$  dB

$$V_n = 10^{-6} \sqrt{(22)^2 + (30)^2} = 37.2 \mu\text{V} \quad (5)$$

When we compare the results of Equation (3) and (5) we see that at  $-20$  dB gain setting we are getting only 8.4 dB noise reduction compared to 10 dB at maximum gain! Since the volume control is not normally set to maximum, this is a significant loss.

Active tone and volume controls are not the only circuits that can contribute to a loss in noise reduction. Most modern automotive radios use I/C power amplifiers delivering in excess of 6 watts into 4  $\Omega$  loads—and even more if bridge amplifiers are employed. With a 12  $V_{DC}$  supply, the output signal swing is limited to less than 4 Vrms if clipping is avoided. Typical amplifiers have an input referred noise level of 2  $\mu$ Vrms, and with a gain of 40 dB (a typical value) the intrinsic output noise level is 200  $\mu$ Vrms, or 86 dB below clipping. For a normal listening level, the signal amplitude will be 20 dB below clipping which yields a S/N ratio of only 66 dB—which is just better than the noise reduced input to the amplifier.

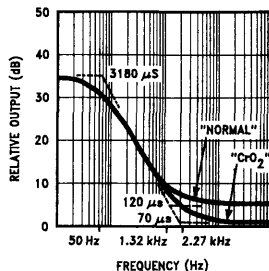
Many manufacturers recommend using I/C power amplifiers with gains of 60 dB. This will always result in unacceptable noise performance at moderate listening levels since the amplifier generated noise is now over 2 mV. For a signal 20 dB below clipping the output S/N ratio is only 46 dB!

It is interesting to note that the inclusion of just 10 dB noise reduction is sufficient to put pressure on the performance standards of the remaining circuits in the audio path of an automotive radio. If more noise reduction is available, such as a combination of Dolby B and DNR, or Dolby C, then the subsequent gain distribution must be considered even more carefully. The power amplifier gain may have to be reduced to 20 dB to avoid degrading the noise performance. In fact it may be impractical to realize the full noise performance capability of systems providing high levels of noise reduction in many automotive stereo radios.

## MODIFICATIONS TO THE STANDARD APPLICATIONS CIRCUIT

### 1. TAPE DECKS WITH EQUALIZATION SWITCHES:

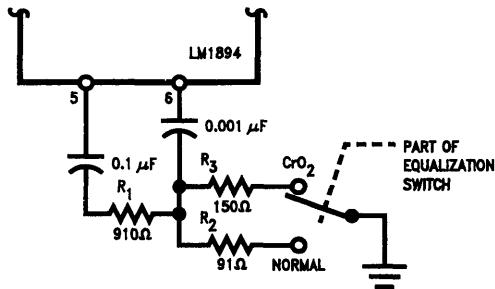
Many modern cassette tape decks and automotive radio cassette players offer at least two types of equalization in the head-preamplifier in order to optimize the frequency response of various tape formulations. These are often identified on the equalization switch as "Normal" and "CrO<sub>2</sub>" corresponding to 120  $\mu$ s and 70  $\mu$ s time constants in the equalization network. This difference in time constants can mean that the noise floor from a cassette tape in the "CrO<sub>2</sub>" mode can be up to 4 dB lower than for a tape requiring the "Normal" mode, *Figure 5*.



TL/H/8420-5

FIGURE 5. Tape Playback Equalization Including Integration

Although a compromise setting can be found for the DNR threshold setting to accommodate both types of tape, a single pole, double throw switch ganged to the equalization switch will optimize performance for each mode. In the example given in *Figure 6*, the resistor values shown are from an application that yielded a 400 mVrms input to the LM1894 when the tape flux density was 200 nW/m. For different tape-head amplifiers the resistors  $R_1$  and  $R_2$  are selected using a "Normal" tape as a source, and then  $R_3$  is selected according to the relationship given in Equation (5).



