Interconnections in Audio

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Definition

cable: a potential source of trouble interconnecting two other potential sources of trouble.

Introduction

The life of an audiophile was never easy, but in the good old days of non-portable equipment and analogue power supplies at least you didn't have to worry about interconnecting cables — that is, if you didn't belong to the "Golden Ears" community.

These days, with the ubiquitous laptop PC being the *de facto* standard signal source, and often also the signal recorder, that peace of mind has gone into a virtual reality, and most of us are now forced to relearn the well proven practices of studio audio work. As long as you stay in the digital domain, probably using S/Pdif or TosLink, there is a fair chance that all will be well, maybe only apart from the infamous L-bit defining your license to copy the material. However, once in the analogue domain, you may be practically sure of having to deal with all sorts of interference breaking into the audio signal path.

Ordinarily the trouble is caused by the switching power supply of one or both signal processing units on either side of the connection, with a spectrum from the mains frequency upward. Often the offending unit is floating, and sometimes shorting its case to ground helps. However in the majority of cases the grounding will only reduce the interference, not eliminate it. And in other cases it may even worsen the situation.

A common solution is to put a small audio transformer in the signal path. This breaks the interfering loop, but for a small parasitic capacitance across the transformer, so the passing interference band will hopefully be above the audio spectrum. This may then be eliminated by either filtering the signal, or by placing a cylindrical ferrite bead (like the one found on PC monitor cables) around one or both ends of the cable, if the dominant disturbance is a common mode signal.

Sometimes placing a pair of series inductances coupled magnetically with a ferrite toroidal core (thus forming a "balanced-unbalanced" or "balun" connection), aided by capacitors at each side, may reduce the common-mode interference.

Another way of reducing the interfering signal is to place a small resistance in the ground path inside one of the connectors, increasing its resistance to something between 10 and $100\,\Omega$, thus reducing the loop circulating current, and making the problem tolerable in some instances.

For a serious audiophile none of this is satisfactory. The audio transformer may eliminate the interference, but you have invested heavily in equipment having lots of zeros before the first significant distortion digit, so you are probably not willing to tolerate more than a ppm of spurious air pressure in your ears caused by the transformer's hysteresis. Therefore an active solution is sought.

It is beyond the scope of this little article to discuss all the possible circuits, this has been done elsewhere extensively, see the references. Assuming that you don't want to violate the warranty of the rest of your equipment, your only option is to make something in between. So we shall discuss a couple of simple differential to single-ended receivers, a couple of single-ended to differential converters, and a couple of fully-differential line drivers.

Single-Ended Line Receivers with CMR

The preferred approach is often to KISS (*keep it simple & stupid*). If you think you'll never have to interconnect equipment which requires fully differential signaling, you may want to consider a single opamp solution to common mode interference rejection.

In **Fig.1** the signal source, capable of driving a $600\,\Omega$ line, is being powered by a switching supply, which generates a ground loop current i_g via the transformer's primary-to-secondary stray capacitance C_c. This current consists of short pulses modulated in width following the available mains voltage, so the interference spectrum starts from the mains frequency upward.

For a moment imagine the resistors R_b and R_d being of zero Ohms. The interference voltage v_{PS} is usually independent of the impedance it is driving, so the ground loop current is determined by the stray capacitance impedance $1/j\omega C_c$ in series with a small cable shield resistance. Because the ground current is relatively high, the error voltage v_e at the driving side of the cable will be higher than at the receiving side, and the difference will sum up with the signal voltage v_s. The amplifier will then sense this sum of the signal and the error voltage.

By making R_b and R_d much higher than the cable shield resistance, but still of low value, say $100\,\Omega$, the ground loop current i_g is greatly reduced, and as a consequence the error voltage v_e at both sides of the cable is essentially equal. As the amplifier is now configured like an ordinary differential amplifier, the remaining error becomes a common-mode voltage and is reduced by the common-mode rejection capability of the amplifier, though lowered somewhat by the value tolerances of the four resistors. Nevertheless the error voltage is now suppressed by a large factor, usually enough to be imperceptible.

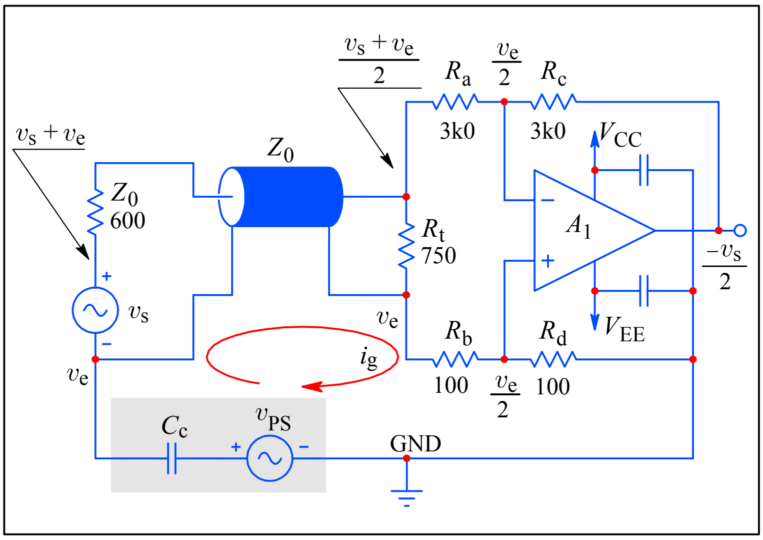

Fig.1: A single opamp line receiver with common-mode interference rejection.

