

Balanced Lines in Audio Systems: Fact, Fiction, and Transformers*

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The theoretical benefits of balanced audio interconnection schemes are discussed as well as fundamental but widely unrecognized mechanisms that limit real-world performance. The venerable audio transformer enjoys several inherent advantages over the so-called active circuits offered to replace it in audio equipment. Today, as improved digital recording systems demand increased system dynamic range, these advantages are becoming ever more important.

0 INTRODUCTION

The use of balanced line drivers, shielded balanced twisted-pair cables, and balanced line receivers is a long-standing practice in professional audio systems. Although, in theory, it is a perfect solution to the ground noise contamination problem, very important details of reducing the theory to practice are widely misunderstood, and the resulting *system* performance is often very poor. Traditional or industry standard practices sometimes create more problems than they solve. More often than not, the reduction or elimination of system hum and buzz is the result of a long series of experiments that stop when someone says "I can live with that." If several mode conversion mechanisms are at work simultaneously, we can find ourselves caught in a delicate balancing act of interactions, usually with an unpredictable outcome. However, if we understand the conversion mechanisms, most of them *can* be prevented. Ground noise is very often *the* most serious problem in an audio system. As Hofer wrote [1]: "Many engineers and contractors have learned from experience that there are far more audible problems in the real world than failing to achieve 0.001% residual distortion specs or DC-to-light frequency response."

1 THE INTERCONNECTION PROBLEM IN AUDIO SYSTEMS

In an audio system, delivering an audio signal voltage from the output of device A to the input of device B

may sound simple, but doing so without adding hum, buzz, clicks, and pops coupled from the ac power line is not easy. (We will use the term "audio system" to mean two or more physically separated devices having cabled connections between them to convey analog audio signals. We will also assume that at least two of the devices are powered from common utility ac power.) The basic problem is that ground in one device nearly always has a noisy voltage with respect to ground in another device. It is this voltage that can contaminate the clean audio signal. Each device may work perfectly on the test bench, but become an annoying headache when connected into a system.

Inside each device, small but significant alternating currents flow from the power line through interwinding capacitances of the power transformer and the capacitors in the RFI filter to the chassis (Fig. 1). Relative to an external reference point, such as the ac outlet safety ground, the chassis of each device will assume an ac voltage. If the devices have three-wire power cords, these small alternating currents will flow through the safety ground wire to the ac outlet ground. Since this wire has both resistance and inductance, each chassis will assume a small voltage with respect to the ac outlet ground. Because the coupling capacitance and the wire resistance and inductance effectively form a high-pass filter, the resulting chassis voltage will generally be a rich mixture of high-frequency power-line distortion components, commonly known as buzz. The high-frequency distortion components of the ac power-line voltage originate in loads that draw current during only a portion of each voltage cycle, which includes most electronic equipment, fluorescent or dimmer-controlled lights, and intermittent or sparking loads such as switches, relays, or brush-type motors.

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Even if both devices are plugged into the same ac outlet, because of differing leakage capacitances, they will very likely have differing chassis voltages. If they are plugged into different branch circuits of the ac power system, the voltage differences will generally be much greater. If audio cables connect the chassis of device A to the chassis of device B and other cables connect to still more devices, the flow of power-line related currents and, therefore, the distribution of chassis voltages can get very, very complicated. In all but the smallest audio systems, significant voltages will exist between grounds within the system. If system devices are distant from each other, it may be nearly impossible to eliminate these voltages completely.

We could use an unbalanced scheme and simply connect a wire from device A ground to device B ground and a wire from device A signal output to device B signal input, as shown in Fig. 2. Because the wire connecting grounds has both resistance and inductance, the circulating alternating currents between devices will cause a voltage drop across it. The signal seen at the input of device B will consist of this buzz voltage added to the output signal of device A. Although simple, this unbalanced scheme is generally useful only in the smallest of systems because it is totally unable to reject the noise caused by circulating alternating ground currents.

When dealing with any ac line powered audio system, we must assume that these voltages between grounds will exist and can contaminate the audio signal.

2 DEFINITION AND THEORY OF THE BALANCED LINE

The IEEE Dictionary defines a *balanced line* as "a transmission line consisting of two single or two interconnected groups of conductors capable of being operated in such a way that when the voltages of the two groups of conductors at any transverse plane are equal in magnitude and opposite in polarity with respect to ground, the total currents along the two groups of conductors are equal in magnitude and opposite in direction" [2, p. 50]. In order to satisfy these conditions, the impedances must also be equal with respect to ground. Although indirectly stated in the definition, this property is extremely important to the rejection of common-mode noise voltage.

It is critically important to understand that the effective impedance of each conductor to ground is determined by everything connected to it. Therefore *every* component connected to a balanced line must rigorously maintain its balance in order to realize maximum *system* benefits. This includes the line driver, the transmission line or cable itself, and the line receiver. This is especially true if we wish to interconnect various devices freely and interchange cables, as is usually the case.

The ground noise received from device A, since it exists on, or is common to, both wires, is called the common-mode voltage. The line receiver uses a differential amplifier to reject common-mode voltages. The IEEE Dic-

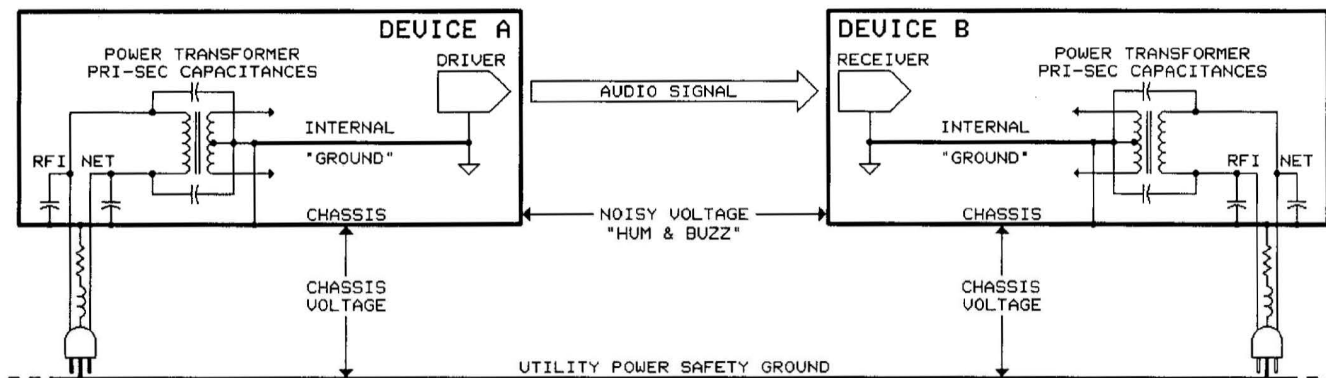


Fig. 1. Origin of hum and buzz voltages.

