

DAC1020,LM10,LM118,LM3046,LM329,LM394

Audio Applications of Linear Integrated Circuits



Literature Number: SNVA530

Audio Applications of Linear Integrated Circuits

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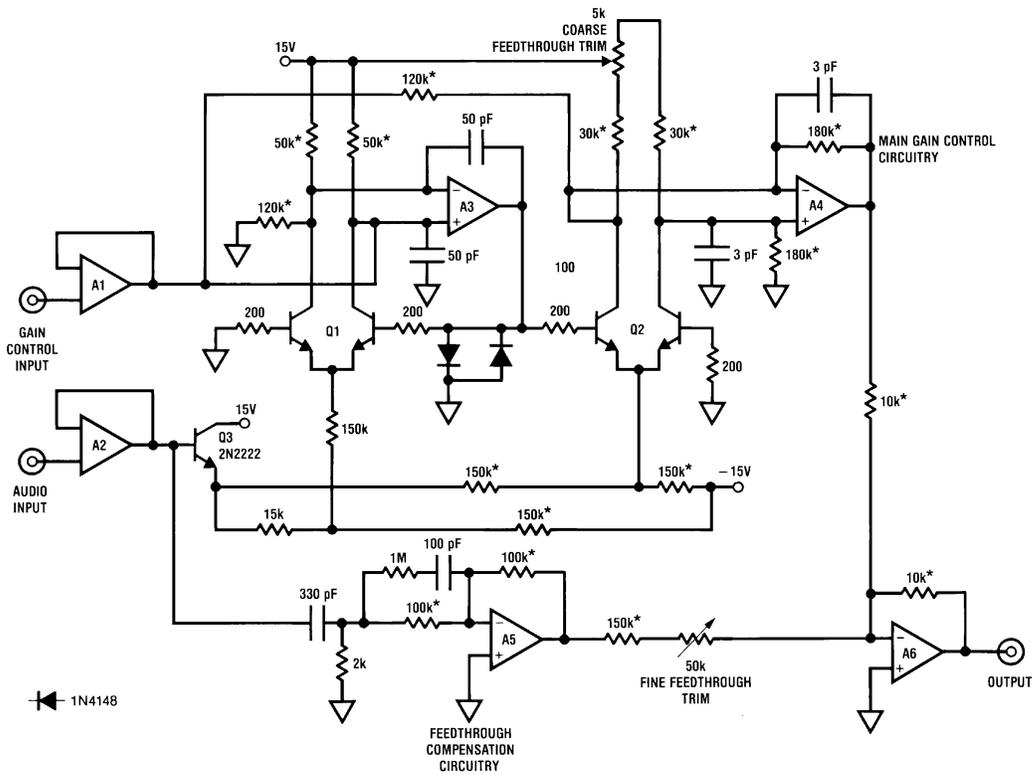
Although operational amplifiers and other linear ICs have been applied as audio amplifiers, relatively little documentation has appeared for other audio applications. In fact, a wide variety of studio and industrial audio areas can be served by existing linear devices. The stringent demands of audio requirements often mean that unusual circuit configurations must be used to satisfy a requirement. By combining off-the-shelf linear devices with thoughtful circuit designs, low cost, high performance solutions are achievable. An example appears in *Figure 1*.

EXPONENTIAL V-F CONVERTER

Studio-type music synthesizers require an exponentially responding V-F converter with a typical scale factor of 1V in per octave of frequency output. Exponential conformity requirements must be within 0.5% from 20 Hz–15 kHz. Almost all existing designs utilize the logarithmic relationship between V_{BE} and collector current in a transistor.