It is a good practice to terminate the cable at both ends by its own characteristic impedance. In audio, the standard $600\,\Omega$ was inherited from the analogue telephone era, and later the microphone technology and the transformer signal coupling. Since all professional grad equipment and even some top commercial grade equipment still rely on that standard, it can hardly be ignored. Here at the amplifier input the termination resistor R_t should be such that $R_t \| R_a$ gives the required $600\,\Omega$. A suitable choice is achieved by making $R_a = 3\text{k}0$ and $R_t = 750\,\Omega$.

Because of the $600\,\Omega$ termination half of the signal voltage is lost. This is normal, but if necessary, it is always possible to double the value of R_c and R_d, making the system gain equal to 2, thus compensating the loss.

When the cable termination is not an issue, and if the signal level is rather high, it is possible to employ a non-inverting configuration, with optional gain, as shown in **Fig.2**. Provided that the input resistance is much higher than the source impedance, the circuit will suppress the error voltage v_e, and amplify only the signal v_s.

The ground resistor R_g provides the DC reference level when the cable is disconnected. The value of R_g is not important, as long as it is much larger than the resistance of the cable shield, a condition met easily in most occasions. However, to minimize the noise gain use R_g larger than R_f, so that the circuit gain approaches unity.

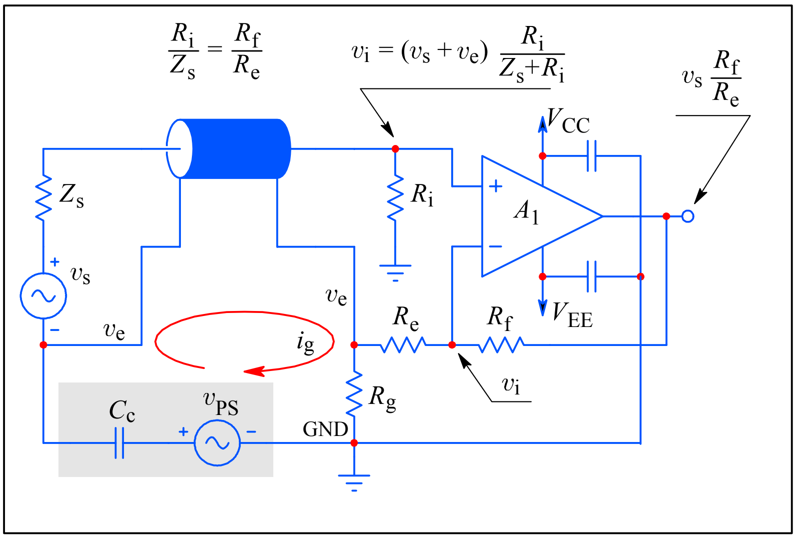

Fig.2: Non-inverting configuration. If $R_i \gg Z_s$, the feedback is referenced to the cable shield, subtracting the error voltage. The ground resistor R_g is necessary to provide a DC reference level when the cable is disconnected, and it can be of any suitable value, preferably between 1 kΩ and 1 MΩ.

The circuits of **Fig.1** and **Fig.2** are not ideal, but are simple to build for a quick fix. If the interference level is not severe or its spectrum not too broad these circuity may provide adequate suppression in most situations. However it is important to keep the ground loop current i_g low.

Component Tolerances & CMR:

Suppose we take the resistors of 1% tolerance and we are so unlucky to select the following values, which unbalance each resistor of the differential amplifier in **Fig.1** by 1%, but in the opposite direction:

$$R_a = 3030\,\Omega,\ R_c = 2970\,\Omega,\ R_b = 99\,\Omega,\ R_d = 101\,\Omega$$

The common-mode gain, with the input shorted, $R_t = 0$, is:

$$A_{cm} = \frac{v_o}{v_e} = \frac{R_d}{R_b + R_d} \cdot \frac{R_c + R_a}{R_a} - \frac{R_c}{R_a} = 0.0198 \tag{1}$$