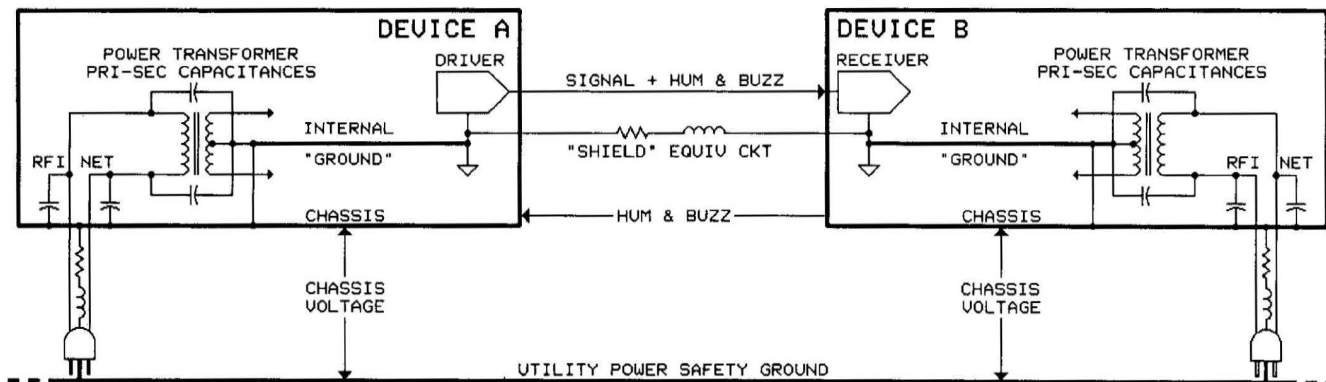


Fig. 2. Unbalanced interconnect scheme.

tionary defines a *differential amplifier* as “an amplifier that produces an output only in response to a potential difference between its input terminals (differential-mode signal) and in which outputs from common-mode interference voltages on its input terminals are suppressed” [2, p. 177]. Note that since a transformer has intrinsic differential response, any amplifier preceded by a transformer also becomes a differential amplifier.

The basic theory of the differential amplifier is straightforward. (For purposes of this discussion we will assume that the ground reference of distant device A has a noise voltage, which we will call “ground noise,” with respect to the local device B ground reference, and that device A has two antiphase outputs.) If we connect a wire to one output of device A, we will see, with respect to local ground, the audio reference polarity signal plus the ground noise. If we connect another wire to the other output of device A, we will see, with respect to local ground, the audio signal of opposite polarity plus the ground noise. All we have to do is subtract the voltage on the second wire from the voltage on the first, which completely cancels the ground noise and effectively adds the two audio signals. If the subtractor inputs have infinite input impedances, the scheme is perfect. In reality, however, device A will have output impedances greater than zero and device B will have input impedances less than infinity, thus forming a voltage divider on each line. To achieve 120 dB of noise rejection, these voltage dividers must match to within 0.0001%, or 1 ppm.

In engineering terms, the subtractor provides common-mode rejection. The ratio of the differential or normal-mode (signal) gain to the common-mode (ground noise) gain of the subtractor is the common-mode rejection ratio (CMRR), also called longitudinal balance, and is usually expressed in decibels. There is an excellent treatment of this subject in Morrison’s book [3, pp. 55–61].

The CMRR will be infinite as long as the common-mode voltages at the inputs of our perfect subtractor are exactly equal. Anything that attenuates these voltages unequally will degrade the CMRR. This unequal attenuation process is called mode conversion since it converts a fraction of the common-mode noise to a normal-mode signal via the impedance unbalances in any of the balanced system components.

Thus far in our discussion we have barely mentioned the normal-mode signal. It is significant that, in order to reject common-mode noise, only the impedances to ground need to be balanced. It is not necessary to have equal-amplitude and opposite-polarity signals in this ideal scheme. When subtracted, the asymmetrical signals $(+1.0) - (0.0)$ or $(0.0) - (-1.0)$ will produce exactly the same output as the symmetrical signals $(+0.5) - (-0.5)$. Signal symmetry is a practical consideration we will consider in Section 6.

3 BACKGROUND OF BALANCED CIRCUITS IN AUDIO

The first widespread users of balanced circuits were the early telephone companies. Earliest systems had no

active amplification, yet needed to deliver maximum audio power from one telephone to another up to 20 mi (32 km) away. It is well known that, with a signal source of a given impedance, maximum power will be delivered to a load with the same, or matched, impedance. It is also well known that reflections and standing waves will occur in a transmission line unless both ends are terminated in its characteristic impedance. Because the signal propagation time through 20 mi (32 km) of line is a significant fraction of a signal cycle at the highest signal frequency, the equipment at each end had to match the line impedance to avoid frequency response errors due to reflections. It appears that the system impedance choice was probably driven by the miles of open wire-pair transmission lines strung along poles originally intended for telegraph use. Typical lines used no. 6 AWG wire at 12-in (0.3-m) spacing, and the characteristic impedance was about 600 Ω , varying by about $\pm 10\%$ for commonly used variations in wire size and spacing [4]. Therefore 600 Ω became the standard impedance for these balanced duplex (bidirectional) wire pairs and telephone equipment in general.

Electrostatic and electromagnetic interference, caused by the ac power lines frequently running parallel to the telephone lines for miles, was largely eliminated through balanced operation of the lines. Balanced and impedance-matched transmission lines were clearly necessary for acceptable operation of the early telephone system. Later, as the telephone network grew, it became necessary to separate the duplexed send and receive signals for unidirectional amplification and long-distance transmission. A hybrid is used for this purpose, and its proper operation depends critically on matched 600- Ω source and load impedances. Telephone equipment and practices eventually found their way into radio broadcasting and, later, into recording and professional audio in general.

In professional audio, however, the goal of the signal transmission system is to deliver maximum voltage, not maximum power. To do this, devices need low differential (signal) output impedances and high differential (signal) input impedances. This practice is the subject of a 1987 IEC standard requiring output impedances to be 50 Ω or less and input impedances to be 10 k Ω or more [5]. Sometimes called voltage matching, it minimizes the effects of cable capacitances and also allows an output to drive multiple inputs simultaneously with minimal level losses. With rare exceptions, such as telephone equipment interfaces, the use of matched 600- Ω sources and loads in professional audio is generally unnecessary and compromises performance.

4 LINE RECEIVERS

Since performance of the differential line receiver is the most important determinant to system CMRR performance and can, in fact, reduce the effects of other degradation mechanisms, we will discuss it first. There are two basic types of differential amplifiers—active circuits and transformers. Active circuits are made of operational amplifiers and precision resistor networks to

perform algebraic subtraction of the two input signals. The transformer is an inherently differential device which provides electrical isolation of input and output signals.

