OP AMP APPLICATIONS

Walter G. Jung
Editor

ANALOG DEVICES

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▶ OP AMP APPLICATIONS

CHAPTER 6: SIGNAL AMPLIFIERS

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SECTION 6-1: AUDIO AMPLIFIERS

Walt Jung

Audio Preamplifiers

Audio signal preamplifiers (preamps) represent the low-level end of the dynamic range of practical audio circuits using modern IC devices. In general, amplifying stages with input signal levels of 10mV or less fall into the preamp category. This section discusses some basic types of audio preamps, which are:

■ *Microphone*— including preamps for dynamic, electret and phantom powered microphones, using transformer input circuits, operating from dual and single supplies.

■ *Phonograph*— including preamps for moving magnet and moving coil phono cartridges in various topologies, with detailed response analysis and discussion.

In general, when working signals drop to a level of ≈1mV, the input noise generated by the first system amplifying stage becomes critical for wide dynamic range and good signal-to-noise ratio. For example, if internally generated noise of an input stage is 1µV and the input signal voltage 1mV, the best signal-to-noise ratio possible is just 60dB.

In a given application, both the input voltage level and impedance of a source are usually fixed. Thus, for best signal-to-noise ratio, the input noise generated by the first amplifying stage must be minimized when operated from the intended source. This factor has definite implications to the preamp designer, as a "low noise" circuit for low impedances is quite different from one with low noise operating from a high impedance.

Successfully minimizing the input noise of an amplifier requires a full understanding of all the various factors which contribute to total noise. This includes the amplifier itself as well as the external circuit in which it is used, in fact *the total circuit environment must be considered*, both to minimize noise and maximize dynamic range and signal fidelity.

A further design complication is the fact that not only is a basic gain or signal scaling function to be accomplished, but *signal frequency response* may also need to be altered in a predictable manner. Microphone preamps are an example of wideband, flat frequency response, low noise amplifiers. In contrast to this, phonograph preamp circuits not only scale the signal, they also impart a specific frequency response characteristic to it. A major part of the design for the RIAA phono preamps of this section is a systematic analysis process, which can be used to predictably select components for optimum performance in frequency response terms. This leads to very precise functioning, and excellent correlation between a computer based design and measured lab operation.

Audio Buffers and Line Drivers

Audio line drivers and buffer amplifiers can take on a wide variety of forms. These include both single-ended and differential output drivers, as well as transformer isolated drivers. Within these general formats there are also many different performance options, and many of these are covered in this section. Note that a later section of this chapter also discusses buffers for a general context, as for video and instrumentation applications.

Many op amps useful as video drivers and/or buffers do well for audio drivers/buffers, because of the high current output stages necessary for good linearity over video bandwidths (see References 1-4). Some examples of video IC amplifiers that are audio-useful are the AD810, AD811 and AD812, AD815, AD817 and AD826, AD818, AD829, AD845, and AD847. Other types notable for either high or unusually linear output drives or other performance features useful towards audio are the AD797 and the OP275.

Some High Current Buffer Basics

As a preliminary to detailed application discussions, some basic circuit principles germane to high current buffers and drivers should be treated first. With output currents up to 100mA or more, "housekeeping" details of bypassing, grounding and wiring also become important, and must be considered to achieve high performance. These are briefly discussed here in the context of high current audio buffers, using the unity gain buffer circuit of Figure 6-35 below, as a point of departure.

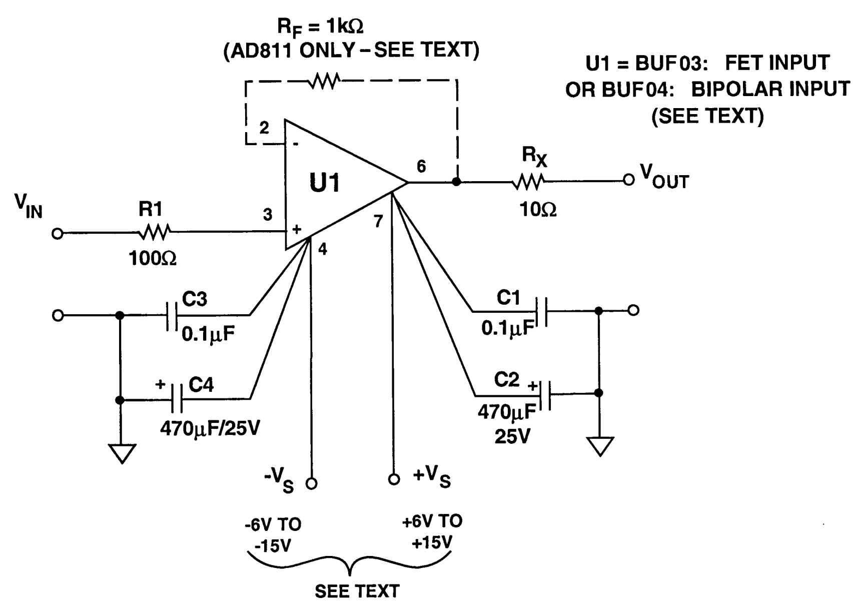


Figure 6-35: *A unity-gain, standalone buffer circuit*

First, despite which IC is used for U1, close attention should be given to making buffer stages free from parasitic effects, at both input, output, and supplies. Physical construction of buffer-drivers and other high current stages should be in accordance with high speed rules. A heavy copper ground plane is preferred, and circuit layout should be compact, with low capacitance high-Z nodes. Signal and ground runs should be laid out with signal coupling and load current flow in mind (see References 5-7 and Chapter 7).

In addition, the power supplies should be well bypassed close to the high current supply pins. In Figure 6-35 this is indicated by the Kelvin connections of C_1-C_4 to the U1 ±Vs pins. This should be used as standard practice for all high current stages, and is intended as a given for all the driver applications of this section.

As a minimum, local low inductance/low ESR RF bypass caps should used within 0.25" of the device supply pins, shown as C_1 and C_3. These are preferably 0.1µF stacked polyester film, or other low inductance capacitor type, preferably films. In addition, for high peak current loads, the high frequency bypasses are paralleled by local, short lead/large value, low ESR electrolytics such as C_2 and C_4, in a range of 470µF/25V and up. Note that capacitor ESR reduces in inverse proportion to electrical size and voltage rating, so larger size and/or voltage units help. These capacitors carry transient output currents, and should be aluminum electrolytic types rated for high frequency use, that is switching supply types. Such types tend to have a broad range of lowest high frequency minimum impedance and are thus less likely to cause power line resonance than are tantalum units.

DC power management and dissipation can also be important with buffer ICs. For example, the BUF03 and the AD811 ICs can dissipate fairly large power levels even with light loading, for supplies above ±12V. This is because the quiescent current of these devices is 15-18mA, relatively independent of operating voltage.

As a conservative general rule of reliability, any IC with a power dissipation above 300mW should not be used without a heat sink. For buffer or driver circuits using this power or more, use the lowest thermal resistance package possible, and add the appropriate heatsink (Thermalloy 2227 for the BUF03 or other TO-99 ICs, or Aavid #5801 for the BUF04, AD811 or other high dissipation 8 pin DIP ICs).

Output resistor R_X in this circuit should be 10 ohms or more, to isolate the buffer from capacitive loading (more on this, elsewhere in this chapter). For an extra safety margin against possible de-stabilization due to capacitive loads, make this resistor as high as feasible from a voltage loss point of view.

The input resistor R_1 is a "bullet-proof" safety item, and can serve two purposes. One is as a parasitic suppression device, which may be required for stability with some amplifiers (not absolutely essential for those here). A subtler feature of this resistance comes about when the buffer is operated within a feedback loop, and is driven from an op amp output. Internally, many buffer ICs have clamping diodes from input to output, and under overload conditions, these diodes act to clamp overdrive. With the inclusion of R_1, this prevents excess current drive into the buffer IC under this clamping condition.

Because of this stage's very high bandwidth, low phase shift, and low output impedance, fast buffers such as this can be used both "stand alone" just as shown, or as a more conventional "in loop" buffer as well, to minimize loading of a weaker, slower amplifier. The improvement raises the linear output up to ±100mA with the AD811 or the BUF04, while maximizing linearity, preserving gain, and lowering distortion.

Buffer THD+N Performance

Operating in a pure stand-alone mode, THD+N tests on several unity gain buffers are shown in Figure 6-36 below. These tests were for common conditions of 10Vrms output into a 600Ω load, operating from ±18V power supplies.

The BUF03, an open loop design, shows a distortion for these conditions of about 0.15%. The BUF04, a closed loop current feedback design buffer, shows a very low distortion of about 0.004%. The AD811 is also a current feedback amplifier, but it is externally configured as a unity gain follower, with $R_F = 1k\Omega$. Note that *all current feedback ICs will require such a resistor*, but the value required may vary part to part. The AD811 shows an intermediate distortion level, under 0.01%.

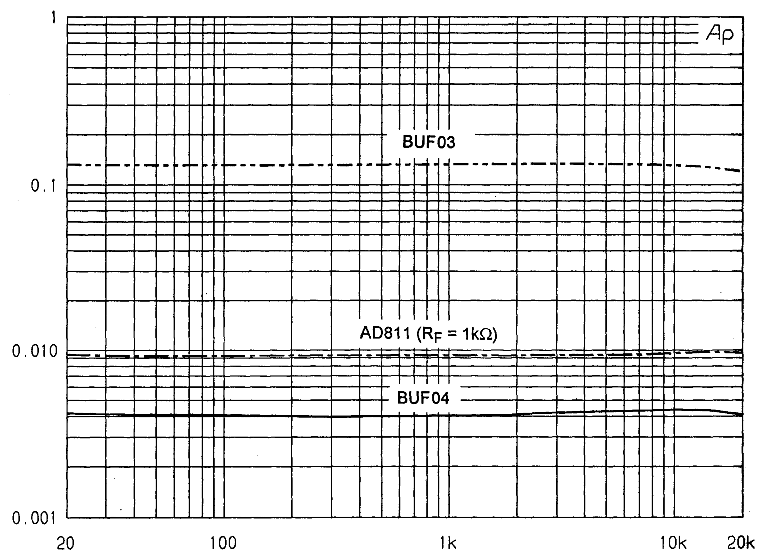


Figure 6-36: *THD+N (%) vs. frequency (Hz) for various buffer ICs, for $V_{OUT} = V_{IN} = 10Vrms$, $R_{LOAD} = 600\Omega$, V_S ±18V*

As a choice among these types, both the BUF04 and AD811 are capable of more than ±100mA of output, with input currents on the order of 1-2µA. The BUF03 has a lower output current (±70mA), but the advantage of a much lower input current (~200pA).

Dual Amplifier Buffers

In addition to standard operation of the various single op amps as unity-gain buffers, certain high output current *dual* op amp ICs also work exceedingly well as buffers. Using a dual IC to buffer a signal has the advantage of doubling the output drive while using basically the same package size, an obvious benefit. Two design steps allow this to be implemented successfully, The first is the selection of a basically linear single device that also is available as a dual. The second is to devise a method of combining the outputs of the two op amps in a linear fashion, without any side effects.

Among the suitable candidates for this task are the dual version of the AD817, the AD826, as well as the AD811 dual, the AD812. Figure 6-37 below shows a hookup that is useful towards increasing buffer output current to more than 100mA.

Ignoring Q1-Q2 for the moment, the circuit can be seen as a pair of unity-gain followers paralleled at the output, through small value resistors R_3 and R_4. These resistors provide balanced drive from the U1A and B sections, linearly combining the signals. With the use of the voltage feedback AD826 op amp, the circuit is quite simple, since R_1 and R_2 reduce to zero. If the AD812 current feedback device is used, these two resistors should be 1kΩ (shown dotted). The Q1-Q2 bi-directional clamp circuit is optional, and when used can provide protection against input overdrive, and/or adjustable current limiting via R_5.