TL/H/8420-6

FIGURE 6. Optimizing the Control Path Threshold for Different Tape Formulations

Notice that only one additional resistor is required over the standard application, and it is easy to substitute transistor switching in place of the spdt switch.

$$R_1/(R_1 + R_2) = 0.63 R_3/(R_1 + R_3) \quad (5)$$

### 2. TAPE DECKS WITH COMPLEMENTARY NOISE REDUCTION:

Most cassette decks available today employ some form of complementary (companding) noise reduction system, usually Dolby B Type. DNR can be used in conjunction with these noise reduction systems as a means to provide yet more noise reduction on decoded tapes and still provide

noise reduction for unencoded tapes. The LM1894 is located after the companding system and provision must be made for the drop in noise level when the compandor is being used. The DNR threshold sensitivity is increased by the appropriate amount so that the lower noise levels are still able to activate the audio filters. For example, the circuit in *Figure 7* shows a switching arrangement to compensate for the 9 dB lower noise floor from a Dolby B decoded tape. Notice the change in resistor values  $R_1$  through  $R_3$  to raise the sensitivity (yet keeping the sum of  $R_1$  and  $R_2$  to 1k) and the 9 dB pad formed by the 3 k $\Omega$  resistor and the 1.5 k $\Omega$  resistor in parallel with the control path input Pin 6, for use when the compandor is switched off. Since the output level from the compandor is usually around 580 mV for a flux density of 200 nW/m, the ratio of  $R_1$  to  $R_2$  and  $R_3$  is changed by only 5.6 dB compared to that shown in the previous Figure where the input level was 400 mVrms.

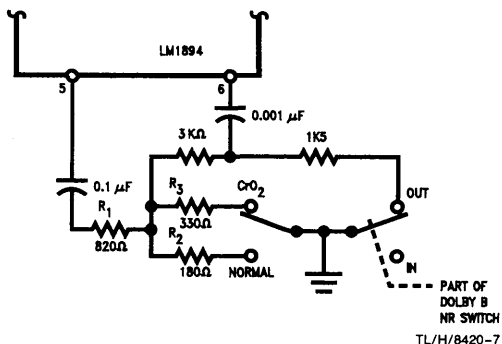


FIGURE 7. Switching with Other NR Systems

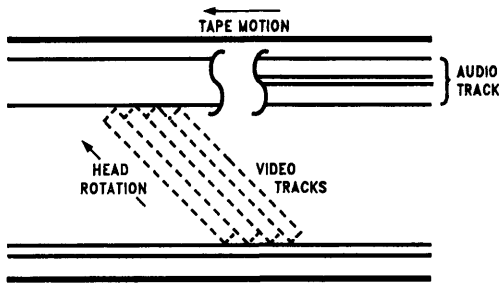


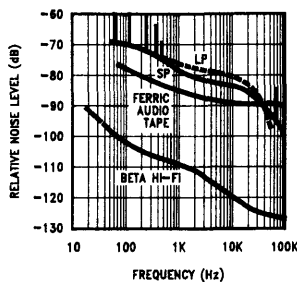
FIGURE 8. Video Magnetic Tape Format

### 3. VIDEO TAPE RECORDERS:

The audio track of a video cassette tape is similar to an audio cassette and appears along one edge of the tape. Although provision is made for two tracks, each 0.35 mm wide, a large number of recordings are monaural with a track width of 1 mm (0.04 inches).

Unlike the video heads, which are mounted on a rotating drum and angled to the direction of tape travel in order to give a much higher recording speed, the audio is recorded longitudinally with a separate head at 33.35 mm/sec for standard play, 16.88 mm/sec for long play, and 11.12 mm/sec for the very long play mode (VHS format tape machines). The noise spectrum is similar to an audio cassette but with a couple of differences. The typical frequency response from the head pre-amplifier does not extend beyond 10 kHz in the SP mode and is less in the LP and VLP modes. Even so, this bandwidth is enough to ensure the presence of the familiar tape "hiss" when played