The active differential amplifier, sometimes called an actively balanced input, is realizable in several circuit topologies. These circuits are well known and have been analyzed and compared in some detail by others [3, pp. 70–75], [6, pp. 201–207], [7], [8]. In our discussion here we will assume that operational amplifiers and resistors are ideal and that resistor values are exact and not a source of error. Fig. 3 shows the schematics of four popular versions in their most basic forms, stripped of ac coupling, RFI filtering, and so on. Because the common-mode input impedances, from either input to ground [6, p. 441], are all 20 kΩ, these four circuits have identical CMRR performance. Even when perfectly matched, these impedances are the downfall of this approach. To quote Morrison: “Many devices may be differential in character but not all are applicable in solving the basic instrumentation problem” [3, p. 58]. Fig. 4 shows the extreme sensitivity of 60-Hz CMRR versus source impedance unbalances for these circuits. These circuits are invariably tested and specified with perfectly balanced sources or short-circuited inputs. However, in the real world, unbalances in the 0.2–20-Ω range are common, and advertised CMRR performance in a system is rare. There are other problems, too.

1) The single and current-mode dual operational amplifier circuits of Fig. 3 must trade off common-mode input impedance for noise. For example, because of the high-value resistors, the single operational amplifier circuit will have a noise output of about -105 dBu, where 0 dBu = 0.775 V rms. If it operates on ±15-V rails, it will have a maximum output of about +20 dBu, giving it a total dynamic range of 125 dB. This may be marginal

in some recording systems. If the resistor values are doubled, which will decrease the CMRR sensitivity to source impedance unbalance, noise will increase by 3 dB.

2) Many circuits use electrolytic coupling capacitors at their inputs. These capacitors have inherently loose tolerance and drift with age, degrading the low-frequency CMRR by unbalancing the common-mode input impedances.

3) Suppression of RF common-mode voltages, to prevent subsequent demodulation by the operational amplifiers, is another tradeoff for these circuits. Often 1000-pF capacitors are added from each input to ground to attenuate radio frequency (RF). Unless very carefully matched, these capacitors unbalance the common-mode input impedances and degrade the high-frequency CMRR. Also, because they lower common-mode input impedances, they increase the high-frequency CMRR

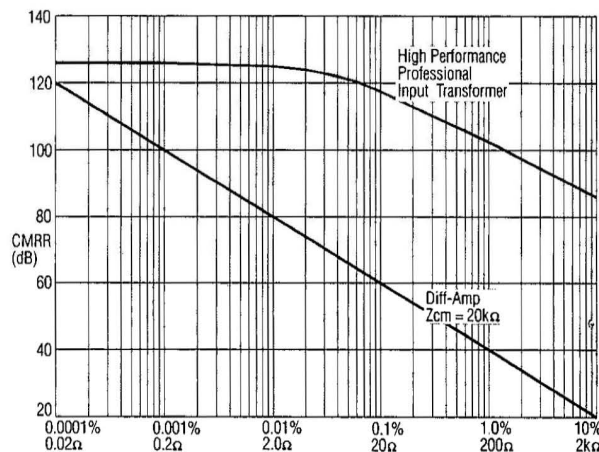


Fig. 4. 60-Hz CMRR versus source unbalance for differential amplifier and transformer.

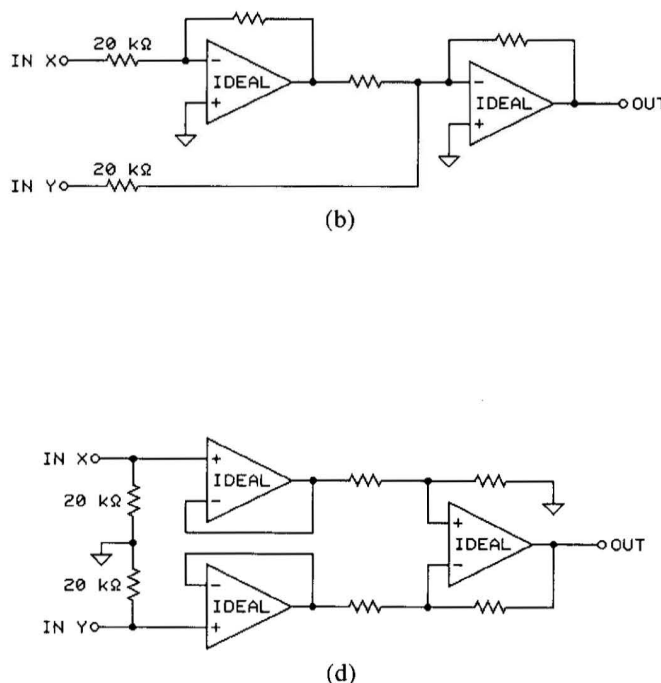
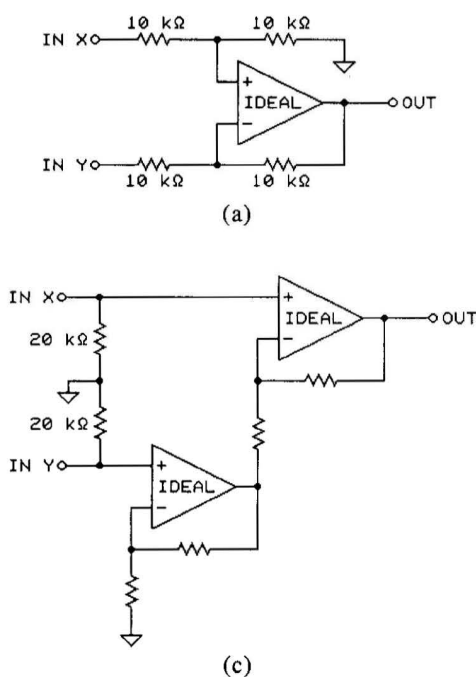


Fig. 3. Popular active differential amplifier circuits. (a) “Simple” single operational amplifier. (b) Current-mode dual operational amplifier. (c) Voltage-mode dual operational amplifier. (d) Instrumentation or triple operational amplifier.

sensitivity to source impedance unbalance. This is a very tricky tradeoff.

4) The common-mode voltage range is limited to ± 10 to ± 15 V for most circuits. At high signal levels the common-mode range can approach zero because the limit applies to the sum of the peak signal and the peak common-mode voltages [9], [10, p. 42]. This can cause problems in electrically hostile environments such as remote recording trucks or sound reinforcement systems operating near high-power lighting equipment or cables.

5) The single operational amplifier design of Fig. 3 also has a property that seems confounding [11]. Although its common-mode input impedances are identical (when the voltages at inputs X and Y are equal), its differential signal input impedances are not. Obviously, if driven from a zero-impedance balanced ground-referenced source, the voltages at X and Y are forced to be identical. Real-world floating sources, which have high common-mode output impedances, will experience signal magnitude unbalances, typically around 3 dB, when used with this receiver. In fact, if driven by an ideal floating source, all signal voltage will appear at input X and none at input Y. The effect of signal unbalance is included in the discussion of cables in Section 6.