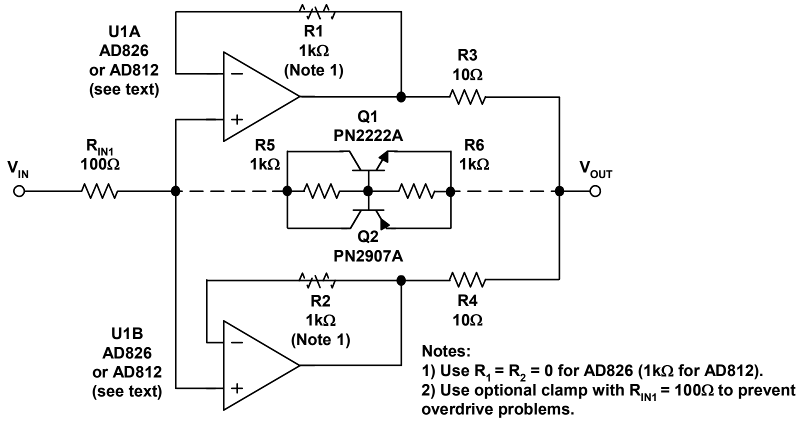


Figure 6-37: *Dual op amp buffer circuit raises output current to more than 100mA with low distortion*

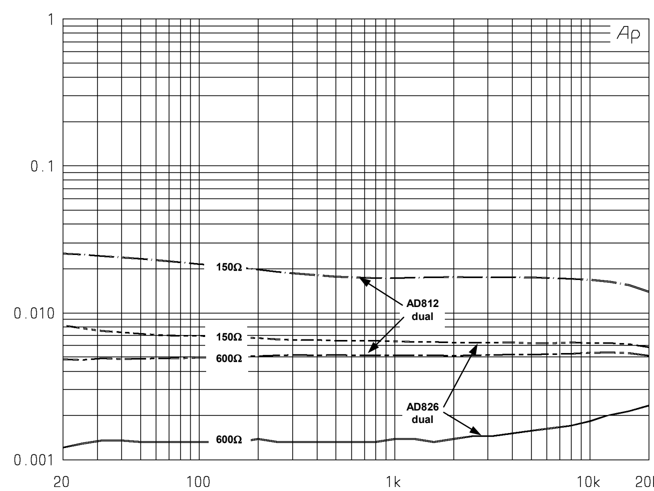


Figure 6-38: *THD+N (%) vs. frequency (Hz) for AD826 and AD812 dual buffer ICs, for $V_{OUT} = V_{IN}$ = 10Vrms, R_{LOAD} = 150/600Ω, V_S ±18V*

This circuit offers excellent performance, as shown in Figure 6-38 above. These THD+N plots show performance with loads of both 600 and 150Ω, at an output level of 10Vrms, from ±18V supplies (without clamping active). The AD826 offers the lowest distortion, due to the voltage feedback architecture, with less than 0.01% THD+N, even when driving 150Ω, which is an approximate 100mA combined peak output.

Capacitive Loading Issues

Audio driver output stages are typically operated as voltage sources feeding high impedance loads. When connected via long transmission lines between stages, the result is that the driver sees an unterminated line, which can appear highly capacitive. Audio driver stability with capacitive loading can be a difficult design issue, but for good reason— it isn't always an easy thing to achieve. If easy, it may be at the expense of performance or circuit complexity. Fortunately, some standard techniques exist for stabilizing op amp drivers with capacitive loads, and these can be implemented in a reasonably direct fashion. These are covered in detail within the next section of this chapter. The discussions immediately following emphasize driver linearity.

Op Amp Device/Topology Related Distortions

Single-ended audio drivers can be built using a linear, non-inverting gain stage as a starting point. Indeed such a circuit, given appropriate op amp choice and gain scaling, can well serve as a basic audio driver. Topologically, a non-inverting gain stage is preferable, since it loads the signal source less, and, it also adds no sign inversion. However, this configuration is subject to certain distortions, which should be understood in order to extract the best performance in an application. Distortion performance for a number of audio op amps in such a line driver circuit are now discussed, in this context.

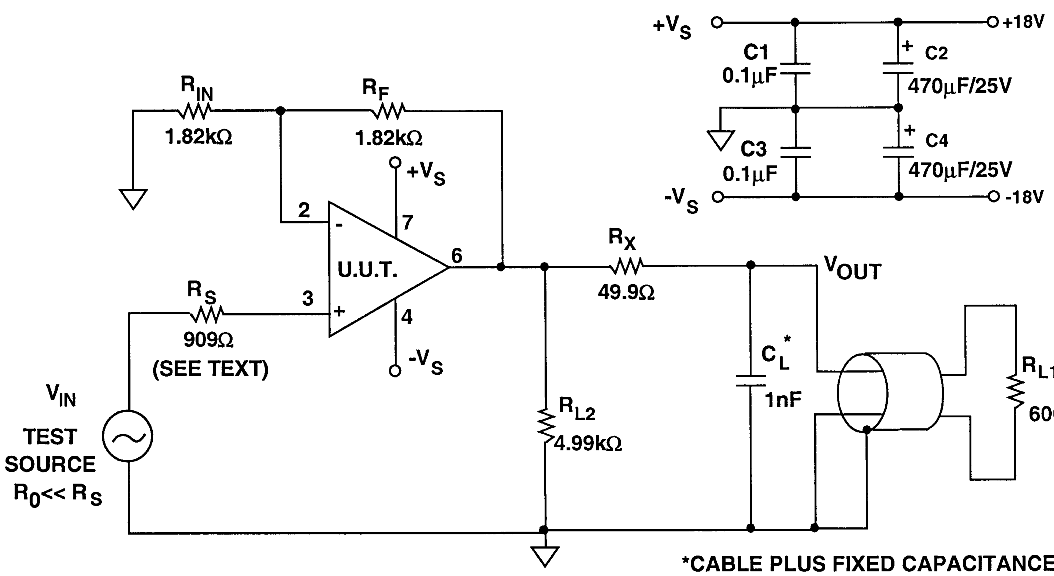


Figure 6-39: *Test circuit for audio line driver amplifiers*

The circuit of Figure 6-39 above is a test configuration that loads the U.U.T. op amp with 500Ω and 1nF. This is a reasonably stressful test load, which can differentiate the distortion of various devices with outputs of 7Vrms or more. A gain of 2 is used, which subjects the U.U.T to a relatively high input CM voltage, thus this configuration is sensitive to CM distortion in the amplifier. For the following tests of this section (except as noted to the contrary), V_S=±18V, and the analyzer bandwidth is 10Hz-80kHz.

Given amplifiers with sufficient load drive and output stage linearity in this circuit, there can still be non-linear effects due to the CM voltage. This distortion is due to the non-linear C-V characteristic seen at the two amplifier inputs, and can be minimized by matching the two impedances seen at the respective (+) and (–) inputs (see Reference 8). When this done, the differential component of the error is minimized, and the distortion seen in V_{OUT} falls to a minimum.

This general point is illustrated by Figure 6-40, a family of plots for an OP275 op amp within the circuit of Figure 6-39. The OP275 use junction FET devices in the input stage, which have appreciable (non-linear) capacitance to the substrate. The test is done with various values of source resistance R_S. As noted, distortion is lowest when R_S is equal to the parallel equivalent of R_F and R_{IN}, or in this case, about 910Ω. For either higher or lower values of R_S, distortion rises. Appreciably higher source impedance (10kΩ) can cause the distortion to rise lower in frequency, making performance much worse overall.

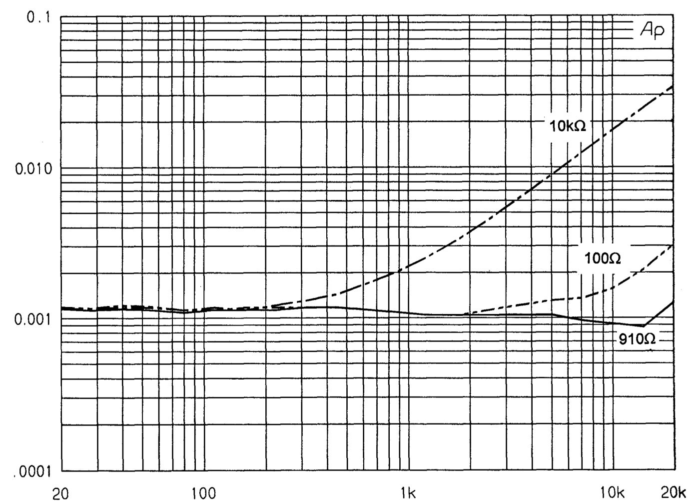


Figure 6-40: *Follower mode R_S sensitivity of OP275 bipolar/JFET input op amp-THD+N (%) vs. frequency (Hz), V_{OUT} = 7Vrms, R_L = 500Ω, V_S = ±18V*

It is therefore suggested that, whenever possible, amplifiers operated as voltage followers should have their source impedances balanced for lowest distortion. Note that the OP275 device is just one example, and its sensitivity to CM distortion effects is not at all unique in this regard.

While the balancing of the two source impedances is most helpful, *lowering the absolute value* can also minimize this distortion. With R_S low, this has the effect of moving the high frequency breakpoint of the distortion rise upwards in the spectrum, were it is less likely to be harmful. The best overall control of this distortion mechanism with an amplifier subject to it is the use of the lowest practical, balanced source impedances.

It is important to understand that virtually all IC op amps, *particularly those using JFET inputs*, as well as discrete JFET and bipolar transistors are subject to non-linear C-V effects, to some degree. In the tests of other amplifiers within the Figure 6-39 circuit, R_S was maintained at 910Ω, so as to minimize the effects of this distortion mechanism.

With high output, high slew rate linear amplifiers, the distortion generated for these test conditions can parallel that of the test equipment residual, as shown in Figure 6-41. Here the AD817, AD818 and AD845 amplifiers show THD+N which is essentially equal to the residual for these conditions, and appreciably below 0.001%.

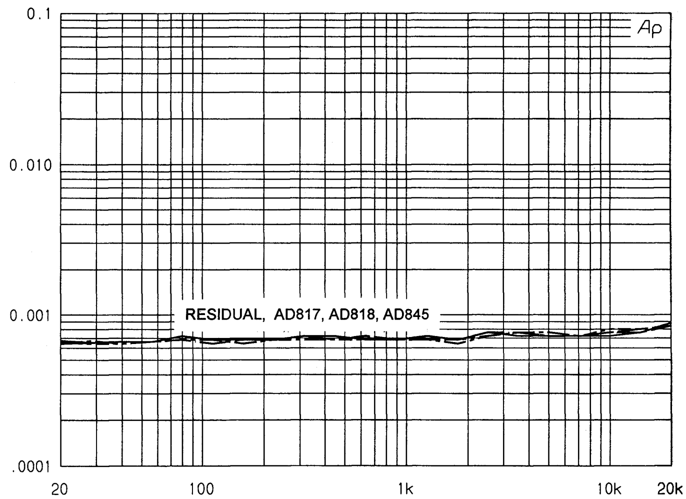


Figure 6-41: *A driver group, THD+N (%) vs. frequency (Hz), for V_{OUT} = 7Vrms, R_S = 909Ω, R_L = 500Ω, V_S = ±18V*

Amplifier types expressly designed for audio use also do well for these THD+N tests, as shown in Figure 6-42. The industry standard 5534 is near or just above the residual level, while the OP275 plot falls just above the 0.001% level and the 5532 is slightly higher.

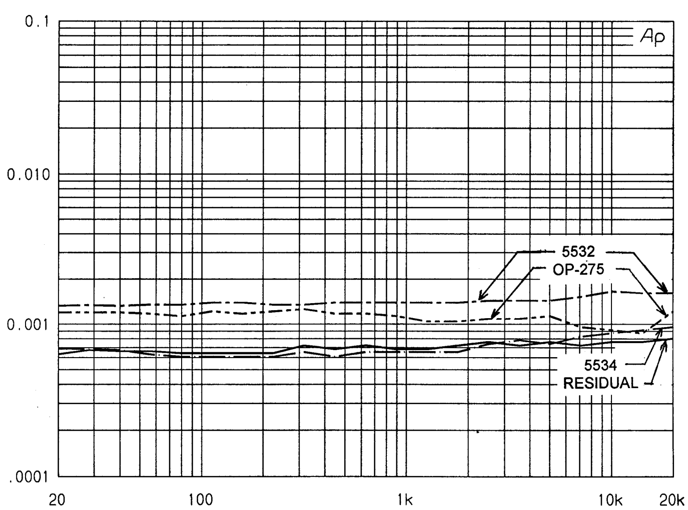


Figure 6-42: *B driver group, THD+N (%) vs. frequency (Hz), for V_{OUT} = 7Vrms, R_S = 909Ω, R_L = 500Ω, V_S = ±18V*

These tests reflect performance of a variety of single amplifiers, as exercised for matched-source test conditions, with medium output loading of 600Ω. Varying test conditions may change the absolute levels of performance. So also may different samples, or in the case of industry standard parts, alternate vendors.

The data of Figs. 6-41 and 6-42 reflect older (but available) op amp devices capable of very high performance in these tests. Several more recent devices also do well for this driver test. Figure 6-43 below shows performance of newer FET input op amps, the AD825, the AD8610 and the AD8065. The AD825 was tested under conditions identical to those of Fig. 6-41 and 6-42. The AD8610 and AD8065 were tested under similar conditions, but with ±13V power supplies, reflecting a lower maximum supply rating.

Interestingly, note that the latter two amplifiers still can accommodate more than a 7Vrms output swing, even with the reduced supplies. Under these conditions, the AD8065 distortion is near the test set residual and the AD8610 slightly higher at high frequencies. The AD825 has somewhat higher distortion, but this is almost frequency independent.