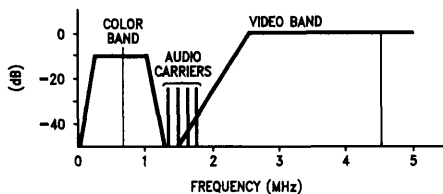
through modest or better Hi-Fi systems. Although the mono track width (twice as wide as an audio cassette stereo track) should help the S/N ratio, the slower tape speed does not, as shown in the curves of *Figure 9*. For the SP mode the S/N ratio is approximately 5 to 10 dB lower than the audio cassette and worsens by 3 to 5 dB in the extended play modes. Some "spurs" or "spikes" may be observed at harmonics of the video field frequency (60 Hz) and at the video line scan frequency of 15.734 kHz. The low frequency spikes will not affect DNR operation since the control path sensitivity decreases sharply below 1 kHz, but the presence of the 15.734 kHz component could cause improper sensitivity settings to be obtained. If this is the case, the pilot frequency notch filter for FM, described later, can be retuned by changing the capacitor from 0.015  $\mu$ F to 0.022  $\mu$ F.



TL/H/8420-9

FIGURE 9. Video Tape Noise Spectrum Levels

*Figure 9* also shows the noise spectrum with the new Beta Hi-Fi format. This is clearly superior to both the standard format and audio cassette tapes and is realized by using the two video record/play heads simultaneously for audio, thus taking advantage of the substantially higher relative tape speed. The audio is added in the form of four FM carriers, *Figure 10*. Four carriers are necessary for two audio channels since the azimuth loss between the normal video heads (reducing crosstalk between the heads at video frequencies) is not enough at the lower audio carrier frequencies. Each head therefore uses different carriers for the left and right channel signals.



TL/H/8420-10

FIGURE 10. Beta Hi-Fi Carrier Frequencies

A quite different technique is used for VHS Hi-Fi, which is similar to that for 8 mm video. Separate audio heads are mounted on the same rotating drum that is carrying the video heads, but with a much larger azimuth angle compared to the video heads. The sound signal is written deep into the tape coating and then written over by the video signal which causes partial erasure of the audio—about a 10 dB to 15 dB loss. The difference in azimuth angle prevents crosstalk and the much greater writing speed still yields an S/N of over 80 dB.

Both Hi-Fi formats provide excellent sound quality with hardly any need for noise reduction but DNR can still play a role. Conventionally recorded tapes are and will be popular for quite a while, and even with Hi-Fi recording capability much

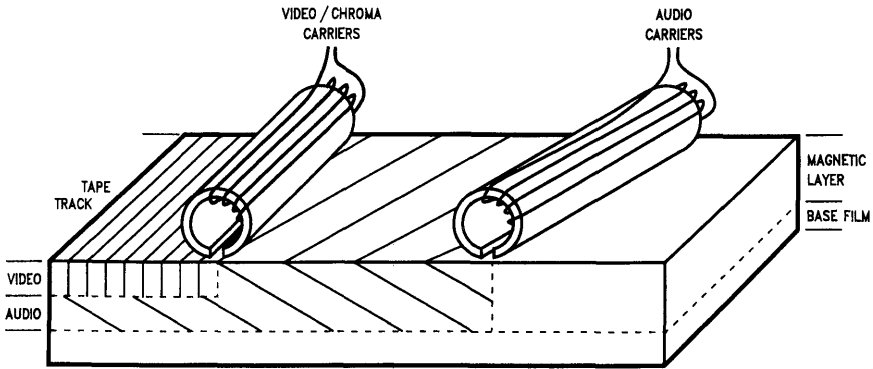


FIGURE 11. VHS Hi-Fi Recording Format

TL/H/8420-11

recording will be done with television sound as a source—and the source noise will dominate now instead of the tape noise. As discussed later, DNR can be very effective in dealing with television S/N ratios, allowing much of the benefit of improved recording techniques to be enjoyed.