An audio transformer converts an input signal to a magnetic field and back to a signal that is electrically isolated from the input. It is an inherently differential device, requiring no ultraprecise nulling or trimming, and its differential properties are stable for life. Fig. 5 shows a circuit simulation model for a high-performance professional line input transformer. Its common-mode input impedances are determined by the 50-pF capacitances of the primary to the Faraday shield, which is grounded, and small parasitic capacitances to the secondary, which is also effectively grounded. These high common-mode input impedances, about 50 M Ω at 60 Hz and 1 M Ω at 3 kHz, are responsible for the relative insensitivity to the large source impedance unbalances shown in Fig. 4. There are other advantages, too.

1) A transformer can transform or match the impedance of the balanced line to the optimum source resistance for the subsequent amplifier to maximize the signal-to-

noise ratio. The noise figure is a measure of the signal-to-noise degradation attributed to the amplifier, and it reaches a minimum value when operating from the optimum source resistance [12]. Although this is especially relevant to microphone preamplifiers, it applies to line input stages as well. A transformer coupled line input stage operating from ± 15 -V rails can easily attain 140 dB of total dynamic range.

2) RF common-mode attenuation is inherent in transformers with Faraday shielding. CMRR is not a useful measure of this attenuation since it compares normal-mode to common-mode responses. Although the normal-mode -3 -dB bandwidth is usually 100–200 kHz for this type of transformer, its common-mode attenuation is typically over 30 dB from 200 kHz to 10 MHz. This attenuation can be supplemented, if necessary, with a simple external network that does not compromise CMRR performance.

3) Limits on the input common-mode voltage in a transformer are imposed by the materials used to insulate the windings. Breakdown typically exceeds ± 350 V peak.

5 LINE DRIVERS

There are three basic types of line drivers—ground referenced, active floating, and transformer floating. Fig. 6 shows simplified schematics of each type connected to an ideal line receiver having a common-mode input impedance of exactly 20 k Ω per input. (Differential or signal voltage generators are shown for clarity, but for common-mode noise analysis the generators are considered short circuits. The receiver ground is considered the zero signal reference and the driver ground is at common-mode voltage with respect to the receiver ground.) Fig. 7 compares the simulated CMRR performances of the three sources with this receiver setup.

1) The *ground-referenced source* has two antiphase voltage sources, each referenced to the driver ground. The resistive common-mode output impedances are R_{S1} and R_{S2} . The differential output impedance R_{OD} is simply $R_{S1} + R_{S2}$. The common-mode voltage V_{CM} is fed into

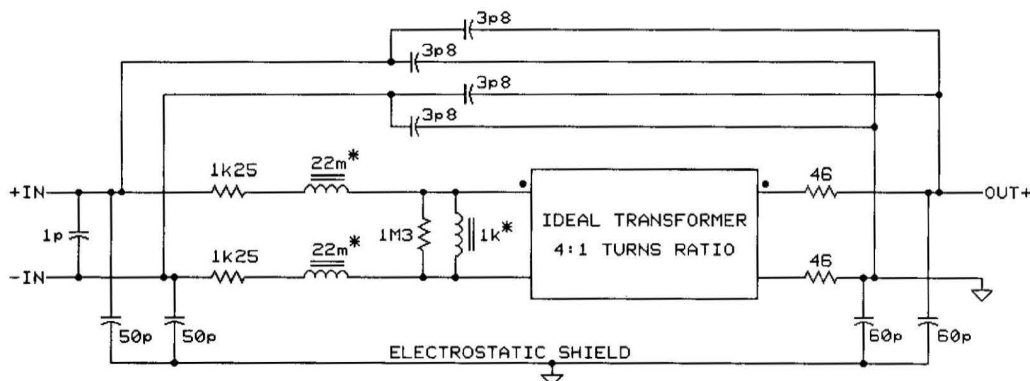


Fig. 5. Circuit simulation model for high-performance input transformer. All resistance in ohms, capacitance in farads, and inductance in henrys. Unit prefix used in place of decimal point: 3p8 = 3.8 pF, 1k25 = 1.25 k Ω , and so on. *Inductances vary inversely with frequency, increasing at ~ 3 dB/oct down to ~ 1 Hz. Value indicated as 1 kH applies at 20 Hz; 22 mH applies at 100 kHz. Values shown are typical only.

both line branches through R_{S1} and R_{S2} . V_{CM} appears at the line receiver attenuated by two voltage dividers formed by R_{S1} and $20\text{ k}\Omega$ in one branch and by R_{S2} and $20\text{ k}\Omega$ in the other. As discussed previously ratio matching errors in these two voltage dividers will degrade the CMRR. (It could be argued that R_{S1} need not equal R_{S2} and that the common-mode input impedances need not match because this condition is not necessary for ratio matching. However, equality is necessary if we wish to allow the interchange of system devices.) Since typical values for R_{S1} and R_{S2} are $20\text{--}100\ \Omega$, each with independent tolerances of ± 1 to $\pm 10\%$, the worst-case source impedance unbalance could range from 0.4 to $20\ \Omega$. With these unbalances the system CMRR will degrade to 94 dB for $0.4\ \Omega$ or to 60 dB for $20\ \Omega$. Since the unbalances are resistive, the CMRR is constant over the audio-frequency range.

2) The *active floating source* is built around a basic circuit consisting of two operational amplifiers cross-coupled with both negative and positive feedback to emulate a floating voltage source. The resistive common-mode output impedances are R_{CM1} and R_{CM2} . The differential output impedance is R_{OD} . The common-mode voltage V_{CM} is fed into both line branches through R_{CM1} and R_{CM2} . V_{CM} appears at the line receiver attenuated by two voltage dividers formed by R_{CM1} and $20\text{ k}\Omega$ in one branch and R_{CM2} and $20\text{ k}\Omega$ in the other, with R_{OD} across the line. R_{OD} is typically $50\text{--}100\ \Omega$. Since the common-mode output impedances of this circuit are increased by precise balancing of the resistor ratios

which also interact with the output signal balance, adjustment is difficult and the values for R_{CM1} and R_{CM2} are not specified directly. One manufacturer of this circuit specifies output common-mode rejection (OCMR) by the B.B.C. test method [13, pp. 7-139–7-144]. The results of this test can be used to determine the effective values of R_{CM1} and R_{CM2} using computer-aided circuit analysis. Values of 5.3 and $58.5\text{ k}\Omega$ were found for a simulated part having OCMR and SBR (signal balance ratio) performances slightly better than the typical specification. For this simulated part the system CMRR was degraded to 57 dB . Since the unbalances are resistive, the CMRR is constant over the audio-frequency range.