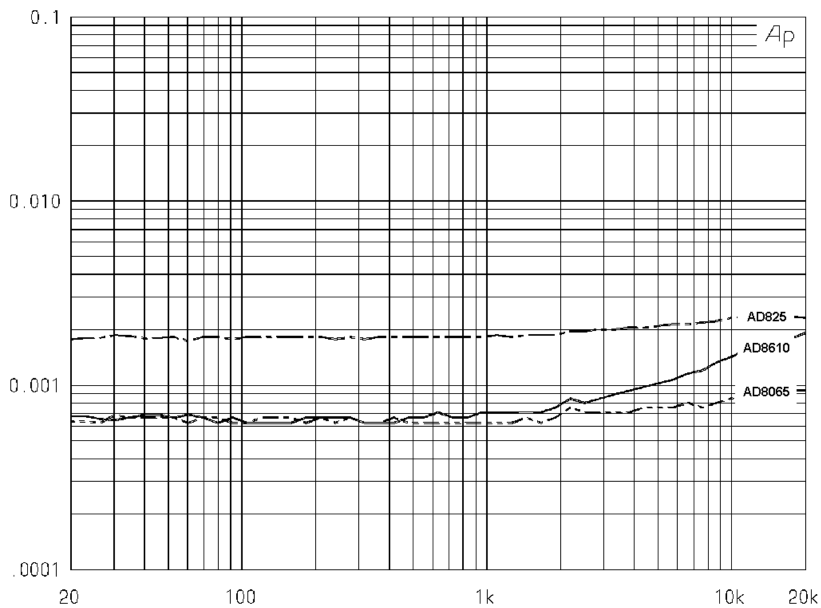


Figure 6-43: *C driver group, THD+N (%) vs. frequency (Hz), for V_{OUT} = 7Vrms, R_S = 909Ω, R_L = 500Ω, V_S = ±13V or ±18V*

Single-Ended Line Drivers

This section discusses a variety of line driver circuit examples that drive single-ended lines, optimized for different operating environments, supply voltages, and performance.

Consumer Equipment Line Driver

One common driver application is a line output stage for consumer preamps, CD and DVD players, etc. This is typically an economical audio stage with a nominal gain of 5 to 10 times, operating from supplies of ±10V to ±18V, usually with a rated output of 2-3Vrms, and a capability of driving loads of 10kΩ or more.

For simplicity of biasing and minimum output DC offset, AC coupling is used, and the circuit is typically fed from a volume control. For stereo operation, a dual channel device is typically sought for this type application, one which is also optimized for audio uses.

Such a stage is shown in Figure 6-44 below, and it uses an OP275 dual op amp as the gain element. In this circuit input and feedback resistors R_1 and R_3 are set equal, which makes the nominal DC bias at U1's output close to zero. The U1 bias current flowing in these resistors also serves to polarize coupling capacitors C_1 and C_2 positively, as noted. This bias is due to the sign of the OP275's PNP input stage bias currents, so reverse C_1-C_2 if an NPN input amplifier is used.

R_2 sets the gain of the stage in conjunction with R_1. The stage gain is nominally 5 times for the values shown. C_2 sets the low frequency rolloff along with R_2, which in this case is ≈0.3Hz. Although this frequency is quite low, it does allow some range of gain increase if desired, simply by lowering R_2. Output capacitor C_3 must be non-polarized, since the worst case DC at the output of U1 is ≤10mV (and can be bipolar). Typically it will be about ¼ this, so if a few mV can be tolerated, C_3 can be eliminated.

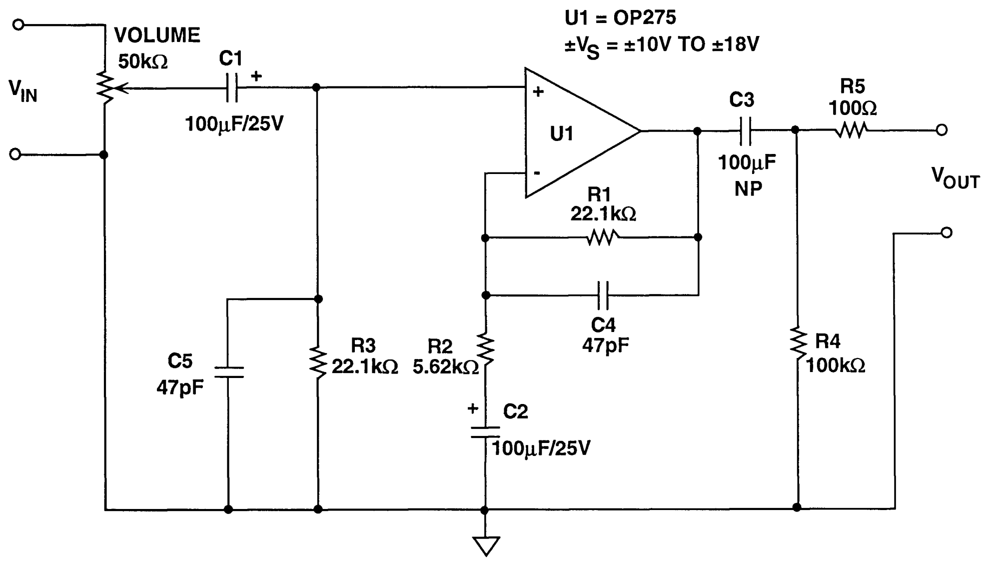


Figure 6-44: *Consumer equipment line driver stage*

THD+N performance of the stage (not shown) was measured for outputs of 1-3Vrms into a 10kΩ/600pF load, using ±18V supplies with an R_S of 1kΩ. At lower output levels performance is noise limited, measuring less than 0.002%. At the 3V output level a slight increase in high frequency distortion is noted. Although this application is an example where the amplifier ± source impedances cannot be matched (due to the variations of the volume control), nevertheless the performance is still quite good.

Noise is the limiting factor for lower level signals, so if lower noise is desired, R_2 can be reduced. The ultimate practical limit to noise is the volume control's finite output impedance. This causes higher noise at positions of high output resistance, interacting with the noise current from U1. For example if the effective R_S from the volume control is 10kΩ, a 1.2pA/√Hz noise current from U1 will produce an input referred 12nV/√Hz noise voltage, from this source alone. The Fig. 6-44 driver is a flexible one, and operates at supplies as low as ±10V with outputs up to 3Vrms, with slight distortion increases. With ±5V supplies up to 2Vrms is available, with higher distortion (but still ≤0.01%).

Paralleled Output Line Driver

Often a modest increase in output may be needed for a driver, but circumstances may not warrant the use of additional buffer devices. Figure 6-45 shows how a second section of a dual op amp can be used to provide additional load drive.

In this circuit using an OP275 dual op amp, the U1A section is a gain-of-five voltage amplifier, while the U1B section is a voltage follower, used simply to provide additional current to the load. Current sharing is determined by output summer resistors R_4 and R_5, and the parallel stage drives 600Ω loads with less distortion than a single OP275 section.

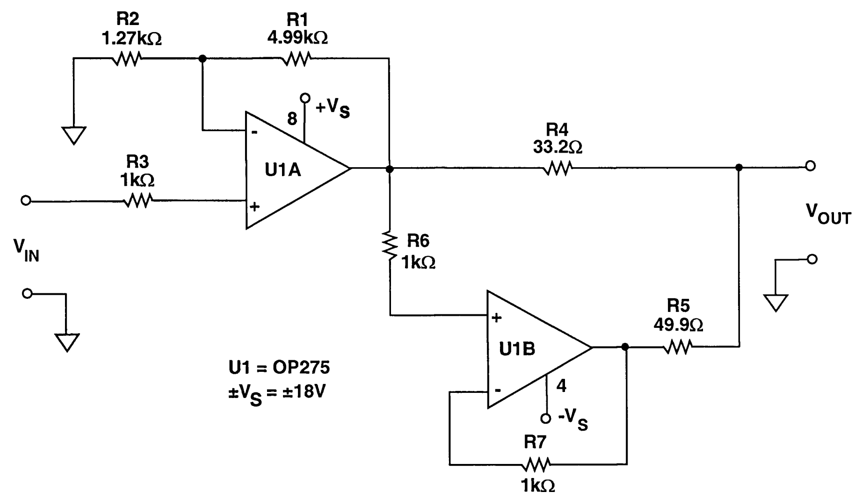


Figure 6-45: *Paralleled output dual op amp line driver*

THD+N performance data is shown in Figure 6-46 below, with the driver operating from ±18V supplies, and for output levels of 1, 2, 5, and 9Vrms into 600Ω.

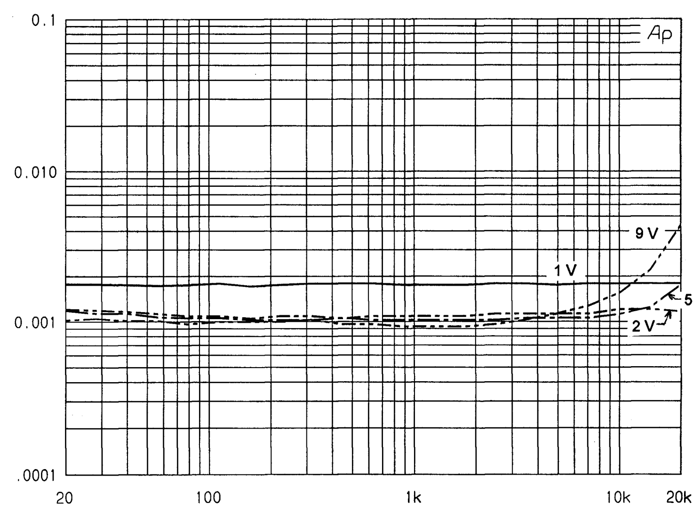


Figure 6-46: *Paralleled output dual op amp line driver, THD+N (%) vs. frequency (Hz), for V_{OUT} = 1, 2, 5, 9Vrms, R_L = 500Ω, V_S = ±18V*

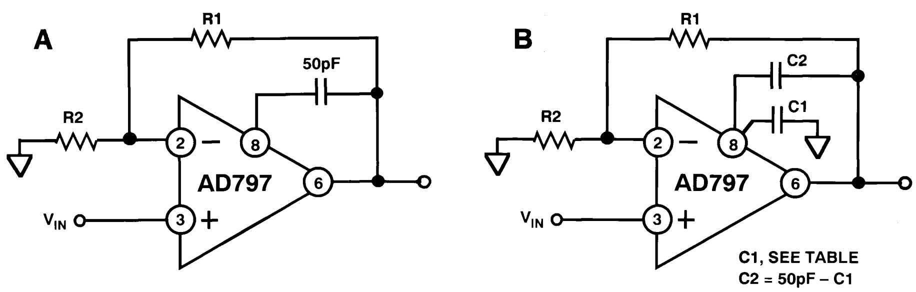
This general scheme can be used with any unity gain stable dual op amp, and also can be adapted for various gain levels, via R_1-R_2. For different devices and/or gains, the ratio of R_4 and R_5 may need adjustment, for lowest distortion into the load.

A Wide Dynamic Range Ultra Low Distortion Driver

Single ended line drivers are simple conceptually, but when pushed to performance limits in dynamic range and distortion, they challenge device choice. The AD797 answers this challenge with its input noise of ≤1nV/√Hz and a distortion canceling output stage. These features allow low and high extremes of dynamic range to be pushed simultaneously.

The AD797 uses a single voltage gain stage, comprised of a folded cascode input combined with a boot strapped current mirror load, allowing the high incremental impedance necessary for a 146dB gain. This buffered single-stage topology is a departure from past devices using multiple stages, with performance benefits in terms of bandwidth, phase margin, settling time, and input noise (see Reference 9).

For standard uses, the AD797 is employed like any 5 pin op amp, such as shown in Figure 6-47 below (neglecting the capacitors for the moment). From the A/B part of the table, relatively low values for resistors R_1-R_2 are recommended for lowest noise. Selecting these resistors should be done with care, since values ≥100Ω will degrade noise performance. Suggested values for gains of G = 10-1000 are noted. The AD797 can drive loads of up to 50mA, and is rated for distortion driving loads of 600Ω.



	A / B		A			B		
	R1 Ω	R2 Ω	C1 pF	C2 pF	3dB BW	C1 pF	C2 pF	3dB BW
G=10	909	100	0	50	6 MHz	0	50	6 MHz
G=100	1k	10	0	50	1 MHz	15	33	1.5 MHz
G=1000	10k	10	0	50	110 kHz	33	15	450 kHz

Figure 6-47: *Recommended AD797 connections for distortion cancellation and/or bandwidth enhancement*

For amplifier applications requiring more top grade performance, optional capacitors C_1 and C_2 can be used. In the Fig. 6-47A configuration, superior performance is realized due to distortion cancellation with the use of a single extra capacitor, enabled simply by adding the 50pF unit as shown. This provides compensation for output stage distortion without effecting the forward gain path, effective over the range of gains noted.