For the same conditions, but with the input impedance matching, $R_t = 750\,\Omega$, $Z_0 = 600\,\Omega$:

$$A_{\mathrm{cm}} = \frac{v_{\mathrm{o}}}{v_{\mathrm{e}}} = \frac{R_{\mathrm{d}}}{R_{\mathrm{b}} + R_{\mathrm{d}}} \cdot \frac{R_{\mathrm{c}} + R_{\mathrm{a}}}{R_{\mathrm{a}} + R_{\mathrm{t}}||Z_0} - \frac{R_{\mathrm{c}}}{R_{\mathrm{a}} + R_{\mathrm{t}}||Z_0} = 0.0178 \tag{2}$$

These are the worst case conditions. Because of the subtraction, the common-mode gain is approximately only one half of the sum of the tolerances, or $\approx 2\%$. This means that the error signal will be attenuated by at least 34dB. By selecting the resistor pairs to match to ~0.1% an attenuation of >56dB is easily achieved. A similar calculation holds for other circuits.

Differential to Single-Ended Line Receiver

Most professional signal sources offer a balanced differential output. In **Fig.3** a balanced line receiver is shown. It uses an instrumentation amplifier configuration with some high frequency filtering.

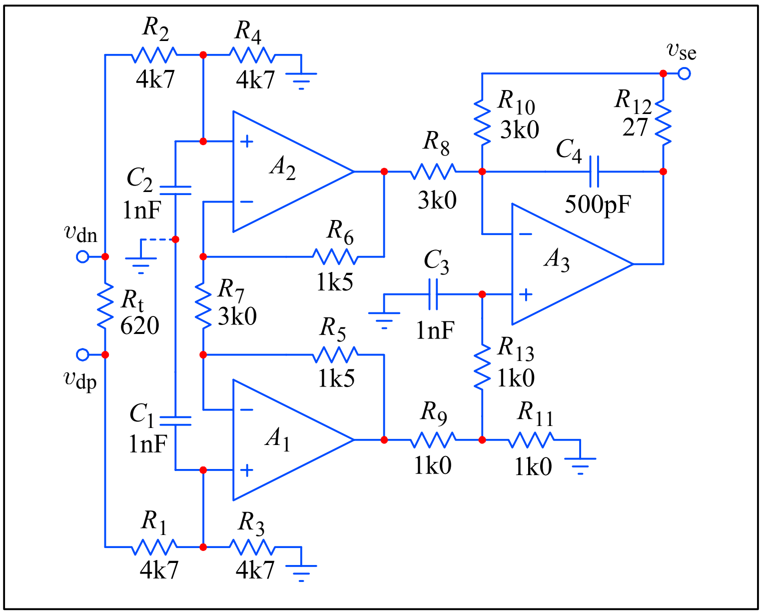

Fig.3: Balanced differential line receiver to single-ended converter.

The correct line termination is achieved by R_{t} of $620\,\Omega$ paralleling the four amplifier input 4k7 resistors R_{1-4}. This reduces the signal and the common-mode error by 2, increasing the input signal handling range, so it is limited only by the amplifier's output range. The gain around the amplifiers A_1 and A_2 compensates this, and if necessary the resistor R_7 can be halved to compensate for the line termination loss.

The input filter capacitors C_1 and C_2 can be either tied to ground, or they can be replaced by a single capacitor of half the value, or you can use two 500pF capacitors connected to ground and a third 250pF capacitor connected between the A_{1-2} inputs, whichever suits your needs in a particular situation. The resistor arrangement at the A_3 input has been chosen to give equal loading to A_1 and A_2, as well as equal filtering by C_{3-4}. The small output resistance R_{12} decouples the A_3 output from large capacitive loading; it can be shorted if the following stage is easy to drive.

This circuit can accept either a single-ended or a differential signal source. In the differential case it is advantageous to connect the shield of the differential line to ground, either the driving or the receiving end, but not both. Such a break in the ground loop reduces inductive interference from external sources.

In **Fig.4** a similar circuit is shown, however here the common-mode signal is sensed by a separate non-inverting input via the R_{5-6} divider. The common-mode signal is being reduced by both the A_{1-2} stage and the A_3 stage, increasing the rejection. It is important that the signal input and the common-mode input are filtered equally, if good common-mode rejection is expected up to and beyond the system's bandwidth. If the differential cable shield is connected to the v_{cm} input, it must also be connected directly to the driving circuit ground. Otherwise, a third line can be used for this connection.