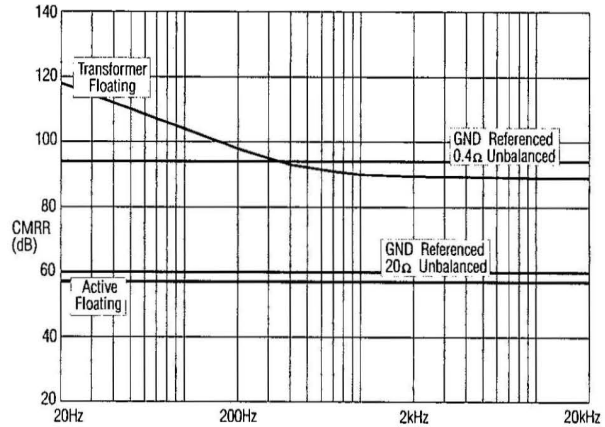


Fig. 7. Simulated response of drivers with $20\text{-k}\Omega$ ideal differential amplifier.

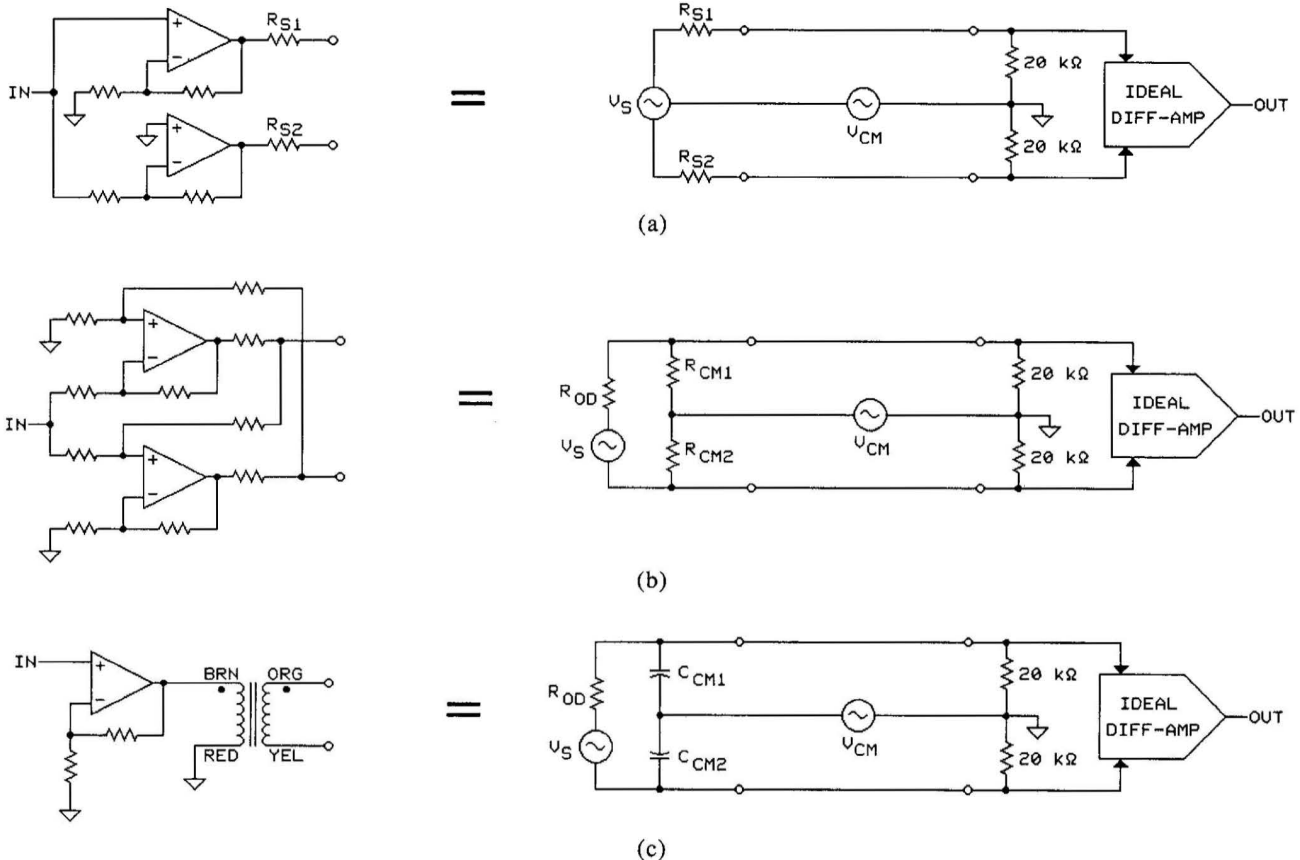


Fig. 6. Basic line drivers and equivalent circuits driving ideal $20\text{-k}\Omega$ differential amplifier. (a) Ground-referenced source. (b) Active floating source. (c) Transformer floating source.

3) The *transformer floating source* consists of a transformer whose primary is driven by an amplifier whose output impedance is effectively zero by virtue of conventional negative feedback. The common-mode output impedances C_{CM1} and C_{CM2} consist of the interwinding capacitance for multifilar wound types, or the secondary-to-shield capacitance for Faraday shielded types. The differential output impedance R_{OD} is the sum of secondary and reflected primary winding resistances. For typical bifilar transformers C_{CM1} and C_{CM2} range from 7 to 20 nF each, matching to within 2%. A typical R_{OD} range is 35–100 Ω . The system CMRR will be 110–120 dB at 20 Hz, decreasing by 6 dB per octave, since the unbalances are capacitive, to 85–95 dB at 500 Hz, above which it becomes frequency independent. If, instead of the active receiver, a high-performance input transformer is used, its full CMRR capability of about 125 dB at 60 Hz and 85 dB at 3 kHz is realized with any of the sources and conditions described.

The grounded load behavior of these three sources is often an important consideration. Of course, for any line driver either output should be capable of withstanding an accidental short circuit to ground or to the other line indefinitely without damage or component failure. This is best accomplished with current limiting and thermal shutdown features.

The ground-referenced source will deliver abnormally high, and usually very distorted, current into a remote ground such as an unbalanced input. This current must return to the driver, and as it does, it may circulate through other system ground paths, creating common-mode voltages at other device inputs. Symptoms of this problem are usually described as crosstalk.

The active floating source compromises CMRR, output magnitude balance, and high-frequency stability in quest of a transformerlike ability to drive a grounded or single-ended input. However, to remain stable, the grounded output must be carefully grounded at the driver [13, p. 7-144], [14]. Since this makes the system completely unbalanced, it is a serious disadvantage.

The transformer floating source breaks the ground connection between the driver and the unbalanced input. Because the transformer secondary floats, it is able to reference its output to the unbalanced input ground. Power-line hum is reduced by more than 70 dB in the typical situation of Fig. 8. Because the ground noise is capacitively coupled through C_{CM1} , reduction decreases

with frequency, as shown in Fig. 9.