An additional option with the AD797 is the use of *controlled decompensation*, available with the Fig. 6-47B option and the use of capacitors C_1, C_2. At gains of 100 or more, adding C_1 as in column B of the table enhances amplifier open loop bandwidth, allowing very high gain-bandwidths to be achieved— 150MHz at G=100, and 450MHz at G=1000. For high gain operation this extra gain-bandwidth can be very effective.

A family of distortion curves for various AD797 gain configurations driving 600Ω is shown in Figure 6-48 below. As noted, at low frequencies the data is limited by noise, while at high frequencies distortion is measurable, but still extremely low. The distortion for a gain of 10 times at 20kHz for example, is on the order of ≈0.0001%, implying a dynamic range of about 120dB re 3Vrms, or even more for higher level signals.

An additional point worth making at this point is one regarding the AD797's special distortion cancellation ability. Referring to Fig. 6-47A (again), it should be noted that the 50pF capacitor is connected between pins 8 and 6 of the AD797. Pin 6 is of course the output of the device for standard hookups. However, in special situations, even greater output current may be required, and a unity gain buffer amplifier can be added between pin 6 and the load. For example, one of the buffers of Fig. 6-36 or 6-37 could be used to extend output current to ≥100mA.

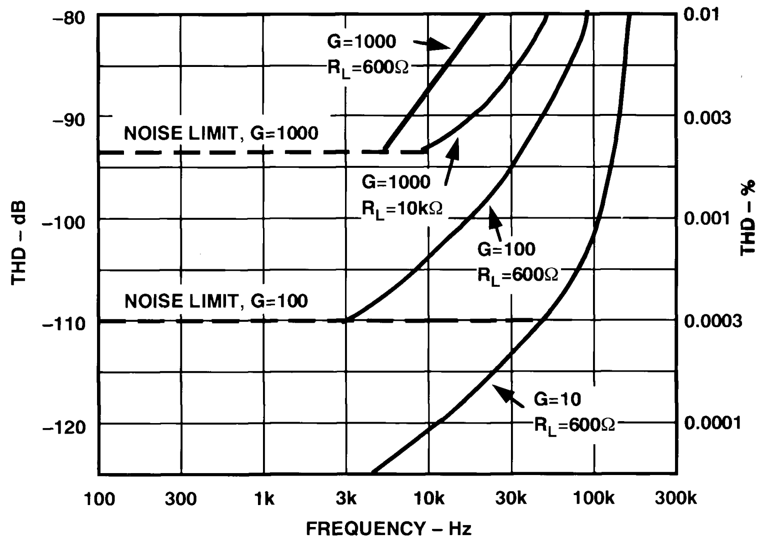


Figure 6-48: *THD vs. frequency at 3Vrms output for AD797 distortion cancellation and/or bandwidth enhancement circuit of Fig. 6-48*

The special point worth noting for this situation is that the 50pF distortion cancellation capacitor should then be connected between the AD797 pin 8 *and the output of the buffer* (not the AD797 pin 6). This step allows the distortion correction to be applied not just to the AD797 internal circuits, but also extends it to include the buffer.

A case in point where this would come to useful purpose lies with the AD797 mic preamp, discussed in some detail earlier in the chapter (Fig. 6-5, again). As mentioned therein, a BUF04 would work well as just such a buffered output option for the AD797 preamp. It would be connected as described above, with the 50pF distortion cancellation capacitor. Details of this are left as a reader exercise (but should even so be obvious).

Current Boosted Buffered Line Drivers

When load drive capability suitable for less than 600Ω in impedance is required, it is most likely outside the output current and/or linearity rating of even the best op amps. For such cases, a current boosted (buffered) driver stage can be used, allowing loads down to as low as 150Ω (or less) to be driven. Another example would a driver for long audio lines, i.e., lines more than several hundred feet in length.

Figure 6-49 below is a high quality current boosted driver example, using an AD845 at U1 as a gain stage and voltage driver, in concert with a unity voltage gain current booster stage, U2. The overall voltage gain is 5 times as shown, but this is easily modifiable via alternate values for R_1 and R_2. In any case, for lowest CM distortion effects, input resistor R_3 should be set equal to $R_1 \parallel R_2$ (this assumes a low impedance source for V_{IN}).

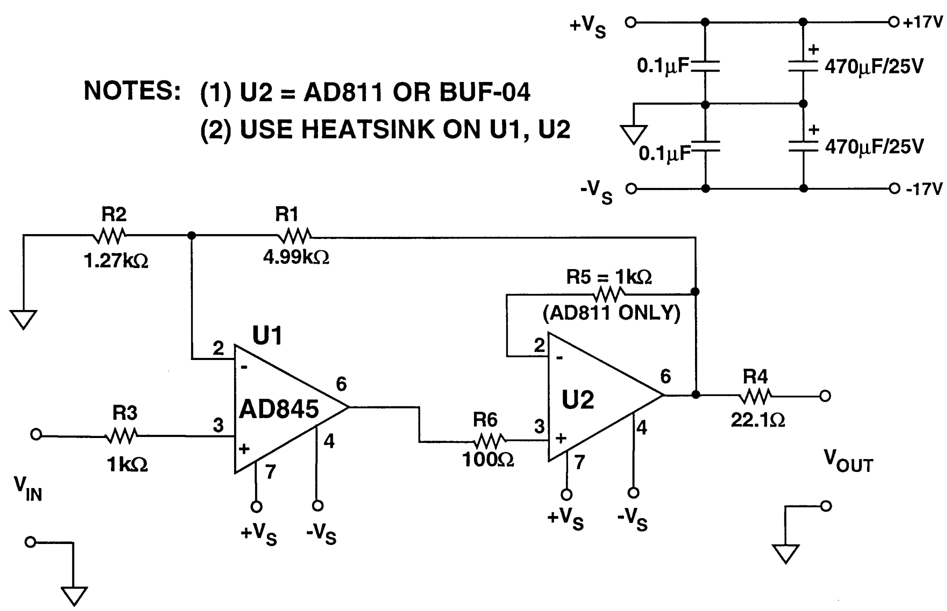


Figure 6-49: *Current boosted line driver*

The amplifier used for U2 can be either the AD811 op amp, or the BUF04 buffer for simplicity. If the AD811 or similar CFB op amp is used (AD812 etc.), it needs to be configured as a follower, with R_5 connected as shown. Since the BUF04 is internally connected as a follower, it doesn't need the R_5 external feedback resistor.

Because of the high internal dissipation of the AD845 and AD811, these devices *must* be used with a heat sink on supplies of ±17V. But, such high supplies are only justified for extreme outputs. Supplies of ±12V also work, and will eliminate need for a heat sink (with lower maximum outputs). In any case, power supplies should be well bypassed.

A special note is applicable here— Always *observe maximum device breakdown voltage ratings within applications. Production versions of this circuit should use supplies of ±17V or less, for 36V(max) rated parts. Similarly, 24V(max) rated parts should use supplies of ±12V or less. In all cases, use only enough supply voltage to achieve low distortion at the maximum required output swing.*

For loads of 150Ω, the output series isolation resistor R_4 is lowered to 22.1Ω to minimize power loss, and to allow levels of 7Vrms or more. The THD+N data for this circuit is shown in Figure 6-51 below, using an AD811 as the U2 buffer. The test conditions are input sweeps resulting in 1, 2, 4 and 8Vrms output, using ±18V power supplies.

For the AD811 operating as U2, the Figure 6-50 data below shows THD+N dominated by noise and residual distortion at nearly all levels and frequencies driving 150Ω, up to 8Vrms. At this level, a slight distortion rise is noted above 10kHz, yet it is still ≈0.001%. With the BUF04 as U2 (not shown) THD+N is comparable at lower output levels, but does show a distortion rise with 8Vrms output at high frequencies (yet still below 0.01%).

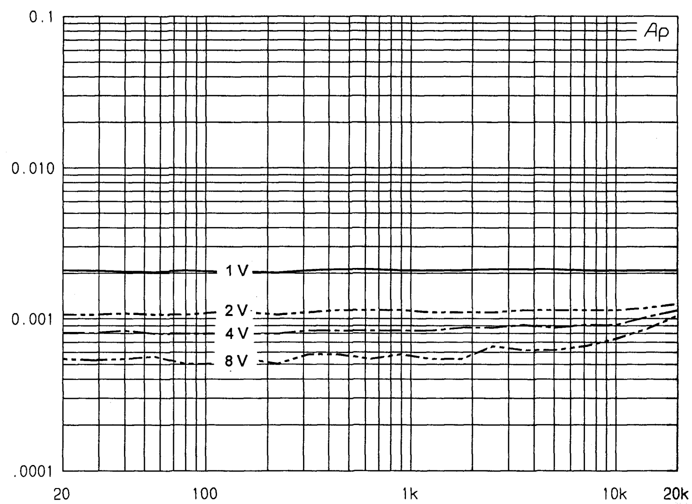


***Figure 6-50:** Current boosted driver of Fig. 6-50 using AD811 as U2, THD+N (%) vs. frequency (Hz), for V_{OUT} = 1, 2, 4, 8Vrms, R_L = 150Ω, V_S = ±18V*

There is a power/performance tradeoff involved with the choice between the two mentioned U2 devices which should be understood. The BUF04 has a standby dissipation of about 200mW on ±15V, while the AD811 is more than double this dissipation, at 500mW. So while the AD811 does yield the lower distortion, it also should be operated more conservatively from a power standpoint. As noted above, only the minimum (±) supply voltage required to sustain a given output should be used with the circuit in general, and particularly with the AD811 employed at U2.

As for U1 in this circuit, other amplifiers can be used, but only with due caution against poor performance. Quite simply, it is difficult to improve upon the AD845's performance in this application. Three possible candidates would include the "Group C" op amps of Fig. 6-43, operating on suitable power supplies; ±17V or less for the AD825, and ±12V or less for the AD8610 and AD8065. Of these, the AD8065 would seem to hold the greatest promise, having shown the lowest wideband distortion in the Fig. 6-43 tests.

But, as considered within the buffered driver circuit of Fig. 6-50, the AD8065 will be operated in an even more linear fashion; that is it is operating essentially unloaded at the output. This is a key factor towards highest performance, as it moves the burden of linear load drive to the buffer stage. Further variations of this circuit technique will be reprised later, within other driver applications to be discussed.

Composite Current Boosted Drivers

Another useful current-boosted circuit technique combines the positive aspects of two different amplifiers into a single composite amp structure, producing a very high performance line driver (see References 10-13 for several variations of this basic circuit). With an FET input IC used as the input stage, DC offset change from source resistance variations of a typical volume control of ≈50kΩ is nil, allowing total direct coupling. As noted previously, with a high current, wide band booster output stage, line impedances down to 150Ω can be driven with excellent linearity.

This type of composite amplifier allows good features of two dissimilar ICs to be exploited; each optimized for the respective input and output tasks. Figure 6-51 shows a low distortion composite amplifier using two op amps ICs with such performance.

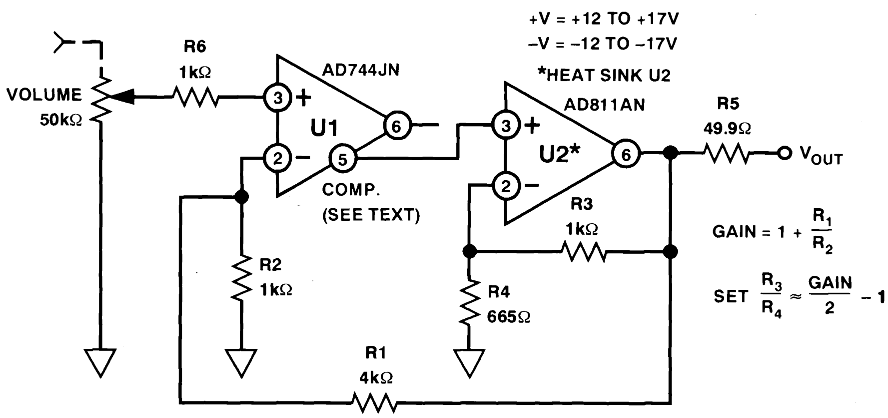


Figure 6-51: *Composite current boosted line driver one*

A factor here aiding performance is that the U1 AD744 stage operates unloaded, and also that the AD744's compensation pin (5) drives U2. This step (unique to the AD744) removes any possibilities of U1 class AB output stage distortion. Another key point is that the overall gain bandwidth and SR of U1 are boosted by a factor equal to the voltage gain of U2, an AD811 op amp, which itself operates at a voltage gain. These factors enhance this circuit by providing both high and linear load current capability, providing a composite equivalent of an FET input power op amp.