#### 4. FM RADIOS:

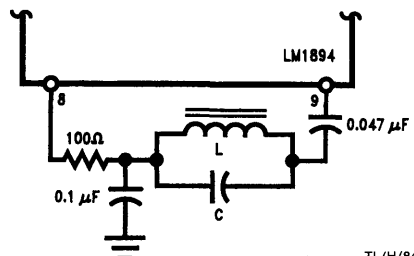
FM sources can present special problems to DNR users. The presence of the 19 kHz stereo pilot tone can be detected in the DNR control path and cause improper threshold settings (the problem is not so much that the 19 kHz tone gives the wrong setting, but that if the threshold is adjusted with the tone present, then the threshold is wrong when the tone is absent—as in a monaural broadcast). Secondly, for FM broadcasts the noise level at the receiver detector output is dependent on the r.f. field strength when this field strength is under  $100 \mu\text{V}/\text{meter}$  at the antenna terminal. With a fixed DNR threshold, as the noise level increases with decreasing field strength, the minimum audio bandwidth becomes wider and a loss in noise reduction is perceived. This latter problem occurs primarily with automobile radios where the signal strength can vary dramatically as the radio moves about. For the home receiver, re-adjustment of the DNR threshold setting for an individual station will compensate for the weaker signals.

To understand how much the pilot tone can affect the DNR control path, we can take a look at some typical signal levels. For an FM broadcast in the U.S., the maximum carrier deviation is limited to  $\pm 75 \text{ kHz}$  with a pilot deviation that is 10% of this value. A high quality FM I/C such as the LM1865 will produce a 390 mVrms output at the detector with this peak deviation, so the pilot level at 19 kHz will be 39 mVrms. If the receiver does not include a multiplex filter, after de-emphasis 4 mV will appear at the inputs to the LM1894. Typically for FM signal noise floors, the resistive divider at Pin 5 will attenuate the pilot by 20 dB leaving 0.4 mVrms at Pin 6. This input level to the LM1894 control path is sufficient to cause the audio bandwidth to increase by over 1 kHz compared to the monaural minimum bandwidth. Of course, if the receiver does have a multiplex filter, which is common in high quality equipment or receivers that include Dolby B Type noise reduction, this problem will not happen, but otherwise we require an extra 15 dB to 20 dB attenuation at 19 kHz. This is obtained with a notch filter tuned to the pilot frequency connected between Pins 8 and 9 of the LM1894. Although a tuned inductor is shown, a fixed coil of similar inductance and Q can be used since with

normal component value tolerances ( $\pm 7\%$  inductance,  $\pm 10\%$  capacitance) the pilot tone will be attenuated by at least 15 dB.

Handling the signal strength dependence of the FM signal noise floor is not quite as easy — at least if pre-set DNR sensitivity settings are used. A look at the quieting curves for an FM radio will show why. At strong signal levels, greater than  $1 \text{ mV}/\text{meter}$  field strength at the antenna, the IF amplifier of the radio is in full limiting and the noise floor is between 60 dB and 80 dB below the audio signal. However, as the field strength starts to decrease below  $1 \text{ mV}/\text{meter}$ , the noise level begins to increase, even though the IF amplifier is still in limiting. Worse yet, since the demodulated output includes the noise from the stereo difference signal channel (L-R), the noise level is increasing more rapidly in the stereo mode than in the monaural mode. By the time the field strength has fallen to  $100 \mu\text{V}/\text{m}$  the stereo noise is over 20 dB higher than the equivalent mono noise. If the DNR sensitivity is pre-set such that noise at the  $-45 \text{ dB}$  to  $-55 \text{ dB}$  level is activating the control path detector, when weaker stations are tuned in the noise level will increase and less noise reduction will be obtained. On the other hand, for stronger stations the noise level will drop below the detector threshold and a possibility exists that high frequency signals will be attenuated. Fortunately this latter occurrence is unlikely with commercial FM broadcasts since substantial signal compression is common, and the relatively high mid-band signals will be adequate enough to open the audio bandwidth sufficiently. In any event, with very strong r.f. signals, the need for noise reduction is minimal and DNR can be switched out.