Although this method works well, it requires careful attention to temperature compensation to achieve good results. *Fig-*

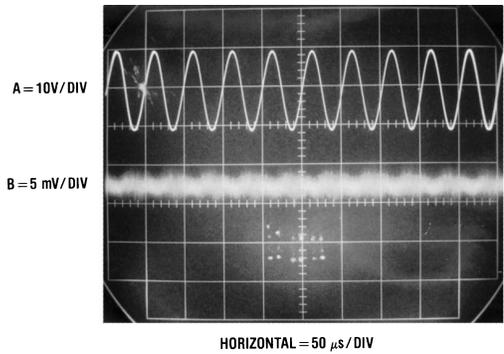
ure 1 shows a circuit which eliminates all temperature compensation requirements. In this circuit, the current into A1's summing junction is exponentially related to the circuit input voltage because of the logarithmic relationship between Q1's V_{BE} and its collector current. A1's output integrates negatively until the Q2-Q5 pair comes on and resets A1 back to 0V. Note that opposing junction tempcos in Q2 and Q5 provide a temperature compensated switching threshold with a small (100 ppm/°C) drift. The -120 ppm/°C drift of the polystyrene integrating capacitor effectively cancels this residual term. In this fashion, A1's output provides the sawtooth frequency output. The LM329 reference stabilizes the Q5-Q2 firing point and also fixes Q1's collector bias. The 3k resistor establishes a 20 Hz output frequency for 0V input, while the 10.5k unit trims the gain to 1V in per octave frequency doubling out. Exponential conformity is within 0.25% from 20 Hz to 15 kHz.



*1% film resistor
 A1-A2, A3-A4, A5-A6 = LF412 duals
 Q1-Q2 = LM394 duals

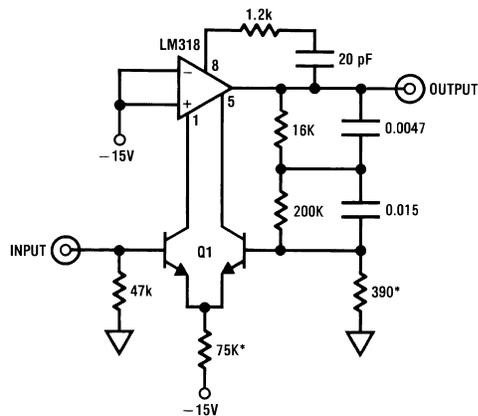
AN007496-2

FIGURE 2.



AN007496-3

FIGURE 3.



Q1 = LM394

AN007496-4

Frequency	Total Harmonic Distortion				
	<0.002	<0.002	<0.002	<0.002	<0.002
20	<0.002	<0.002	<0.002	<0.002	<0.002
100	<0.002	<0.002	<0.002	<0.002	<0.002
1000	<0.002	<0.002	<0.002	<0.002	<0.002
10000	<0.002	<0.002	<0.002	<0.0025	<0.003
20000	<0.002	<0.002	<0.004	<0.004	<0.007
Output Amplitude (Vrms)	0.03	0.1	0.3	1.0	5.0

FIGURE 4.

ULTRA-LOW NOISE RIAA PREAMPLIFIER

In *Figure 4*, an LM394 is used to replace the input stage of an LM118 high speed operational amplifier to create an ultra-low distortion, low noise RIAA-equalized phono preamplifier. The internal input stage of the LM118 is shut off by tying the unused input to the negative supply. This allows the LM394 to be used in place of the internal input stage, avoiding the loop stability problems created when extra stages are added. The stability problem is especially critical in an RIAA circuit where 100% feedback is used at high frequencies. Performance of this circuit exceeds the ability of most test equipment to measure it. As shown in the accompanying chart, harmonic distortion is below the measurable 0.002% level over most of the operating frequency and amplitude range. Noise referred to a 10 mV input signal is -90 dB down, measuring 0.55 μ Vrms and 70 pArms in a 20 kHz bandwidth. More importantly, the noise figure is less than 2 dB when the amplifier is used with standard phono cartridges, which have an equivalent wideband (20 kHz) noise of 0.7 μ V. Further improvements in amplifier noise characteristics would be of little use because of the noise generated by the cartridge itself. A special test was performed to check for transient intermodulation distortion. 10 kHz and 11 kHz were mixed 1:1 at the input to give an rms output voltage of

2V (input = 200 mV). The resulting 1 kHz intermodulation product measured at the output was 80 μ V. This calculates to 0.0004% distortion, quite a low level, considering that the 1 kHz has 14 dB (5:1) gain with respect to the 10 kHz signal in an RIAA circuit. Of special interest also is the use of all DC coupling. This eliminates the overload recovery problems associated with coupling and bypass capacitors. Worst-case DC output offset voltage is about 1V with a cartridge having 1 k Ω DC resistance.