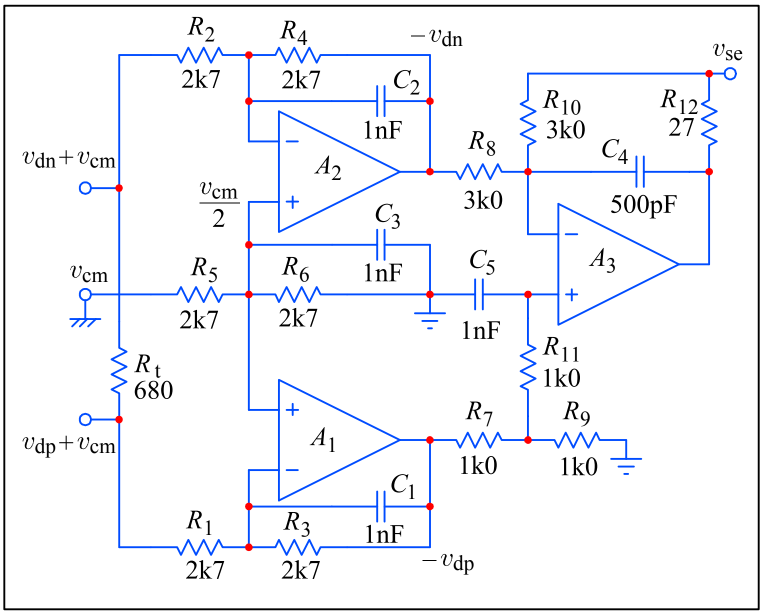

Fig.4: Balanced differential line receiver with separate common-mode sensing.

Single-Ended to Differential Conversion

It is often necessary to convert a single-ended signal to a differential one, and a simple circuit in **Fig.5** can be used for this task. The circuit uses a single filtered divider, making the two complementary output signals equal in magnitude and phase up and beyond the desired signal bandwidth.

If this circuit will be used to drive a 600 Ω line, two 300 Ω resistors should be put at the amplifier's outputs. However, this circuit can not cope with unbalanced loads or loads which are not floating with respect to both the driving circuit ground and the receiving circuit ground. Any such load imbalance will result in signal amplitude reduction, and if the load is reactive also in a degradation in the frequency response. Consequently this circuit is recommended only as an internal converter driving a predictably behaving load.

If necessary, it is possible to modify this circuit for a gain of, say, 2×, by doubling the value of R_3, and inserting a resistive divider in the feedback of A_1 using, say, another pair of 3k0 resistors.

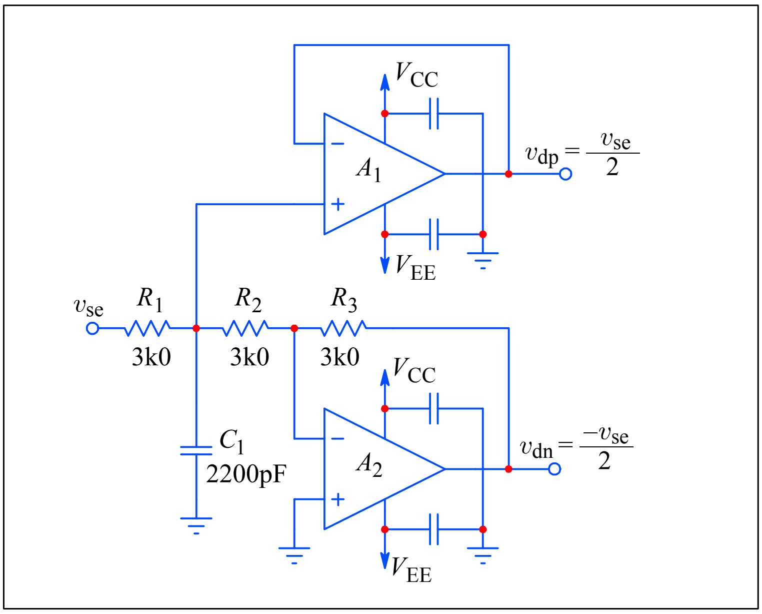

Fig.5: Simple single-ended to differential converter.

As a general rule, it is a good practice to keep the resistances low in order to keep the circuit noise low. Observe however that the majority of opamps on the market, which are capable of driving a 600 Ω load, won't be able to drive anything much lower, so the feedback resistances should be kept some 5× to 10× higher in order to allow maximum output driving capability.

In contrast with the simplicity of the circuit in **Fig.5**, the circuit in **Fig.6** employs a cross-feedback configuration, improving the adaptability to less than optimally balanced loads, though not by much. The main problem with all such circuits arises when the overall cross-feedback gain is close to unity (or higher). Because of the effectively positive feedback applied via R_{5-6}, such circuits are prone to oscillation.

The problem persists even if the negative feedback resistors R_{3-4}, are bridged by a small capacitance, in fact this may make the matters worse, because of the phase shift imbalance caused by the tolerance of the capacitor value. Bear in mind that by adding these capacitors the positive feedback loop critical phase shift is merely transferred to a lower frequency.