	FM	TV
L	4.7 mH	4.7 mH
C	0.015 $\mu\text{F}$	0.022 $\mu\text{F}$

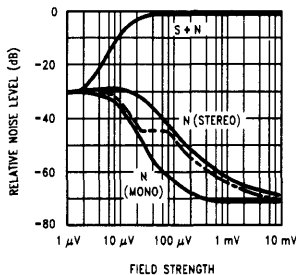


TL/H/8420-12

FIGURE 12. Control Path Notch Filter



Recognizing that a fixed threshold setting is necessarily a compromise for FM, the designer can still elect to use a pre-set adjustment for convenience. The set-up procedure is a little more complicated than for an audio tape source and involves the use of an FM signal generator. The carrier frequency from the generator (between 88 MHz and 108 MHz) is unmodulated except for the stereo pilot tone, and the receiver is tuned to this carrier frequency. Then the carrier level is increased until the stereo demodulator output S/N ratio is that desired for the DNR threshold setting. For example, if the recovered audio output is 390 mVrms for 75 kHz deviation of the carrier frequency, the stereo noise level is 2.2 mVrms for a 45 dB S/N ratio. The generator level is increased until this noise voltage is measured at the demodulator output and the resistive divider at Pin 5 of the LM1894 adjusted correspondingly. A multiplex filter should be inserted between the decoder output and the S/N meter to prevent the pilot tone from giving an erroneous reading. At no time should the pilot tone be switched off since this will allow the decoder to switch into the monaural mode, decreasing the noise level -65 dB instead. A S/N ratio of 45 dB is chosen since many modern receivers incorporate blending stereo demodulators. As the dashed curve of *Figure 13* shows, when the stereo S/N ratio falls to 45 dB, the decoder starts to blend into monaural operation, thus keeping a constant S/N ratio. The loss in stereo separation that inevitably accompanies this blending is far less objectionable than abrupt switching from stereo to mono operation at weak signal levels.



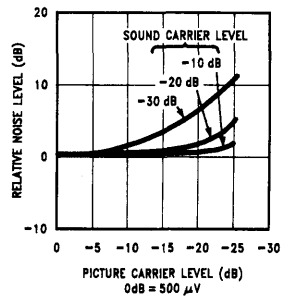
TL/H/8420-13

FIGURE 13. FM Radio Quieting Curves

## 5. TELEVISION RECEIVERS:

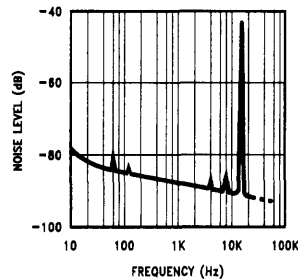
At first it might be thought that television broadcast signals, with an FM sound carrier located 4.5 MHz above the picture carrier frequency, will present the same difficulties as FM radio broadcasts to a DNR system with a pre-set threshold. This conclusion is modified by two considerations. First the TV receiver is unlikely to be mobile and the received signal strength will be relatively constant from an individual broadcast station. Secondly another subjective factor, the picture quality, will largely determine whether the signal strength is adequate enough for the viewer to stay tuned to that station. A representative television receiver will have a VHF Noise Figure between 6 dB and 7 dB such that, with a 75Ω antenna impedance, the picture will be judged noise-free at an input signal level of just above 0.5 mVrms - i.e. a picture signal to noise ratio of 43 dB. Noise will become perceptible to most viewers at a S/N ratio of 38 dB and become objectionable at 28 dB to 30 dB. Therefore 13 dB below 1 mVrms the picture noise is objectionable, and at -25 dB to -30 dB it will probably be totally unacceptable to the majority of viewers. For off-air broadcasts, the audio carrier ampli-

tude is 7 dB to 10 dB below the picture carrier amplitude and for cable services the typical sound/picture carrier ratio is -15 dB. However, due to the FM improvement factor (45.4 dB for equal amplitude carriers compared to the AM picture carrier) audio S/N ratios do not degrade as rapidly as the picture S/N—even with the lower audio carrier amplitudes. *Figure 14* shows the increase in audio noise level as both carrier amplitudes are reduced from the picture carrier level that produces a noise-free picture. When the picture noise is already objectionable the audio noise level has remained virtually unchanged, even for an audio carrier 30 dB below the picture carrier. By the time an unacceptable picture noise level has been reached, the audio noise has increased by less than 3 dB for sound carriers at -10 dB and -20 dB relative to the picture carrier. Therefore it is unlikely that a perceptible increase in noise compared to a strong channel will occur before the viewer switches to another channel.