This design operates at an *overall* voltage gain set by R_1 and R_2 (just as a conventional non-inverting amplifier) which in this case is 5 times. Since the circuit also uses a local loop around stage U2, the R_3/R_4 ratio setting the U2 stage gain should be selected as noted. This complements the overall gain set by R_1 and R_2, and optimizes loop stability.

Also note that the U2's feedback resistor R_3 has a preferred minimum value for stability purposes (again, as is unique to CFB amplifier types). Here with the AD811, a 1kΩ value suffices, so this value is fixed. R_4 is then chosen for the required U2 stage gain. Further design details are contained in the original references (see References 2 and 10, again).

The composite amplifier performance for a typical audio load of 600Ω, THD+N at output levels of 1, 2, 4 and 8Vrms is shown in Figure 6-52 below, while operating from supplies of ±18V for this test. The apparent distortion is noise or residual limited at almost all levels, rising just slightly at the higher frequencies.

Lower impedance loads can also be driven with this circuit, down to 150Ω. Note that for operating supply voltages of more than ±12V, a clip on heat sink is recommended for U2, as previously discussed for the AD811. Practical versions of this circuit can readily use supplies of ±12V, and still operate very well.

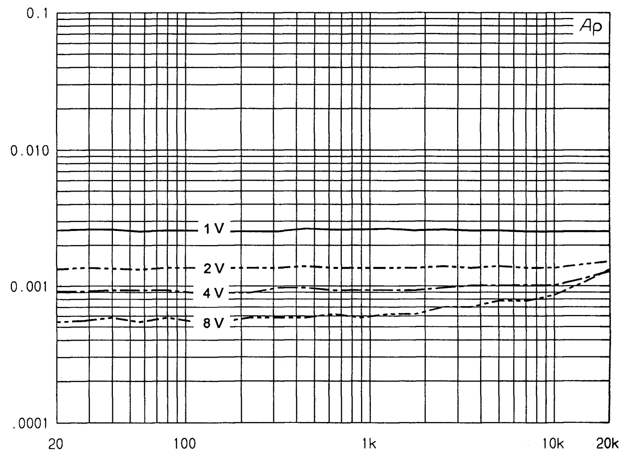


Figure 6-52: *Composite current boosted driver one of Fig. 6-51, THD+N (%) vs. frequency (Hz), for V_{OUT} = 1, 2, 4, 8Vrms, R_L = 600Ω, V_S = ±18V*

The circuit of Fig. 6-51 is a very flexible one, and it can also be adapted a variety of ways. Although the original version shown uses the AD744 compensation pin (5) to drive the output stage U2 device, conventional internally compensated op amps can also be used for U1, and still realize the many features of the architecture.

The ability to adapt the topology to differing devices in single and dual op amp formats allows such dual FET devices as the AD823 to be usefully employed in a stereo realization. Similarly, dual CFB op amps such as the AD812 can be used in the U2 output stage. Thus a complete stereo version of the circuit can be efficiently built, based on only two IC packages.

This topology's flexibility also opens up a diversity of other applications beyond the basic line driver. For example, using a power-packaged dual CFB op amp such as the AD815 for U2, allows very low impedance loads such as headphones to be driven, down to as low as 10Ω (see Reference 12, again).

The composite current boosted line driver two, shown in Figure 6-53 below, summarizes a number of the above mentioned options, and adds some other features as well.

Similarities within this circuit to the predecessor are resistances R_1-R_4, which perform similar functions to the previous version. Overall gain is again calculated via R_1-R_2, while output stage gain is set via R_3, R_4, etc.

Here, note that an additional pair of resistances, R_C and R_D, form a local feedback path around stage U1. This addition allows the effective open loop bandwidth of U1 as it operates within the overall loop to be increased. For the values shown, using an AD823 for U1, the open loop bandwidth is about 100kHz. This means that the open loop bandwidth of the entire circuit is greater than the audio bandwidth, which means phase errors within the passband will be minimized.

An optional small capacitance (C_F, 10-20pF) can be useful for stabilizing the U1 stage, particularly if it employs a wide bandwidth device such as the AD825. When C_F is used, a like capacitor C_{IN} can also be used, to preserve high frequency impedance matching.

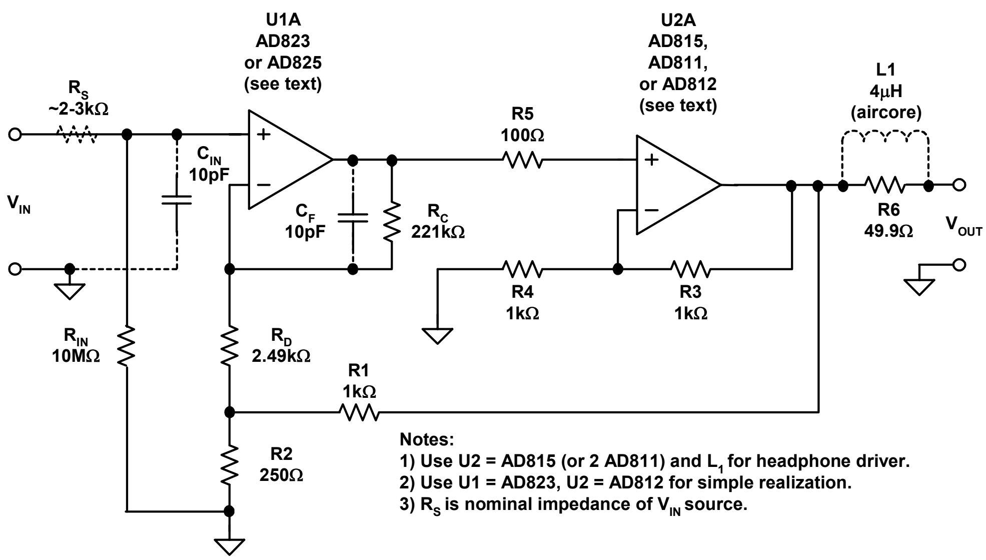


Figure 6-53: *Composite current boosted line driver two*

The primary input impedance balancing of the circuit is accomplished via resistance R_D, which has a dual role. External resistance R_S is the nominal output resistance of a volume control (typical for a 50kΩ audio taper control at listening level). R_D is chosen to match R_S, and R_C will be approximately 100 times the R_D value when using the AD823.

The necessity of inductor output L1 depends upon whether the circuit is to be used with low impedance loads. For headphones, the L1 choke is necessary to prevent excessive voltage loss from a simple R_6 connection. R_6 is used, in either a headphone or line driver case. If configured as a headphone driver, the circuit should use several square inches of PCB area around U2, to heat sink the AD815 device (see device data sheet). The AD811 and AD812 can also be used to drive higher impedance phones, such as 100Ω or more.

Because of the vast number of options with this circuit, no performance is presented here. However, some insight into headphone driver performance is contained in Reference 12.

Differential Line Drivers

Unlike differential line receivers, a standard circuit topology for differential line *drivers* isn't nearly so clear-cut. A variety of different circuit types for driving audio lines in a balanced mode are discussed in this section, with their contrasts in performance and complexity. The virtues of balanced audio line operation are many. The largest and most obvious advantage is the inherent rejection of inevitable system ground noises, between the driver and receiver equipment locations.

There are also more subtle advantages to balanced line operation. Differential drivers tend to inject less noise onto the power supply rails. Related to this, they also produce inherently less noise onto the ground system, since by definition the return path for a differential signal is not ground. This can be a significant advantage when high currents need to be driven into a long audio line, as it can reduce multiple channel crosstalk. The circuits that follow illustrate a variety of methods for differential line driving.

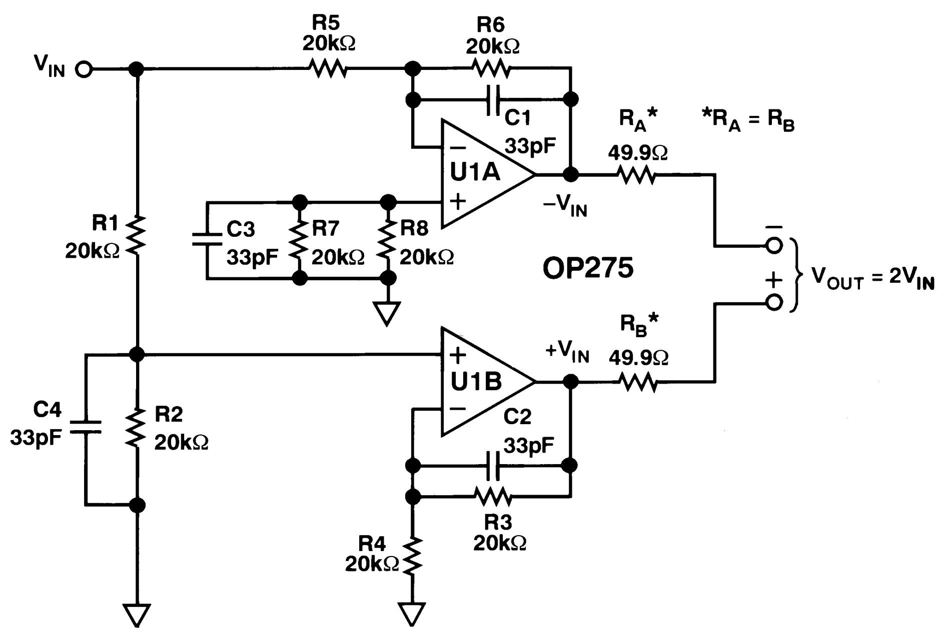


Figure 6-54: *An "inverter-follower" differential line driver*

"Inverter-Follower" Differential Line Driver

A straightforward approach to developing a differential drive signal of $2V_{IN}$ is to amplify in complementary fashion a single-ended input V_{IN}, with equal gain inverter and follower op amp stages. With op amp gains of ±1, this develops outputs $–V_{IN}$ and V_{IN} with respect to common, or $V_{OUT} = 2V_{IN}$ differentially. This "inverter/follower" driver is easily accomplished with a dual op amp such as the OP275, plus an 8x20k film resistor network (or discrete), as shown above, in Figure 6-54. Here U1A provides the gain of –1 channel, while U1B operates at a gain of +1. The differential output signal across the balanced output line is $2V_{IN}$, and the differential output impedance is equal to $R_A + R_B$, or 100Ω. The output resistors $R_A + R_B$ should be well matched, for reasons discussed earlier.

Use of like-value gain resistors around the U1 sections makes the respective channel noise gains match, and also makes their purchase easy. In addition, this forces the source impedances seen by the op amp ± inputs to be matched. Capacitors C_1-C_2 provide a ultrasonic rolloff, and enhance stability into capacitive lines. Overall, this circuit is high in performance for its cost and simplicity. Note that if a resistor network is used for R_1-R_8, the entire circuit can be built with only 8 components.

THD+N performance of the Fig. 6-54 circuit operating on ±18V supplies is shown in Figure 6-55 below, for a series of successive sweeps resulting in output levels of 1, 2, 5 and 10Vrms across 600Ω. The distortion in most instances is about 0.001%, and somewhat higher at a 1V output level (noise limited at this level). Maximum output level is about 12Vrms into 600Ω before clipping (not shown).

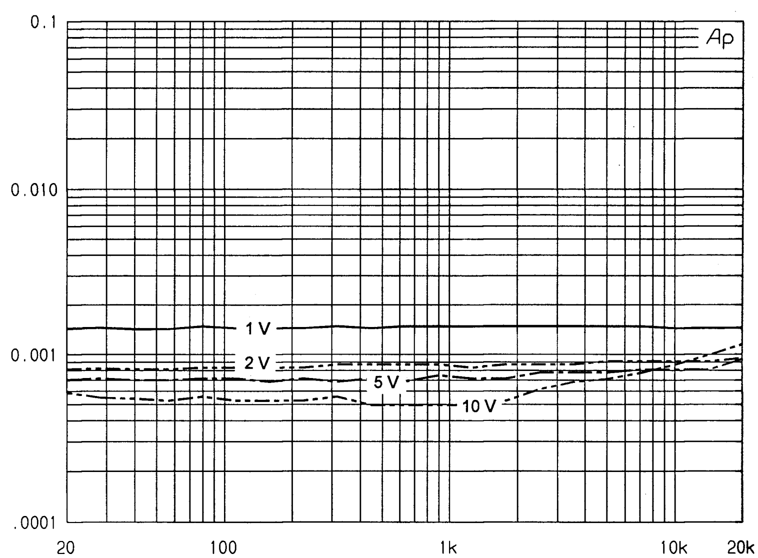


Figure 6-55: *Inverter-follower driver of Fig. 6-54, THD+N (%) vs. frequency (Hz), for V_{OUT} = 1, 2, 5, 10Vrms, R_L = 600Ω, V_S = ±18V*

In system terms, this type of differential line driver can potentially run into application problems, and should be used with some caveats in mind. In reality, this driver circuit uses two mirror-imaged, single-ended drivers, and they produce voltage output signals with respect to the source (V_{IN}) common point.