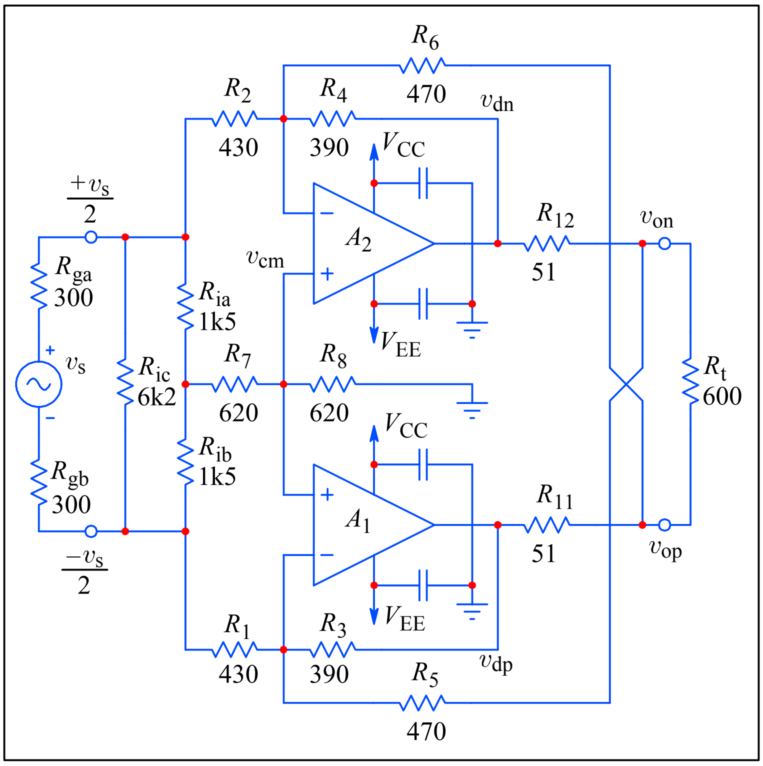

Fig.6: Differential amplifier with cross-feedback for greater dynamic range.

The main purpose of the cross-feedback is to allow the output resistors be smaller that the required 2×300 Ω, and still achieve an effective total 600 Ω output impedance. In this way the circuit can handle almost twice the output signal range and common mode range before clipping. This makes the circuit desirable for low voltage systems, where the signal range is only slightly lower than the supply voltage. However, with a reactive load, as in some ADC inputs, make sure that the positive feedback never comes close to unity, by making a suitable choice of other components.

The input common mode is sensed by $R_{\mathrm{ia-ib}}$ and R_{7-8}, driving the positive input opamp terminals. The divider attenuation should match the opamp gain in order to cancel the common mode substantially.

The calculation of components for the effective output impedance is simple: given the values of the desired termination $Z_\mathrm{t} = R_\mathrm{t}$, the feedback resistor R_3, and the output resistor R_{11}, it is possible to calculate the positive feedback resistor R_5:

$$R_5 = \frac{R_3 R_\mathrm{t}}{R_\mathrm{t} - 2R_{11}} \tag{3}$$

Then from the desired gain A it is possible to determine the input resistance R_1:

$$R_1 = \frac{R_3}{A} \cdot \frac{R_5 R_\mathrm{t}}{2R_5 R_{11} + R_{11} R_\mathrm{t} + R_5 R_\mathrm{t} - R_\mathrm{t} R_3} \tag{4}$$

Of course, the same values hold for the appropriate resistors in the other circuit arm.

Fully Differential Amplifiers

There is a large number of fully differential amplifiers on the market. Most of them are in the form of differential line drivers/receivers, some with a separate common-mode input, some with the ability to accept either a differential or a single-ended signal. They usually handle well any common-mode errors from DC to ultrasound frequencies, but few of them are capable of handling the load impedance imbalance.

Load imbalance is not a feature that anybody would be particularly happy with, but there are occasions where it cannot be avoided. This is where coupling transformers excel, if only the total differential impedance is equal to the required characteristic impedance. So a natural requirement would be an active circuit which behaves just like a transformer, yet without hysteresis. **Fig.7** shows one such option.

The input signal must be supplied by a driver capable of handling a 600 Ω load. The load impedance imbalance and the common-mode error are sensed by A_3, and the amplifiers A_{1-2} provide the necessary compensation of the driving currents, so that the output differential voltage is correctly established. The circuit operation is very simple, the amplifiers A_{1-2} take one half of the input voltage, amplify it by two, and subtract the output common-mode error provided by A_3. The input and output signals are mixed passively via 600 Ω resistors, providing a 300 Ω driving impedance in each arm.