TL/H/8420-14

FIGURE 14. Increase in Audio Noise with Decreasing Carrier Levels



TL/H/8420-15

FIGURE 15. TV Noise Spectrum Level

*Figure 15* shows the noise spectrum level of a strong audio carrier (1 mVrms) referred to 7.5 kHz carrier deviation. The standard peak deviation in the U.S. is 25 kHz so that the spectrum level will be 10 dB lower when referred to the peak audio level, meaning that the noise is not much better than the cassette tape noise levels shown previously. Only the relatively small power capability and limited bandwidth of audio amplifiers and speakers in conventional receivers has made this noise level acceptable. Unfortunately for the listener who hooks up the audio to his Hi-Fi system, or buys a new receiver with wider audio bandwidth and high output power (in anticipation of the proposed BTSC stereo audio broadcasts for television), TV sound will exhibit this noise.

Because the noise floor will be relatively constant, a pre-set threshold can be used for the LM1894 control path (although broadcast of older movies with unprocessed and noisy optical soundtracks might increase the received noise), and the only modification to the standard application circuit is to shift the control path notch filter down to 15.734 kHz. This is done with sufficient accuracy simply by changing the 0.015  $\mu\text{F}$  tuning capacitor to 0.022  $\mu\text{F}$ .

**Note:** The introduction of a stereo audio broadcast (the BTSC-MCS proposal) does not substantially modify the above conclusions, even though dbx noise processing is used. The dbx-TV noise reduction is applied only to the new stereo difference signal channel (L-R) to decrease the additional noise intrinsic in the use of an AM subcarrier along with the normal (L+R) monaural channel. This means that the new stereo signal should have roughly the same characteristics as the present monaural signal.

## 6. MULTIPLE SOURCES:

Multiple sources are best accommodated by keeping the potentiometer in the LM1894 control path and allowing the user to optimize each source. Nevertheless, for convenience, pre-sets are often desired and these can be done in two ways.

- 1) If the sources have widely different S/N ratios, the resistive divider at Pin 5 should be tapped at the appropriate point for each source noise level. This assumes that the source signal levels have been matched at the input to the LM1894 for equal volume levels.
- 2) If the source S/N ratios are not too far different, then the input levels can be trimmed individually to produce the same noise level in the LM1894 control path. A single sensitivity setting is used, and an additional switch pole ganged to the source selector switch is avoided.

Examples of both arrangements are shown in *Figure 16(a)* and *(b)*. To set up the multiple source system of *16(b)*, the DNR control path sensitivity is adjusted for the source with the lowest noise floor. Measure the peak detector voltage (Pin 10) produced by this noise source and then switch to the next source. Adjust (attenuate) the input level of the new source to match the previous Pin 10 detector voltage and repeat this procedure for each subsequent source.

## 7. CASCADING THE LM1894 AUDIO FILTERS

The LM1894 has two matched audio lowpass filters which can be cascaded, providing a single channel filter per I/C with a 12 dB/octave roll-off. This produces slightly more noise reduction (up to 18 dB) but because the steeper filter slope may in some cases produce audible effects on high frequency material, cascaded filters are best used for sources with a relatively restricted h.f. content. When the filters are cascaded the combined corner frequency decreases by 64% according to Equation (6), for  $n = 2$

$$f_c = f_0 \sqrt{100.3/n - 1} \quad (6)$$

Therefore, to retain the original frequency range, the capacitor values must be reduced by the same factor to 0.0022  $\mu\text{F}$ . One of the audio outputs is connected over to the other audio filter input and the summing amplifier in the control path is by-passed by moving the 0.1  $\mu\text{F}$  coupling capacitor from Pin 5 over to the single audio input. If the audio source is unable to drive the 1 k $\Omega$  impedance of the control path input network, this can be scaled up by using a 0.01  $\mu\text{F}$  capacitor and a 10 k $\Omega$  potentiometer.