At the load end of a cable being driven, if the receiver used has a high impedance differential input (such as those discussed in the line receiver section) there is no real problem in application for this driver circuit. However, it should be noted that one side of the differential output from Figure 6-54 *cannot be grounded without side effect*. This is because the source drive V_{OUT} is *not* truly floating, as would be in the case of a transformer winding.

In this sense, the circuit is *pseudo differential*, and it shouldn't be used indiscriminately. Nevertheless, within small and defined systems, is still has the obvious advantage of simplicity, and, as noted, it can achieve high performance. Note also that with the matched source resistances R_A and R_B of 49.9Ω as shown, nothing will be damaged even if one output is shorted— other than a loss of half the signal! Finally, if balanced high impedance differential loading is used at the receiver, there will be no side effects.

Cross-Coupled Differential Line Driver

A more sophisticated form of differential line driver uses a pair of *cross-coupled* op amps with both positive and negative feedback paths. The general form of this type of circuit is a cross-coupled Howland circuit, after the classic resistor bridge based current pump. The cross-coupled form was described by Pontis in a solid-state transformer emulator for high performance instrumentation (see Reference 13).

Application-wise, this configuration provides maximum flexibility, allowing a differential output signal V_{OUT} to be maintained constant and independent of the load common connections. This means that either side can be shorted to common without loss of signal level, i.e., as can be done with a transformer.

Figure 6-56 below shows the SSM2142 balanced line driver IC in an application. The SSM2142 consists of two Howland circuits A2 and A3, cross-coupled as noted, plus an input buffer (A1). The trimmed multiple resistor array and trio of op amps shown is packaged in an 8 pin miniDIP IC with the pinout noted.

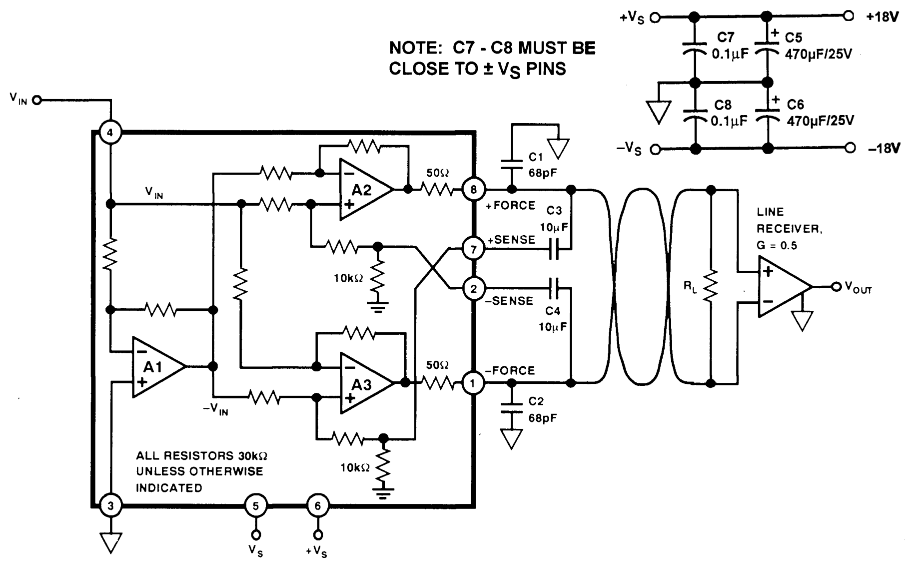


Figure 6-56: *SSM2142 cross-coupled differential line driver used within balanced driver/receiver system*

The SSM2142 line driver is designed for a single-ended to differential gain of 2 working into a 600Ω load. In the simplest use, it is strapped with the respective output FORCE/SENSE pins tied together (7-8, 1-2). Small film capacitors C_1-C_2 preload the IC for stability against varying cable lengths. To decouple line dc offsets, the optional capacitors C_3-C_4 are used as shown, and should be non-polar types, preferably films.

An additional "housekeeping" caveat with the SSM2142 involves the high frequency power supply bypassing. The 0.1μF low inductance bypass caps C_7 and C_8 must be within 0.25" of power supply pins 5 and 6, as noted in the figure. If this bypassing is compromised by long lead lengths, excessive THD will be evident.

In a system application, the SSM2142 is used with a complementary gain of 0.5 receiver, either an SSM2143, or one of the other line receivers discussed previously. The complete hookup of Fig. 6-56 comprises an entire single-ended to differential and back to single-ended transmission system, with noise isolation and a net end-to-end unity gain.

Figure 6-57 shows the THD+N performance of the SSM2142 driver portion of Fig. 6-56, for sweeps yielding output levels of 1, 2, 5 and 10Vrms across 600Ω. While performance is noise limited for the 1V output curve, distortion drops to ≤0.001% and near residual for most higher levels, rising only with higher frequencies and the 10V output curve.

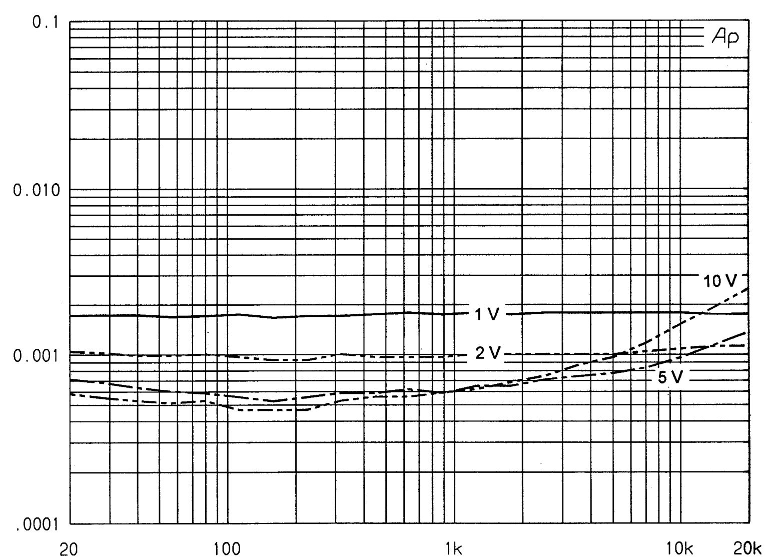


***Figure 6-57:** SSM2142 driver portion of Fig. 6-56, THD+N (%) vs. frequency (Hz), for V_{OUT} = 1, 2, 5, 10Vrms, R_L = 600Ω, V_S = ±18V*

These two differential drivers are suited for 600Ω or higher loads, and, within those constraints, perform well.

As should be obvious, these drivers do *not* offer galvanic isolation, which means that in all applications there must be a DC current path between the grounds of the driver and the final receiver. In practice however this isn't necessarily a problem.

The following circuits illustrate differential drivers that do offer galvanic isolation, and can therefore be used with ground potential differences up to several hundred volts (or the actual voltage breakdown rating of the transformer in use).

Transformer Coupled Line Drivers

Transformers provide a unique method of signal coupling, which is one that allows completely isolated common potentials, i.e., *galvanic isolation*. As noted previously in the line receiver section, transformers are not without their technical and practical limitations, but their singular ability to galvanically isolate grounds maintains a place for them in difficult application areas (see References 15 and 16).

Basic Transformer Coupled Line Driver

The circuit of Figure 6-58 below uses some previously described concepts to form a basic low DC offset, high linearity driver using a high quality nickel core output transformer. U1 and U2 form a high current driver, similar to the Fig. 6-49 current boosted driver.

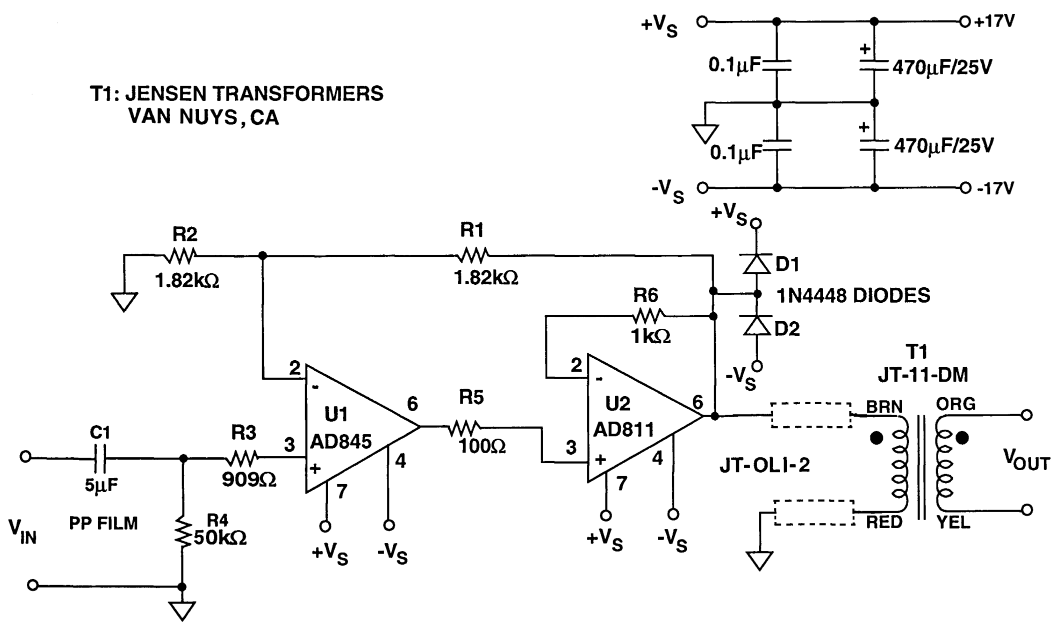


***Figure 6-58:** A basic transformer coupled line driver*

In this circuit U1 is a low offset voltage FET input op amp, for the purpose of holding the DC offset seen at the primary of T1 to a minimum (±12.5mV maximum as shown, typically less). DC current flowing into the primary winding of a transformer should be minimized, for lowest distortion. C_1, a high quality film capacitor, decouples any DC offset present on V_{IN}, for similar reasons.

The U1-U2 device combination is capable of ±100mA or more of output, which aids greatly in the ability of this circuit to drive low impedances. The buffering of U2 is recommended for long lines, or for the absolute lowest distortion. Although T1 is shown with a 1:1 coupling ratio, other winding configurations are possible with transformer variations, that is step-up or step-down, allowing either 600Ω or 150Ω loads.

As can be noted, the T1 primary isn't driven directly, but is isolated by two series isolation devices, Jensen JT-OLI-2s. Each of these is an LR shunt combination of about 39Ω and 3.7µH. The net impedance offers a very low DCR, and an increasing impedance above 1.5MHz for load isolation (see device data sheet and Reference 17). The use of two isolators as shown offers best output CMR rejection for the transformer, but one will also work (with less CMR performance), as will a single 10Ω resistor.

THD+N performance for this driver-transformer combination is shown in Figure 6-59 below, for supplies of ±18V and successive input sweeps, resulting in outputs of 1, 2, 4, and 8Vrms into 600Ω. These data were taken with a single series resistance of 10Ω driving T1 (which could be conservative compared to operation with two isolators).

As with the 2x and 5x basic drivers previously described, these data are essentially distortion free above 100Hz. At lower frequencies there is seen a level dependent, inverse-frequency dependent distortion. The measured distortion reaches a maximum at 20Hz with output levels of 8Vrms (≈20dBm), while at lower levels it is substantially less.

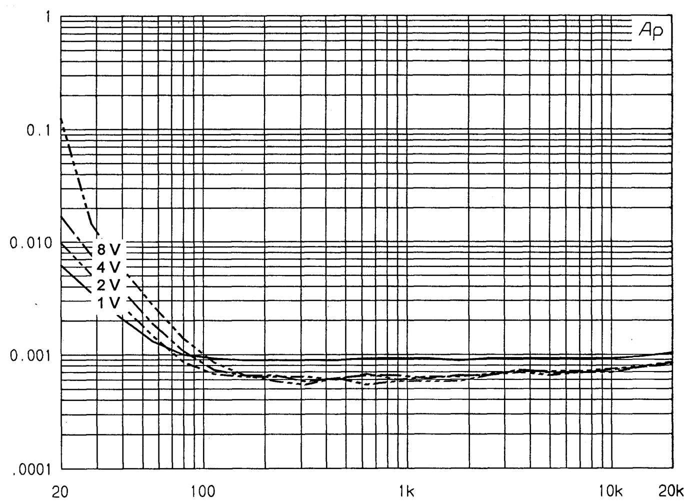


Figure 6-59: *Transformer driver of Fig. 6-58, THD+N (%) vs. frequency (Hz), for V_{OUT} = 1, 2, 4, 8Vrms, R_L = 600Ω, V_S = ±18V*

This distortion phenomenon is basic to audio transformers, to one degree or another. It is lessened (but not totally eliminated) in the higher quality transformer types, such as the nickel-core unit used in the Fig. 6-58 circuit.