MICROPHONE PREAMPLIFIER

Figure 5 shows a microphone preamplifier which runs from a single 1.5V cell and can be located right at the microphone. Although the LM10 amplifier-reference combination has relatively slow frequency response, performance can be considerably improved by cascading the amplifier and reference amplifier together to form a single overall audio amplifier. The reference, with a 500 kHz unity-gain bandwidth, is used as a preamplifier with a gain of 100. Its output is fed through a gain control potentiometer to the op amp, which is connected for a gain of 10. The combination gives a 60 dB gain with a 10 kHz bandwidth, unloaded, and 5 kHz, loaded with 500 Ω . Input impedance is 10 k Ω .

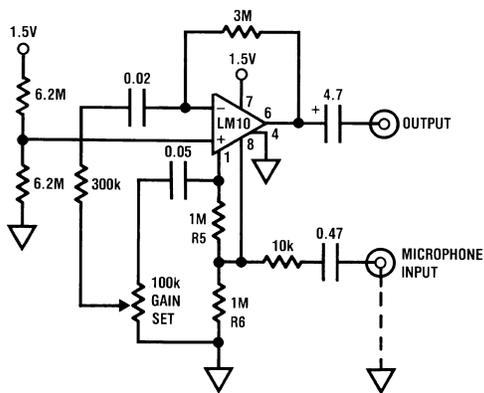


FIGURE 5.

AN007496-5

Potentially, using the reference as a preamplifier in this fashion can cause excess noise. However, because the reference voltage is low, the noise contribution which adds root-mean-square, is likewise low. The input noise voltage in this connection is $40 \text{ nV} - 50 \text{ nV}/\sqrt{\text{Hz}}$, approximately equal to that of the op amp.

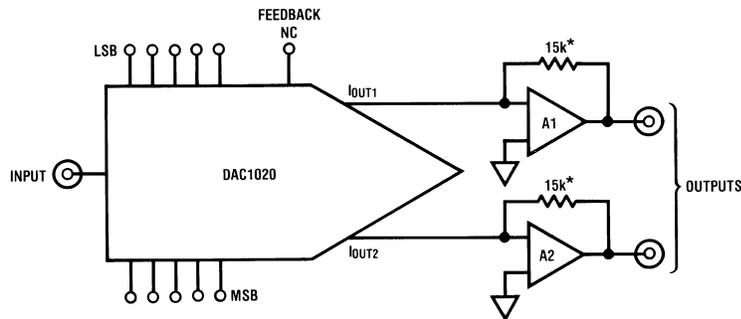
One point to observe with this connection is that the signal swing at the reference output is strictly limited. It cannot swing much below 150 mV, nor closer than 800 mV to the supply. Further, the bias current at the reference feedback terminal lowers the output quiescent level and generates an

uncertainty in this level. These facts limit the maximum feedback resistance (R5) and require that R6 be used to optimize the quiescent operating voltage on the output. Even so, one must consider the fact that limited swing on the preamplifier can reduce maximum output power with low settings on the gain control.

In this design, no DC current flows in the gain control. This is perhaps an arbitrary rule, designed to insure long life with noise-free operation. If violations of this rule are acceptable, R5 can be used as the gain control with only the bias current for the reference amplifier ($<75 \text{ nA}$) flowing through the wiper. This simplifies the circuit and gives more leeway in getting sufficient output swing from the preamplifier.