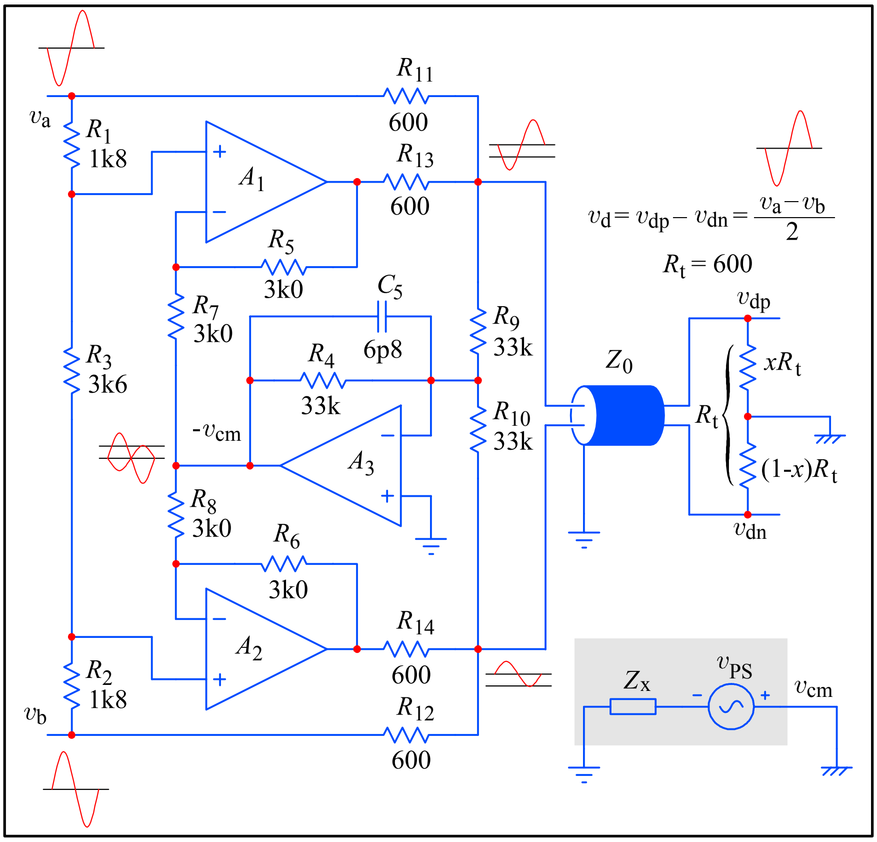

Fig.7: Fully-differential line driver with common-mode rejection and load impedance imbalance compensation. The inputs must be driven by amplifiers capable of handling a 600 Ω load.

The driving stage providing the input signal for the circuit in **Fig.7** would be a good place to implement an active differential filter. The preferred configuration is described in the following chapter.

Differential Filters

In contrast with ordinary filtering circuits, a differential filter should be designed to preserve the common-mode signal equally in both arms, both in amplitude and phase, if the error is to be effectively removed by the successive circuitry. **Fig.8** shows the preferred configuration.

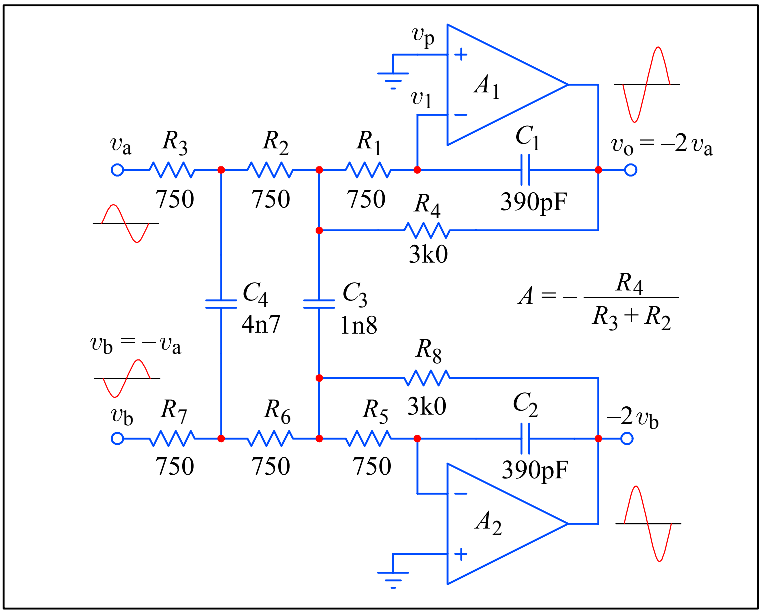

Fig.8: Differential third-order Bessel filter with a 50 kHz cut off, and a gain of two, inverting.

This circuit is an evolution of the usual single-ended inverting multiple feedback design. Inverting, virtual ground filters are preferred to non-inverting types, such as the Sallen–Key configuration, mainly because in non-inverting circuits there is a large common-mode signal driving both the inverting and the non-inverting input to a similar level, but the amplifier is actually processing only the small differential signal. Since in filters the inverting and the non-inverting impedances are very different, a large common-mode sinal may introduce a significant amount of error.

By using a virtual ground configuration this inconvenience is avoided. Again, it is advantageous to use low impedance components in order to limit the circuit noise, since today most good analog-to-digital converters are low-voltage types, their supply often being some 3.3V, and the signal range 1–2V_{pp}. In order to achieve a decent signal to noise ratio, the total circuit noise should not exceed 3μV_{rms}, and the noise density within a 50kHz bandwidth should be less than $15\text{nV}/\sqrt{\text{Hz}}$.