In practice, there are some factors that tend to mitigate the seriousness of the low frequency distortion seen in the performance data of Fig. 6-59. First, rarely will maximum audio levels ever be seen at 20Hz. Thus suitably derated operation of T1 will strongly reduce the incidence of this distortion.

However, if the lowest distortion possible independent of level is desired, then some additional effort will need to be expended on making the transformer driver more sophisticated. This can take the form of actively applying feedback around the transformer, so as to lower its non-linearity to negligible levels. This is design approach is discussed with the next driver circuits.

Feedback Transformer Coupled Line Drivers

While non-premium core transformers are more economical than the nickel core types, as a tradeoff they do have much higher distortion. To further complicate the design issue, the distortion characteristics of most transformers varies with level and frequency in complex ways, rising more rapidly at higher levels and lower frequencies. This behavior is even less forgiving than that of the nickel core types, and complicates somewhat the application of audio transformers. While a nickel core transformer has distortion characteristics sufficiently low so as to allow their use without distortion correction (Fig. 6-58, again) the same simply isn't true for other core materials.

A family of distortion curves for another transformer type illustrates this behavior, shown in Figure 6-60 below. This series of plots is for a Lundahl LL1517 silicon iron C core unit, with successive output levels of 0.5, 1, 2 and 5Vrms into a 600Ω load. Individual device samples will vary, but the general pattern is typical of many audio transformers.

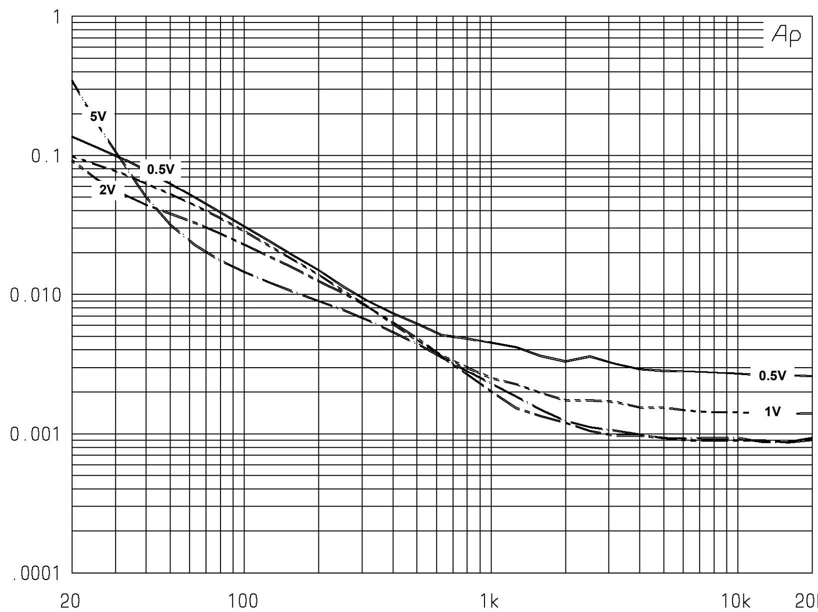


Figure 6-60: *Lundahl LL1517 transformer and driver (without feedback), THD+N (%) vs. frequency (Hz), for V_{OUT} = 0.5, 1, 2, 5Vrms, R_L = 600Ω*

Werner Baudisch (see Reference 18) developed a very effective driver technique for minimization of transformer distortion. The technique involves the use of a drive amplifier, connected to the transformer primary in a direct manner. The amplifier uses conventional negative feedback for gain stabilization. In addition, a primary sensing resistance develops a voltage sample proportional to primary current, and the voltage thus derived is also fed back to the amplifier. This second feedback path is positive feedback, so the arrangement is also known as a *mixed feedback* driver (see Reference 19).

This very useful technique of the mixed feedback driver can be used to advantage to integrate a line driver with the transformer primary within a feedback loop, which cancels the bulk of the objectionable distortion. In practice, with careful driver adjustment it is possible to reduce the distortion of the transformer plus driver almost to *that of the driver stage, operating without the transformer*. The beauty of the principle is that the inherent floating transformer operation is not lost, and is still effectively applied in a highly linear mode. Due to the action of the mixed feedback, the transformer primary resistance is effectively cancelled, thus appreciably lowering the net secondary output impedance.

The circuit of Figure 6-61 below is a basic, single-ended mixed feedback driver using either a Lundahl LL1582 or LL2811 transformer as T1, and an AD845 or an OP275 as the amplifier. These transformers have two 1:1 primaries, as well as two 1:1 secondaries. As used, both primaries are connected in series, and the T1 net voltage transfer is unity.

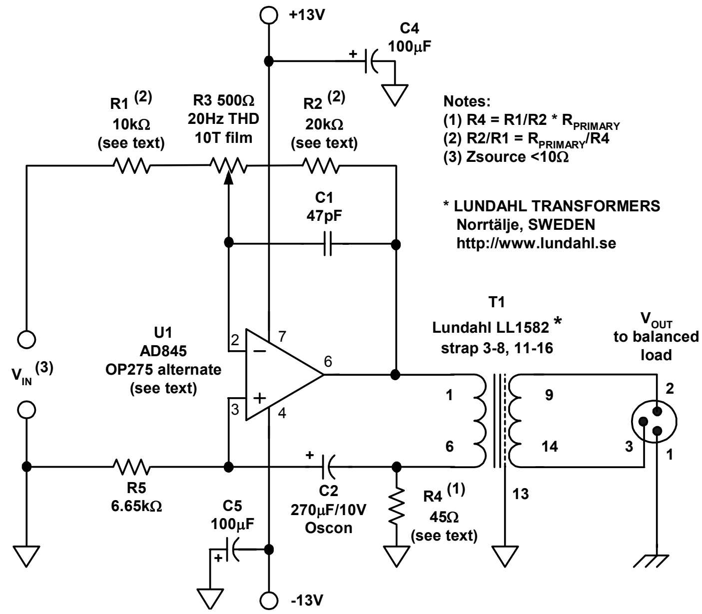


Figure 6-61: *A basic single-ended mixed feedback transformer driver*

To enable correct mixed feedback operation, two key ratios within the circuit must be set to match. One ratio is between the net T1 primary resistance, $R_{PRIMARY}$ and sample resistor R_4, and the other is R_2 and R_1. This relationship is:

$$R_{PRIMARY}/R_4 = R_2/R_1 \qquad \text{Eq. 6-18}$$

It is important to note that $R_{PRIMARY}$ is the total effective DC resistance of T1. As used here, two series 45Ω primaries are used, so $R_{PRIMARY}$ is 90Ω. Gain of the driver circuit is established as in a standard inverter, or the R_2-R_1 ratio. For a gain of 2x, R_2 is then simply 2 times R_1, i.e., 20kΩ and 10kΩ. R_4 may then be selected as:

$$R_4 = R_1/R_2 * R_{PRIMARY} \qquad \text{Eq. 6-19}$$

With the R_1/R_2 ratio of 0.5, this makes R_4 simply ½ $R_{PRIMARY}$, or in this case 45Ω.

Note the value of R_1 is critical, thus the V_{IN} source impedance must be low (<10Ω). This and other subtleties are effective performance keys. One is the sensitivity of the ratio match described by Eq. 6-18. Only when trimmed optimally will the lowest frequency THD be minimum. Thus a multi-turn film trimmer R_3 is used to trim out the various tolerances and the winding resistance of T1. Further, the positive feedback path is AC-coupled via C_2. This provision prevents DC latchup, should positive feedback override the negative. However, a simple time constant of say, 8ms (corresponding to 20Hz) is *not sufficient* for lowest low frequency THD. To counteract this, the C_2-R_5 time constant is set quite long (≈1.8 seconds), which enable lowest possible 20Hz THD. With the suggested AD845 for U1, distortion is lowest, as it is also with an Oscon capacitor for C_2. A larger value ordinary aluminum electrolytic can also be used for C_2, with a penalty of somewhat high distortion. Alternately, an OP275 can also be used for U1 (see below).

With the AD845 FET input op amp used for U1, the maximum DC at the T1 primary is essentially the amplifier Vos times the stage's 3x noise gain, or ≤7.5mV. Since the AD845 can also dissipate ≈250mW, the lowest possible supplies help keep the offset change with temperature as low as possible.

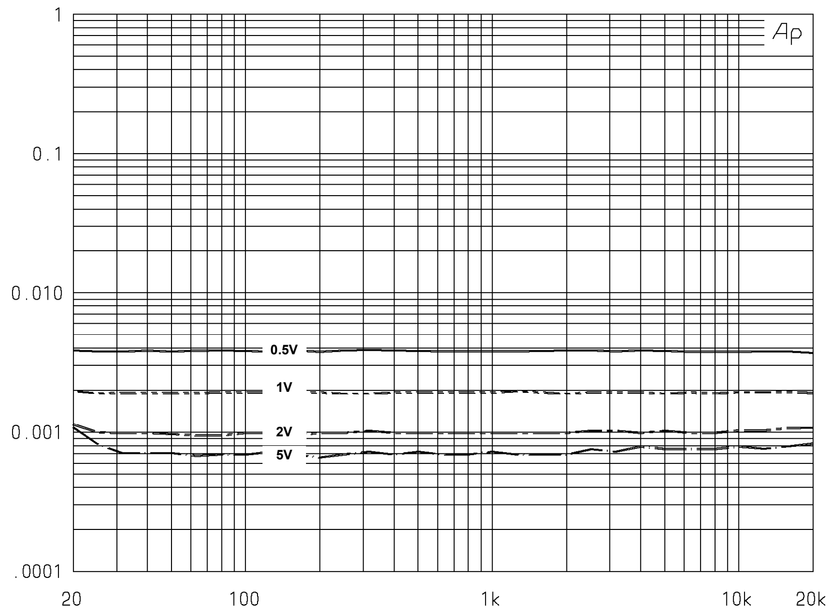


Figure 6-62: *Fig. 6-61 driver with Lundahl LL2811 transformer and AD845, THD+N (%) vs. frequency (Hz), for V_{OUT} = 0.5, 1, 2, 5Vrms, R_L = 600Ω*

Lab THD+N measurements of Fig. 6-61 were made using an LL2811, a transformer like the LL1582, but without a Faraday shield. The two transformers are very similar, but the LL1582 is recommended for single-ended drive circuits. The performance of this feedback driver is shown in Figure 6-62 above, for successive output levels of 0.5, 1, 2, and 5Vrms into a 600Ω load. Comparison of these data with Fig. 6-60 bears out the utility of the distortion reduction; it is decreased by orders of magnitude. More importantly, the level dependence with decreasing frequency is essentially eliminated.

These data do in fact represent almost an ideal THD+N pattern; the distortion level is flat with frequency, and it decreases with increasing output level. An extremely slight increase in THD+N can just be discerned at 20Hz in the 5V curve. The alternate OP275 for U1 also works well, but does have slightly higher distortion (not shown).

Although directly comparable data is not presented for it, it is worth noting that the LL1517 transformer can also be used with the Fig. 6-61 circuit, with R_4 = 9.2Ω, and the two primaries connected in series. However, some additional data on a circuit quite similar to Fig. 6-61 does reveal a potential limitation for this type of driver.

Figure 6-63 below shows a set of high level THD+N curves for a mixed feedback driver using an AD8610 op amp for U1, and the LL1517 transformer. Three THD+N sweeps are made, with the lowest THD+N curve representing the best possible null. The other two curves show increased THD+N at low frequencies, for conditions of R_2/R_1 ratio mismatches of 1 and 5%, respectively. This demonstrates how critical a proper null is towards achieving the lowest possible distortion at the low end of the audio band.

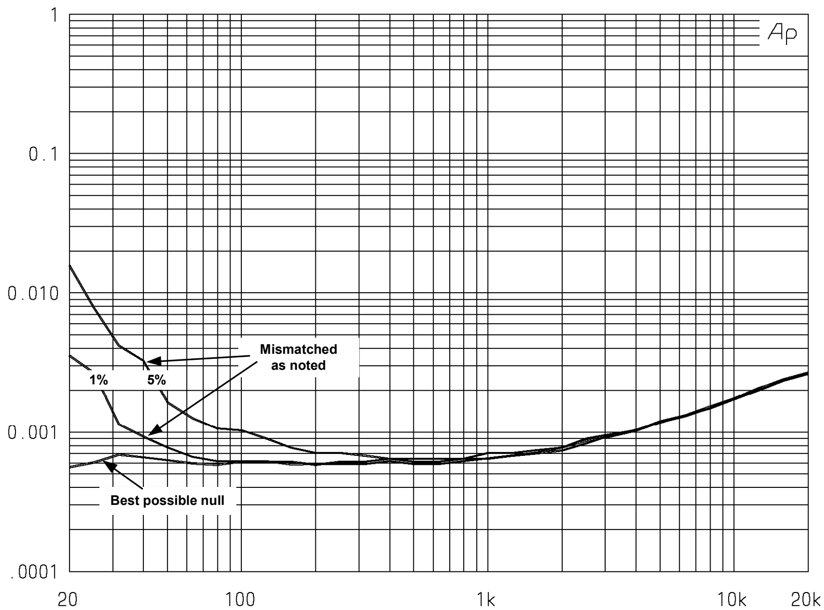


Figure 6-63: *Lundahl LL1517 transformer with mixed feedback AD8610 driver, THD+N (%) vs. frequency (Hz) for various null accuracies*

It is possible to tweak the ratio via R_3 for an excellent 20Hz high level null at room temperature, and this is recommended to get the most from one of these circuits. But, it must also be remembered that the TC of the T1 copper windings is about 0.39%/°C. So, only a 10°C ambient temperature change would be sufficient to degrade the best null by nearly 5%. The resulting performance would then roughly represent the upper curve of Fig. 6-63— still quite good, but just not quite as good as possible in absolute terms.