With the transformer floating source, if it is known that an output line will be grounded, an appropriate step can be taken to optimize performance further. With a differentially driven transformer, drive should be removed from the corresponding end of the primary to reduce the signal current in the remotely grounded output line. In the case of a single-ended driven transformer, simply choose the secondary line corresponding to the grounded end of the primary for grounding. Fig. 8 illustrates the method in a typical system. With ground-referenced or active floating outputs that require grounding at the driver, the system degenerates to the unbalanced interface of Fig. 2 and has no ground noise rejection at all.

6 CABLES

The primary effects on system behavior caused by the interconnecting shielded twisted-pair cable are due to the capacitance of its inner conductors to the shield. The two inner conductors of widely used 22-gauge foil-shielded twisted-pair cable, when driven common-mode, exhibit a capacitance to the shield of about 67 pF/ft. But the capacitance unbalance can be considerable. Measurements on samples of two popular brands of this cable showed capacitance unbalances of 3.83 and 3.89%, with the black wire having the highest capacitance in both cases. On one sample, the insulation thick-

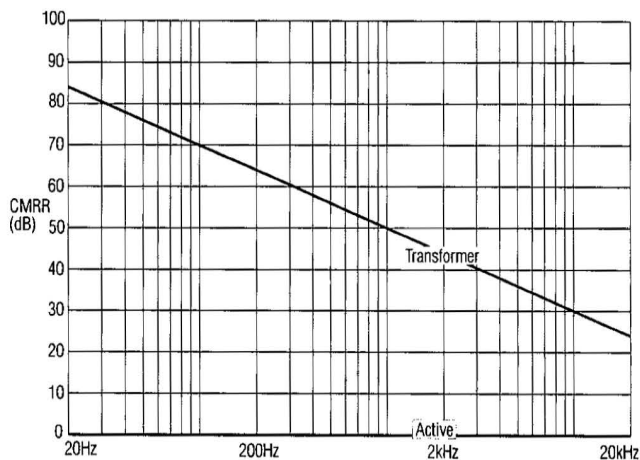


Fig. 9. Transformer versus grounded active driver into unbalanced input.

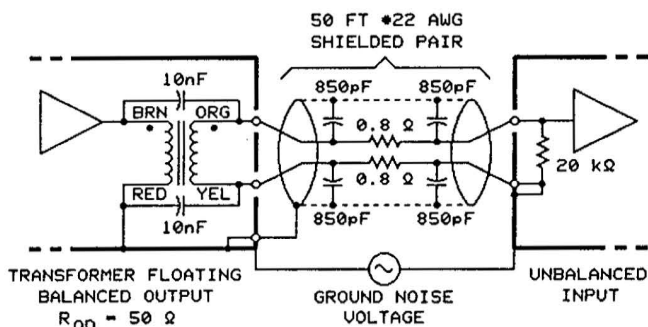
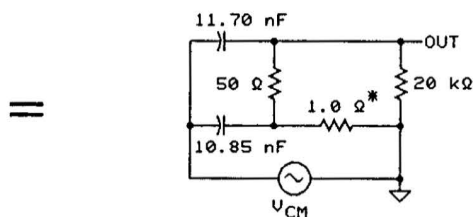


Fig. 8. Transformer floating output to unbalanced input.

EQUIVALENT CIRCUIT AT AUDIO FREQUENCIES



*INCLUDES 2 CONNECTORS AT 0.1 Ω EACH

ness was calculated from outside diameter measurements, and it was assumed that the stranded conductors in both wires were identical. The black insulation was 2.1% thinner, and since the capacitance varies as the inverse square of the thickness, this would seem to explain the unbalance. Perhaps this topic needs some attention from cable manufacturers.

This is important because, if the cable shield is grounded at the receive end, these capacitances and the output impedances of the driver form two low-pass filters. Unless these two filters match *exactly*, requiring an exact match of both driver output impedances and cable capacitances, mode conversion will take place. Such conversion is aggravated by long cables and unbalanced driver impedances. But this conversion can be avoided. The cable shields must be grounded at the driver end only. Examination of Fig. 10 shows how the common-mode noise is low-pass filtered.

Remember that our reference point is the *receiver* ground. If we simply ground the cable shield at the driver end instead, as shown in Fig. 11, no common-mode voltage appears across the cable capacitances and no filters are formed. Since the shield now is at the common-mode voltage and so are both driver outputs, there is no common-mode voltage across the cable capacitances and they effectively disappear. As far as the common-mode voltage on the signal conductors is concerned, the cable capacitances are now in parallel with the source impedances, virtually eliminating the unbalancing effects of the capacitances. Because of its high common-mode output impedances, the active floating

driver is very vulnerable to this conversion mechanism.

The grounding of shields at both the driver and the receiver creates an interesting tradeoff. The cable effects will, predictably, fall between the schemes of Figs. 10 and 11. The advantage is that it can reduce the common-mode voltage itself, even though it may degrade the rejection of it by the receiver. If there is no other conducting path between the chassis of two devices, using the cable shield to connect them will likely reduce the common-mode voltage. It would be far better, of course, to use some other means, such as a dedicated grounding system or even the utility safety ground (ac cord third pin) to limit the common-mode voltage. Devices with no safety ground wire (two-pin ac plugs) are the most offensive in this regard, with their chassis voltage often well over 50 V ac above utility safety ground. The current available is very small, posing no safety hazard, but it creates a very large common-mode voltage unless somehow contained.

A complex ac circuit simulator model of 100 ft (30 m) of the popular shielded twisted-pair cable was generated to study the effects of grounding the shield at both ends. The cable model accounts not only for capacitances, but for resistances, inductances, and conductor-to-conductor transformer coupling effects as well. The effects on the CMRR caused by grounding both ends of the shield are shown in Fig. 12. The simulation was run using a ground-referenced driver having a differential output impedance of 100 Ω and a perfect receiver having infinite input impedances and CMRR. As we might expect, for the perfectly balanced source, the CMRR is well over 200 dB. As we unbalance the system as little as 1 Ω , however, we see the CMRR at 3 kHz fall to about 90 dB. For many systems this may not be a serious practical problem, but remember that this is a limit to the achievable CMRR. If the actual input stage has only 70 dB of CMRR, this cable limitation is probably of no consequence, but if it is capable of 100 dB, the limitation will be real. This degradation is directly dependent on the cable length. A 200-ft (60-m) cable will worsen the effects by 6 dB, 500 ft (150 m) by 14 dB, and so on.