In all filters the effective noise peaks at the filter cut off frequency because the noise gain is directly proportional to the filter Q–factor. Thus the Chebyshev and Butterworth filters with their steeper transition are considerably more noisy in the octaves about the cut off frequency, than the smooth Bessel filter of equal order. Anyway, in audio we prefer Bessel filters for another performance aspect: their flat envelope delay (or equivalently, phase being a linear function of frequency). As all the relevant frequencies pass through the system with equal delay, the waveform transient shape is preserved, and the overshoot, the ringing, and the settling time are minimized. We pay for these goodies by a somewhat less steep transition at the cut off frequency, however that is a modest price indeed!

Hereby I must resist the temptation of writing a full size book on modern filter theory. Large library shelves can be filled by books written on the subject, and I, too, have given my own share, so it would be counterproductive to repeat it here. To spare some paper and a couple of trees, I forward the interested reader to **Ref.6**, where a complete derivation is offered. Now I can almost hear some of the readers saying “oh, not another showing off maths guy, don’t you get it, audio is an art, not science!” Well, I respectfully disagree, audio is a science every bit as it is an art. In order to be able to evaluate and optimize a circuit properly, we have to solve a couple of equations. Not that Mother Nature would be impressed by the endeavour in any way, She will continue to do Her own doings regardless, but if we want to understand how to manipulate the reality to our own taste, we should follow up the often neglected physics. There are no shortcuts here.

Anyway, deriving the filter transfer function in the Laplace space is trivial, only the matching of the components to suit the required system poles is slightly more complicated. The trick to simplify the task of circuit analysis is to consider only the upper arm of the circuit, with the capacitors C_3 and C_4 grounded and of double value (because in the differential case they are being driven by the resistors in both arms, effectively doubling their resistance values).

Here is the final result:

$$\frac{v_\mathrm{o}}{v_\mathrm{a}} = -\frac{R_4}{R_2+R_3}\cdot\frac{K_0}{s^3+s^2K_2+sK_1+K_0} \tag{5}$$

where:

$$K_2 = \frac{1}{C_4R_3}\left(1+\frac{R_3}{R_2}\right)+\frac{1}{C_3R_2}\left(1+\frac{R_2}{R_1}+\frac{R_2}{R_4}\right) \tag{6}$$

$$K_1 = \frac{1}{C_3C_4R_2R_3}\left(1+\frac{R_2}{R_1}+\frac{R_3}{R_1}+\frac{R_2}{R_4}+\frac{R_3}{R_4}\right)+\frac{1}{C_1C_3R_1R_4} \tag{7}$$

$$K_0 = \frac{1}{C_1C_3C_4R_1R_2R_3}\cdot\frac{R_2}{R_4}\left(1+\frac{R_3}{R_2}\right) \tag{8}$$

In **Fig.8** the component values for a third-order Bessel filter with 50 kHz cut off are shown. Those are the nearest standard values, but the deviation from actual Bessel poles is barely noticeable, as shown in **Fig.9**, where the spectral magnitude, phase and envelope delay of the differential output voltage V_d are shown. The positive and negative output voltages V_op and V_on are shown for comparison, being of half magnitude (or $-6\,$dB). Note the envelope delay $\tau_e(V_d)$ being essentially flat up to the cut off frequency of the differential output voltage.