For the best and most consistent performance, wide temperature range applications of this type of circuit should therefore employ some means of temperature compensation for the copper winding(s) of T1. One means of achieving this would be to employ a thermally sensitive device to track the copper TC of T1. The net goal should be to hold the $R_{PRIMARY}/R_4$ ratio constant over temperature. It should also be noted that for this approach to work, it is assumed that the R_2/R_1 ratio is temperature independent. This is possible with the use of close tolerance, low TC metal film resistors, i.e., 50ppm/°C or better (or the use of a low tracking TC network).

It should also be noted that the *output balance* of an audio transformer is a very important factor when designing audio line drivers. Poor transformer balance can lead to mode conversion of CM signals on the output line (see Reference 20). The result is that a spurious differential mode signal can be created due to poor balance. A transformer can attain good balance (i.e., 60dB or better) by the use of sophisticated winding techniques, or the use of a Faraday shield, as is true in the case of the LL1582 and the LL1517.

Transformer drivers can of course also be operated in a balanced drive fashion. This has the advantage of doubling the available drive voltage for given supply voltages, plus lowering the distortion produced. Mixed feedback principles can be extended to a balanced arrangement, which lowers the distortion in the same manner as for the single-ended circuit just described. An example circuit is shown in Figure 6-64, below.

In Fig. 6-64, a U1-U2 low distortion op amp pair drive an LL1517 transformer. U1 is an inverting gain circuit as defined by gain resistors R_1-R_2, which drives the top of T1. Placed in series with the T1 primary, R_3 acts as a current sampling resistor, and develops a correction voltage to drive the second inverter, U2, through R_4.

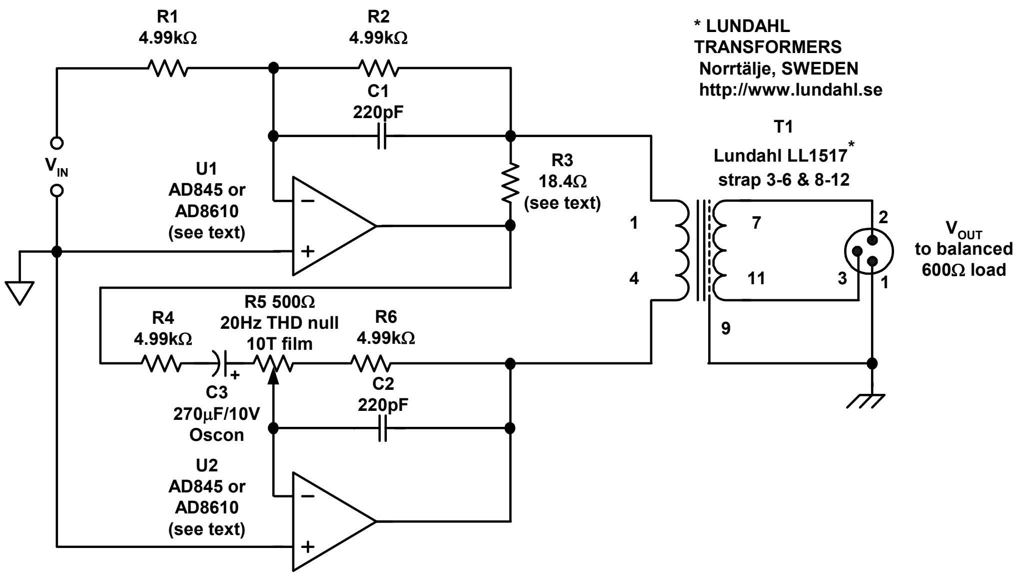


Figure 6-64: *A balanced transformer driver circuit that applies mixed feedback principles of distortion minimization*

This scheme is adapted from the mixed feedback balanced driver circuit of Arne Offenberg (see Reference 21). There are however two main differences in this version. One is operation of the U1 stage as an inverter, which eliminates any CM distortion effects in U1, and the second being the ability to easily set the overall driver gain via R_1-R_2. Within this circuit, it should be noted that the resistances R_1-R_2 do *not* affect the distortion null (as they do in the simpler circuit of Fig. 6-61).

The distortion null in this form of the circuit occurs when the ratios $R_3/R_{PRIMARY}$ and R_4/R_6 match. For simplicity, the second inverter gain is set to unity, so R_3 is selected as:

$$R_3 = R_4/R_6 * R_{PRIMARY} \qquad \text{Eq. 6-20}$$

For the R_4-R_6 values shown, R_3 then is simply equal to $R_{PRIMARY}$, or 18.4Ω when used with the LL1517 transformer with series connected primaries.

As with any of these driver circuits, the exact op amp selection has a great bearing on final performance. Within a circuit using two amplifiers, dual devices are obviously attractive. The distortion testing below discusses amplifier options.

THD+N performance data for the balanced transformer driver of Fig. 6-64 is shown in Figure 6-65 below, using an LL1517 transformer with successive output levels of 0.5, 1, 2, and 5Vrms into a 600Ω load. For these tests, the supply voltages were ±13V, and the U1-U2 op amp test devices were pairs of either the AD8610 or the AD845 (note— two AD8610 singles are comparable to a single AD8620 dual).

An interesting thing about these plots are the fact that the THD+N is both low and essentially unchanged with frequency, which is again, near ideal. Low frequency nulling of the distortion is almost as critical in this circuit as in the previous, and a slight upturn in THD+N can be seen at 20Hz, for the highest level (5V).

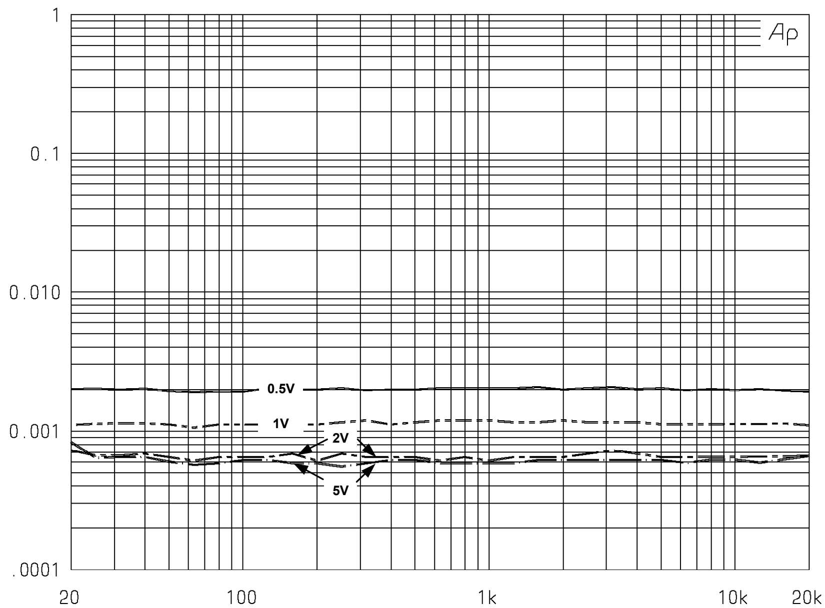


Figure 6-65: *Fig. 6-64 balanced driver with Lundahl LL1517 transformer and two AD8610s, THD+N (%) vs. frequency (Hz), for V_{OUT} = 0.5, 1, 2, 5Vrms, R_L = 600Ω*

For the AD8610 devices shown by these data, the wideband THD+N was slightly lower than a comparable test with the AD845 pair (the latter not shown). Both amplifier sets show essentially flat THD+N vs. frequency characteristics.

Because of the balanced drive nature of this circuit, the realization offers lower distortion than the simpler single-ended version of Fig. 6-61, plus a buffering of the distortion null sensitivity against the input source impedance and gain adjustment. It can thus be considered a more robust method of transformer distortion minimization. For these reasons, the balanced form of driver is recommended for professional or other high performance requirements. Note however that similar caveats do apply with regard to stabilizing the distortion null against temperature.

This driver can be used with the LL1517 and many other transformers, with of course an appropriate choice of R_3. Note that the performance data above reflects use of the op amps operating unbuffered. For very low impedance loads and/or long lines, a pair of the previously described unity gain buffers should be considered, and both the U1 and U2 stages operated with output buffering. This will allow the retention of THD+N performance as is shown in Fig. 6-65 above, but in the face of more difficult loads.

REFERENCES: AUDIO LINE DRIVERS

1. Walt Jung, "Op Amps in Line-Driver and Receiver Circuits, Part 1," **Analog Dialogue**, 26-2, 1992.

2. W. Jung, A. Garcia, "Op Amps in Line-Driver and Receiver Circuits, Part 2," **Analog Dialogue**, 27-1, 1993.

3. Walt Jung, "Walt's Tools & Tips: 'Op Amp Audio: Buffers Part I'" **Electronic Design**, September 1, 1998, pp. 165, 166.

4. Walt Jung, "Walt's Tools & Tips: 'Op Amp Audio: Buffers Part II'" **Electronic Design**, October 1, 1998, pp. 103, 104.

5. Paul Brokaw, "An IC Amplifer User's Guide to Decoupling, Grounding, And Making Things Go Right For a Change," **Analog Devices AN202**.

6. P. Brokaw, J. Barrow, "Grounding for Low and High Frequency Circuits," **Analog Dialogue**, 23-3 (1989).

7. Alan Rich, "Shielding and Guarding," **Analog Dialogue**, 17-1 (1983).

8. Scott Wurcer, "Input Impedance Compensation," discussion within **AD743 data sheet**, Analog Devices.

9. Scott Wurcer, "An Operational Amplifier Architecture With a Single Gain Stage and Distortion Cancellation," presented at **92nd Audio Engineering Society Convention**, March 1992, preprint #3231.

10. W. Jung, S. Wurcer, 'A High Performance Audio Composite Line Stage' within "Applications for Amplifiers in Audio," Ch. 5 in W. Kester, Editor, **1992 Amplifier Applications Guide**, Analog Devices, Inc., Norwood, MA, 1992, ISBN 0-916550-10-9.

11. Walt Jung, 'High Performance Audio Stages Using Transimpedance Amplifiers', within Gary Galo, "POOGE-5: Rite of Passage for the DAC960," **The Audio Amateur**, issue 2, 1992.

12. Walt Jung, "Composite Line Driver with Low Distortion" **Electronic Design Analog Special Issue**, June 24, 1996, pp. 78.

13. Walt Jung, "Walt's Tools & Tips: 'Op Amp Audio: Minimizing Input errors'" **Electronic Design**, December 14, 1998, pp. 80-82.

14. George Pontis, "Floating a Source Output," **HP Journal**, August 1980.

15. "Transformer Application Notes (various)," Jensen Transformers, 7135 Hayvenhurst Avenue, Van Nuys, CA, 91406, (213) 876-0059.

16. Bruce Hofer, "Transformers in Audio Design," **Sound & Video Contractor**, March 15, 1986.

17. Deane Jensen, "Some Tips on Stabilizing Operational Amplifiers," Jensen Transformers AN-001, 7135 Hayvenhurst Avenue, Van Nuys, CA, 91406, (213) 876-0059.

18. Werner Baudisch, "Schaltungsanordnung mit Verstärker mit Ausgangsübartrager," **German patent DE2901567**, issued July 24, 1980.

19. Per Lundahl, "Mixed Feedback Drive Circuits For Audio Output Transformers," Lundahl Transformers, Norrtälje, Sweden, **http://www.lundahl.se**

20. Per Lundahl, "Winding Arrangements of Output Transformers," Lundahl Transformers, Norrtälje, Sweden, **http://www.lundahl.se**

21. Arne Offenberg, "Mixed Feedback Balanced Driver Circuit," LL2811 Audio Output Transformer data sheet, Lundahl Transformers, Norrtälje, Sweden, **http://www.lundahl.se**

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