As mentioned in Section 2, it is not necessary to have

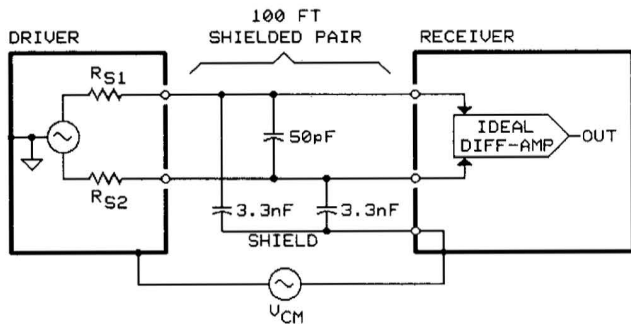


Fig. 10. Incorrect shield connection. A fraction of V_{CM} appears as differential signal unless $R_{S1} = R_{S2}$ and the 3.3-nF cable capacitances are matched exactly.

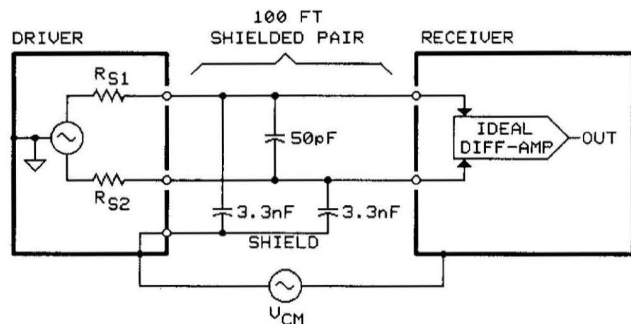


Fig. 11. Correct shield connection. Conversion problem of Fig. 10 eliminated by connecting cable shield at driven end only.

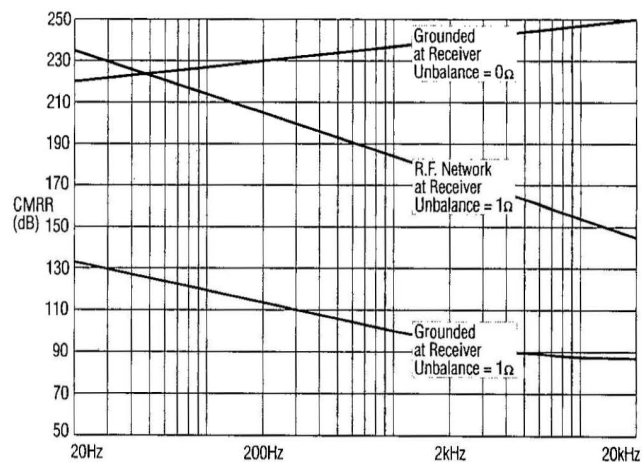


Fig. 12. Effects of shield grounding at driver and receiver. Ground-referenced driver with R_{OD} 100 Ω and ideal receiver.

symmetrical signals on the balanced line in order to reject common-mode noise. Signal symmetry is a practical consideration to cancel capacitively coupled signal currents, which would otherwise flow in the cable shield. In a real system there will be some signal currents flowing in the shield because of either signal asymmetry or capacitance unbalance in the cable. If the cable shield is grounded only at the driver, these currents will flow back to the driver and have negligible system effects. If the shield is grounded to the receiver, these currents, because they rise with frequency and will return to the driver after circulating through portions of the grounding system, can cause very strange symptoms or even ultrasonic oscillations.

Especially with long cables, the scheme of Fig. 10 can have an RF shielding advantage. Because Fig. 11 leaves the shield floating at the receive end, increased RF common-mode voltage may appear at the receiver since the cable itself can become an effective antenna. In these cases a modification to the scheme of Fig. 11 can be made to RF terminate the shield at the receive end [3, p. 86]. We computer modeled this using a series-connected 50- Ω resistor and a 10-nF capacitor connected between shield and chassis at the receiver. Under the same conditions as our previous example, using 100 ft (30 m) of cable and a 1- Ω unbalance, the CMRR limits became 220 dB at 60 Hz and 170 dB at 3 kHz, as shown in Fig. 12. Although AM radio is the usual offender in these cases, pickup and subsequent demodulation of broadcast television signals can result in a buzz remarkably similar to the power-line variety. An oscilloscope with sweep locked to the 60-Hz power line will reveal the "crawl" of the 59.94-Hz vertical sync pulses.

Especially if they are long or bundled with unshielded wiring, cables should have 100% shield coverage. Tight twisting (high twists per unit length) or star-quad construction will greatly reduce magnetic pickup. This is especially important if the cables are near a power transformer, motor, computer CRT, television receiver, or wires carrying high alternating currents.

7 TRANSFORMER PERFORMANCE ISSUES

There is a widespread notion that all audio transformers have inherent limitations such as high distortion, mediocre transient capability, and large phase errors. Unfortunately many such transformers do exist, and not all of them are cheap. Manufacturers of such transformers are simply ignorant of sonic clarity issues, have a poor understanding of the engineering tradeoffs involved, or are willing to take manufacturing shortcuts that compromise performance to meet a price.

Accurate performance in the time domain, or waveform fidelity, is critically important. Accurate time-domain performance, sometimes called transient response, requires low phase distortion. Since a pure time delay exhibits a linear phase versus frequency characteristic, true phase distortions are expressed as deviations from linear phase (DLP). Phase shift is not necessarily phase distortion [15]. In order to achieve a DLP of 5°

or less from 20 Hz to 20 kHz, the -3-dB frequency response must extend from 0.87 Hz to 35 kHz for first-order filter responses. If the low-pass filter is a second-order Bessel, the cutoff frequency can be as low as 25 kHz [16].

Phase distortion not only alters sonic quality, it can have potentially serious system headroom implications. Even though the frequency response may be flat, peak signal amplitudes can increase up to 15 dB after passing through a network with high phase distortion. Even ultrasonic phase distortions caused by undamped resonances can excite complex audible cross-modulation products in subsequent nonlinear (any real-world) amplifier stages [17]. Low-frequency phase distortions are often described as muddy or loose sounding bass and high-frequency phase distortions as harshness or mid-range smear. The complex cross-modulation products are sometimes described as dirty sounding or causing listener fatigue. All Jensen transformers are expressly designed for very low-phase distortion, most measuring well under $\pm 2^\circ$ DLP from 20 Hz to 20 kHz. Typical frequency response is -3 dB at 0.5 Hz and 180 kHz and -0.05 dB from 20 Hz to 20 kHz. Low-pass response, or high-frequency rolloff, has a second-order Bessel characteristic. High-pass response, or low-frequency rolloff, is actually *sub* first order (less than 6 dB per octave) because of permeability versus frequency effects in the core material. This further improves the DLP performance.