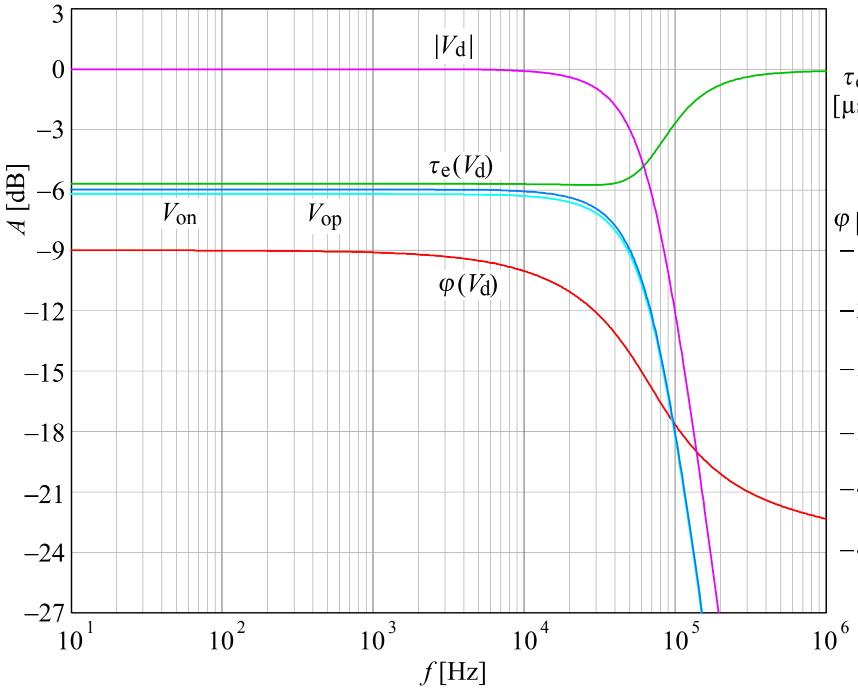

Fig.9: Frequency response magnitude $|V_d|$, phase $\varphi(V_d)$, and group delay $\tau_e(V_d)$ of the differential output voltage $V_d = V_{op} - V_{on}$ of the combination of the differential filter (**Fig.8**) driving the differential line driver (**Fig.7**). The positive and negative output V_{op} and V_{on} are drawn displaced by 0.2 dB to distinguish them better. Note the envelope delay being essentially flat up to the cut off frequency (~50kHz).

In **Fig.10** we show the transient performance of the combined circuits of the differential filter in **Fig.8** driving the common-mode and load unbalance compensating circuit of **Fig.7**. The load imbalance was made to change in $50\,\Omega$ steps throughout the whole $600\,\Omega$ range. Note the compensating currents changing in the opposite direction to the drive currents. The differential voltage remains practically the same in all cases. Apparently the circuit behaves just as it would be expected of an audio transformer.

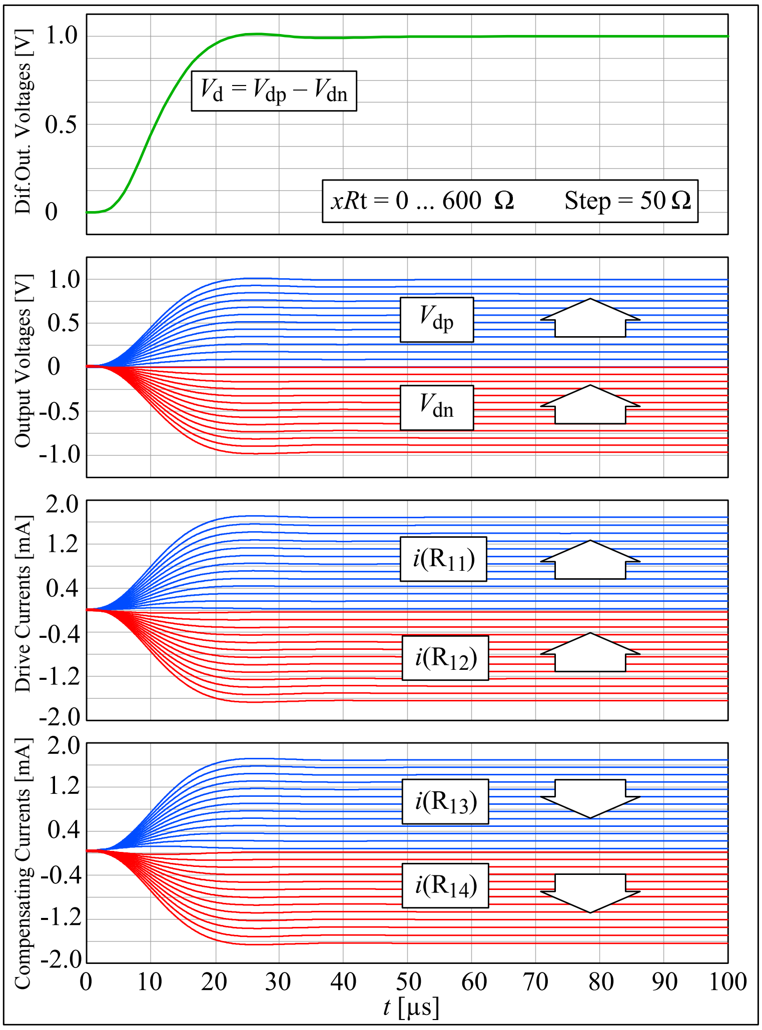

Fig.10: Transient response of the differential filter (**Fig.8**) driving the differential line driver (**Fig.7**) for various load impedance imbalance conditions (50 Ω steps). Note the compensating currents changing in the opposite direction from the drive currents. The differential output voltage remains the same.

Acknowledgment:

I am grateful to Jan Didden for his kind invitation to write an article for Linear Audio. The return to audio was certainly a refreshing experience, after spending the last 20 years designing detectors, amplifiers, and support systems for elementary particle detection, culminating in several systems for the LHC ATLAS Experiment at CERN. LA readers may be surprised to know that the design principles in audio and particle physics are similar, the basic requirements being (in order of importance) reliability, low noise, and low distortion, with a huge difference in bandwidth, of course; though, hopefully, no audiophile will ever need considering the issue of radiation hardness. I hope the LA readers will find this article interesting, and I also hope to be able to return to audio more often in the future.

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This is by no means a complete list, but only an indication where to start.

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