DIGITALLY PROGRAMMABLE PANNER-ATTENUATOR

Figure 6 shows a simple, effective way to use a multiplying CMOS D-A converter to steer or pan an audio signal between two channels. In this circuit, the current outputs of the DAC1020, which are complementary, each feed a current-to-voltage amplifier. The amplifiers will have complementary voltage outputs, the amplitude of which will depend upon the address code to the DAC's digital inputs. Figure 7 shows the amplifier outputs for a ramp-count code applied to the DAC digital inputs. The 1.5 kHz input appears in complementary amplitude-modulated form at the amplifier outputs. The normal feedback connection to the DAC is not used in this circuit. The use of discrete feedback resistors facilitates gain matching in the output channels, although each amplifier will have a $\approx 300 \text{ ppm}/^\circ\text{C}$ gain drift due to mismatch between the internal DAC ladder resistors and the discrete feedback resistors. In almost all cases, this small error is acceptable, although two DACs digitally addressed in complementary fashion could be used to totally eliminate gain error.



*1% film resistor A1, A2 = LF412 dual

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FIGURE 6.

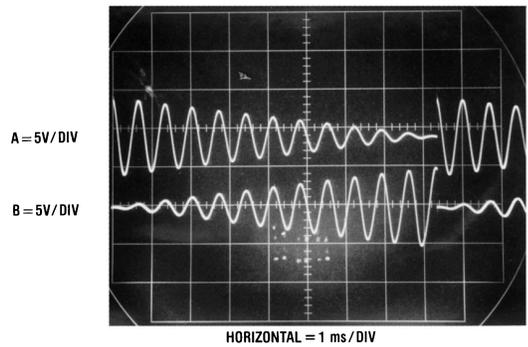
DIGITALLY PROGRAMMABLE BANDPASS FILTER

Figure 8 shows a way to construct a digitally programmable first order bandpass filter. The multiplying DAC's function is to control cut-off frequency by controlling the gain of the A3-A6 integrators, which has the effect of varying the integrators' capacitors. A1-A3 and their associated DAC1020 form a filter whose high-pass output is taken at A1 and fed to an identical circuit composed of A4-A6 and another DAC.

The output of A6 is a low-pass function and the final circuit output. The respective high-pass and low-pass cut-off frequencies are programmed with the DAC's digital inputs. For the component values shown, the audio range is covered.

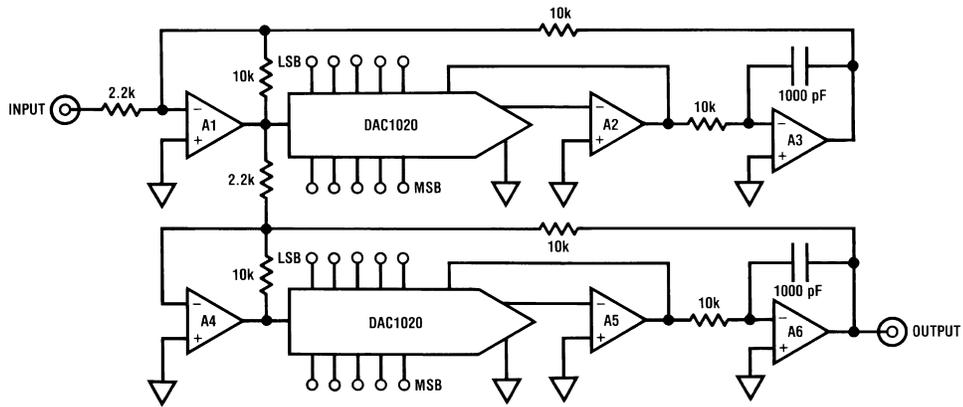
REFERENCES

Application Guide to CMOS Multiplying D-A Converters, Analog Devices, Inc. 1978



AN007496-7

FIGURE 7.



AN007496-8

A1, A2 = LF412 dual
A3, A4 = LF412 dual
A5, A6 = LF412 dual

FIGURE 8.

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