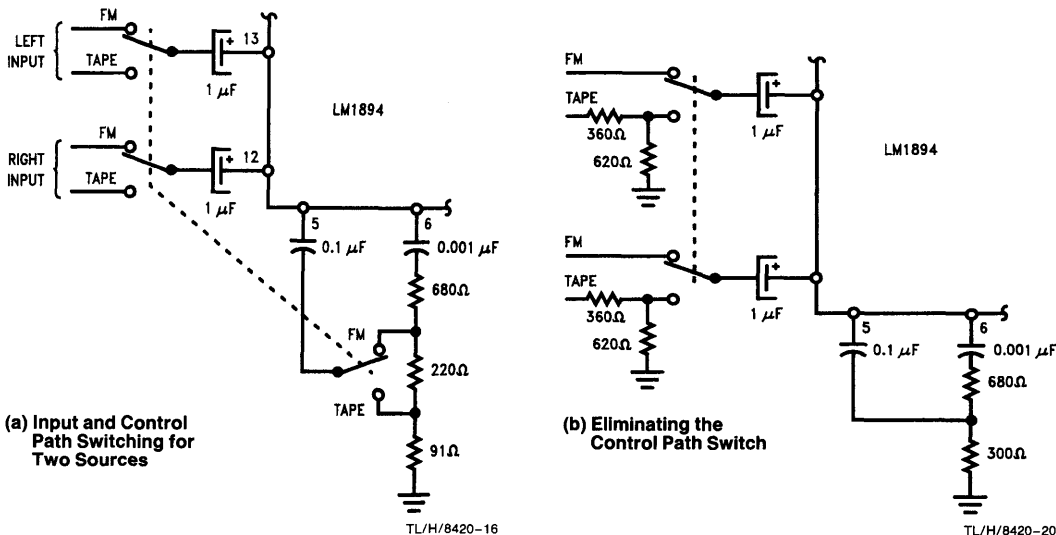


FIGURE 16. Multiple Programme Source Switching

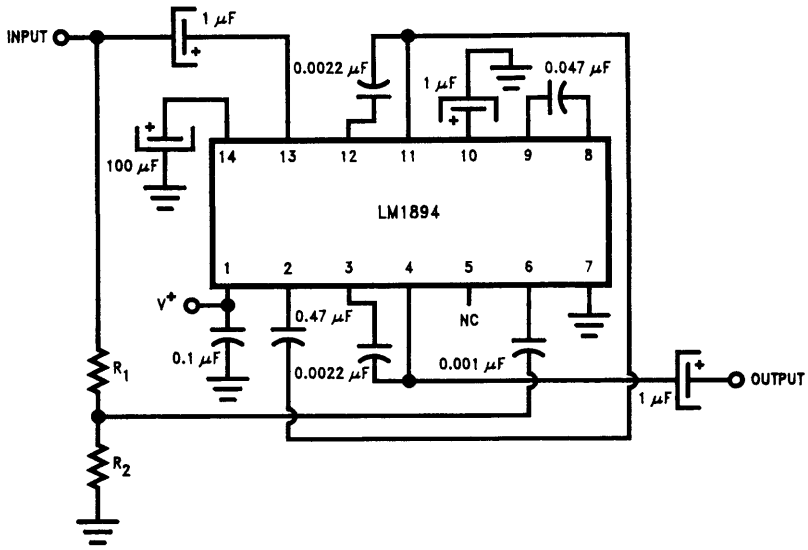


FIGURE 17. Cascading the Audio Filters of the LM1894

TL/H/8420-17