Distortion in audio transformers is unusually benign in sonic character. It is caused by a smooth symmetrical curvature of the magnetic transfer characteristic of the core material and is proportional to flux density. For a constant-voltage input the flux density is inversely proportional to frequency. Since lower frequency and higher level both increase flux density, and therefore distortion, this determines the "maximum 20-Hz level" for a given design. The distortion products are nearly purely third harmonic. Even at high levels the distortion roughly quarters for every doubling of frequency, dropping to less than 0.001% above a few hundred hertz. The distortion mechanism is frequency selective in a way that an amplifier, for example, is not. If an amplifier with a flat frequency response measured 0.1% total harmonic distortion at a given signal level, for example, we would normally expect its SMPTE IMD (intermodulation distortion measured using 60-Hz and 7-kHz sine waves in a 4:1 ratio) to be around 0.3% at an equivalent level. If we test a quality transformer the same way, we find that the IMD is actually several times lower than the total harmonic distortion. For example, at +26-dBu input, the transformer cited earlier has a total harmonic distortion of about 0.03% at 60 Hz. But at the equivalent level, the SMPTE IMD is only 0.01%. To summarize, transformer distortion is almost purely third harmonic, occurs only at high levels of very low frequencies, and is not accompanied by the much more audible intermodulation distortions that occur in most other types of audio devices.

The vast majority of available audio transformers,

even when used as directed, do not achieve professional performance levels. As Perkins wrote: "With transformers, you get what you pay for. Cheap transformers create a host of interface problems, most of which are clearly audible" [10, p. 41]. If well designed and properly applied, however, audio transformers qualify as true high-fidelity devices. They are passive, stable, reliable, and require neither trimming, tweaking, nor excuses.

8 CONCLUSION AND RECOMMENDATIONS

The steadily increasing dynamic range of modern digital recording equipment requires careful attention to analog interfacing in order to realize full benefits at the system level. Rejection of common-mode noise, often the most challenging of the system problems, is easily accomplished using refinements of the balanced line and transformer techniques developed by early telephone companies.

The following are suggestions, not necessarily in order of importance, to help maximize the useful system dynamic range.

- Minimize system common-mode voltages using sensible grounding schemes that do not depend on cable shields as current paths.
- Bond the shield pin of all outputs and microphone inputs as directly as possible to the *chassis* and the safety ground.
- Reduce the pickup of RF fields by routing cables so that they "cling" to large ground-plane areas, such as equipment racks or steel-reinforced concrete floors.
- Do not use longer cables than necessary. Coiling of excess length is a bad idea.
- To minimize hum pickup, use star-quad style cables routed as far as possible from ac magnetic fields.
- Use line receivers with very high common-mode impedances and CMRR. A high-quality input trans-

former gives the best possible CMRR performance from a wide variety of sources, including floating, actively balanced, and even totally unbalanced ones.

- All these foregoing suggestions apply to microphone inputs as well [18].
- Use a three-position "direct-open-RF" option switch at the shield pin of line inputs. Open is the preferred normal condition.
- Use appropriate high output line drivers. Depending on the cable lengths, a driver should be capable of ± 50 to ± 250 mA peak currents to avoid high-frequency distortions due to current-limited slewing [10, p. 46]. Fig. 13 shows three drivers capable of dynamic ranges greater than 140 dB. Damped inductor load isolators, Z_i in Fig. 13, are preferred over resistors to minimize unbalancing effects of both driver output impedance and cable capacitance.
- Keep the operating signal levels as high as possible.

9 ACKNOWLEDGMENT

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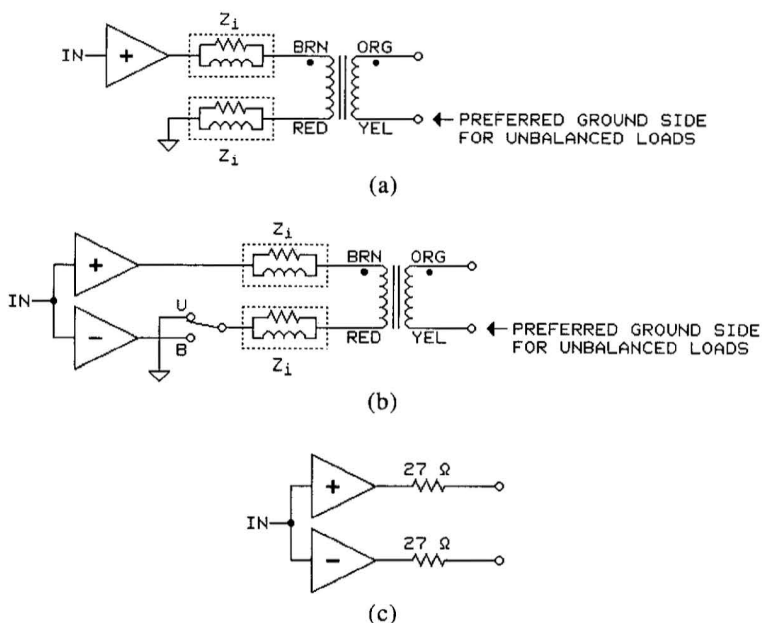


Fig. 13. High-performance line drivers. (a) All-purpose driver. (b) High-output driver. Switch optimizes performance for balanced or unbalanced loads. (c) High output for differential loads only.

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THE AUTHOR



William E. (Bill) Whitlock was born in Harrisburg, IL, on April 14, 1944. In 1950, his family moved to St. Petersburg, FL, where he was reared and educated. Electronics became a passionate avocation by age 8 and by 14, he was operating his own radio/TV repair shop and had won several science fairs. He attended St. Petersburg Junior College and Pinellas County Technical Institute from which he graduated with honors in 1965. Before moving to California in 1971, he held engineering positions with Schlumberger/EMR Telemetry, General Electric Neutron Devices, and RCA Missile Test Project.

Mr. Whitlock's involvement in professional audio began in 1972, when he became chief engineer for console-maker Quad-Eight, replacing Deane Jensen who was leaving to start Jensen Transformers. At Quad-Eight, he conceived and developed COMPUMIX™, a pioneering console automation system. He presented a paper and was a panelist on the console automation forum at the 45th Convention of the Audio Engineering Society in 1973. In 1974, he became chief engineer for Laser Images, producers of LASERIUM™, a planetarium laser light and music show. He designed automated image

synthesis and control systems, several theater sound systems, and, in 1975, a VCR-based digital audio recording system for which he received a patent in 1977. In 1981, he became manager of electronic development engineering for EMI/Capitol Records where he designed the XDR™ high-speed cassette duplicator system electronics, including a high-performance version of Dolby's HX Pro™. From 1986 to 1988, he was project leader for Capitol's digital loop-bin project, which produced high-speed duplicated cassettes directly from the digital master.

Deane Jensen and Mr. Whitlock worked together as a consulting team until Mr. Jensen's death in October, 1989, when, as Jensen had requested, Whitlock became president of Jensen Transformers. In addition to continued refinement of its high-grade audio transformers, he has added high-performance electronics to the Jensen product line. He also does consulting engineering for outside clients as time permits.

Mr. Whitlock's research work currently involves high CMRR audio input circuits, for which he now has a patent pending. He is an active member of the AES, IEEB, and SMPTE.