Subtleties count in wide-dynamic-range analog interfac

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Transporting high-dynamic-range analog signals from one piece of equipment to another is not a trivial task. subtle design variations can make huge differences in the equipment's ability to reject interface from the ac pow and other sources when the equipment connects to real-world system.

Ground noise is often the most serious problem in a system. Reducing or eliminating this noise is usually the a series of experiments that stops when someone says, "I can live with that." If several coupling or conversion mechanisms are working simultaneously in the circuit, these experiments become a delicate balancing act of interactions. However, if you understand the conversion mechanisms, you can prevent most of these interaction

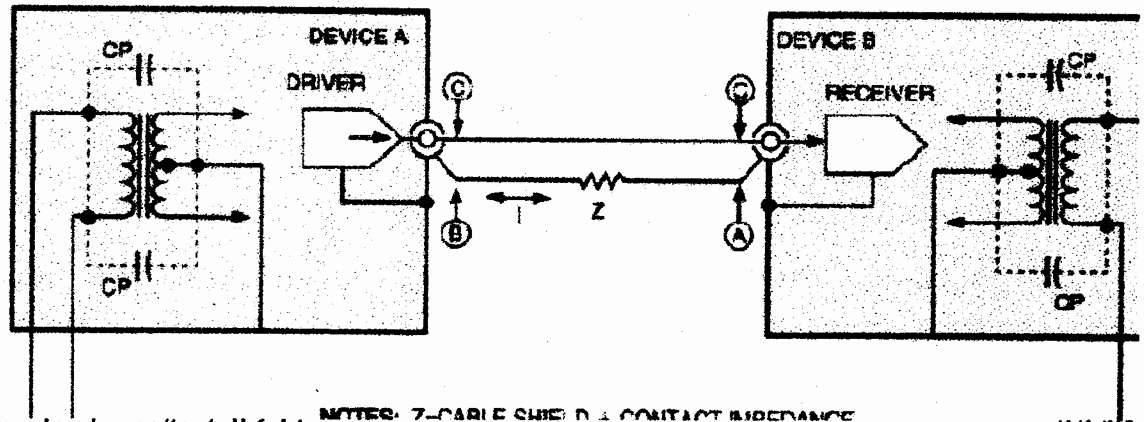
The term "noise" here means any undesired in-band signal interference, rather than the rigorous engineering definition. A "system" is two or more physically separate, ac-powered devices with cabled analog-signal connec between them. Although this discussion concentrates on audio systems because they typically require a large dy range, the principles involved apply to any system. Contrary to popular belief, digital interfaces are also suscept these problems; they simply exhibit different symptoms from analog interfaces.

Noise is pervasive

The fundamental interface problem stems from the fact that once noise contaminates a signal, it's nearly imp to remove the noise. Dynamic range quantifies the ratio of the maximum undistorted signal to the noise floor, w SNR quantifies the ratio of the reference signal to the noise floor. Dynamic range equals SNR plus "head room' ratio of the maximum undistorted signal to the reference signal. These values are generally expressed in decibel

System-dynamic-range requirements depend on the application and on user expectations. The human ear has ab dB of dynamic range, whereas a high-performance audio-reproduction system in a typical home listening enviro may require as much as 120 dB (**Reference 1**). Video systems generally accept 50 dB of dynamic range as the l beyond which expert viewers perceive no further improvement.

Both basic types of interfaces – unbalanced and balanced – use a pair of wires to carry the signal; the impedanc these wires – with respect to a reference point, usually ground – define them. In an ideal unbalanced interface, c has zero impedance, and the other signal-carrying wire has nonzero impedance to ground. In the ideal balanced interface, both wires have equal and nonzero impedances to ground.



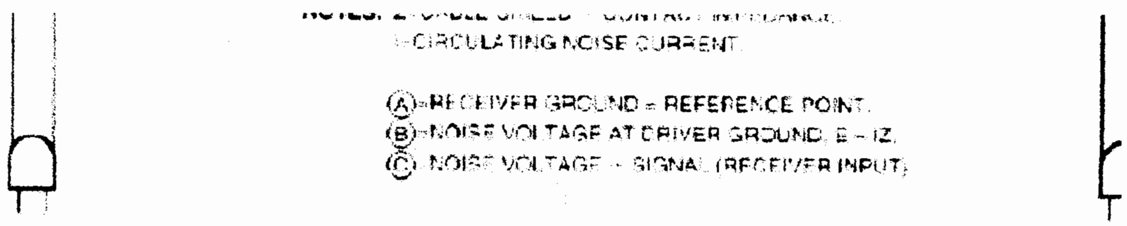


Fig. 1

The offending common impedance of the unbalanced interface is the result of the grounded conductor o interconnecting cable and the contact resistance of any connectors.

When you are dealing with any ac-line-powered system, you must accept the existence of significant ground-vo differences between system components. Although you can sometimes reduce these voltages by carefully desigı executing system-grounding schemes, they are virtually impossible to eliminate. In most systems, these voltages dominant noise source, entering unbalanced signal paths through common-impedance coupling and balanced pa through common-mode conversion. Common symptoms are hums, buzzes, pops, clicks, and other noises in auc systems; hum bars or bands of "sparkles" in video systems; and unexplained data errors or crashes in data systeı

All internal and external power transformers have unavoidable parasitic capacitances from their power-line-conı primary windings to their equipment-ground-connected secondary windings. These parasitic capacitances never on schematic diagrams, and you cannot eliminate them in a practical way. Power-line RFI/EMI filters generally even larger capacitances from their lines to their chassis. The periodic charge and discharge of these capacitanc small but significant ac-power-line currents to flow from the power line to each chassis. System devices are eith "grounded" or "floating."

Grounded devices use three-wire power cords. Parasitic currents flow through the safety ground wire to the ac ground. Because this wire has both resistance and inductance, each chassis assumes a small voltage with respec outlet ground. The series-coupling capacitance and shunt-wire resistance/inductance effectively form a highpass so the resulting chassis voltage generally is a rich mixture of high-frequency power-line noise and distortion components, which you hear as a buzz rather than the more fundamental rich hum in an audio system. Nonlinea on the power line can generate these high frequencies. Such loads can include electronic equipment with a capa input or switching power supplies; fluorescent or dimmer- controlled lights; and intermittent or sparking loads, switches, relays, or brush motors.

Even if you plug both devices into one outlet, because of differing parasitic capacitances, the chassis voltages w be different. Because "grounded" devices connect the chassis to safety ground at low frequencies, each device effectively acts as a voltage source with an impedance less than a few tens of ohms. If you plug two devices int different branch circuits of the ac power system, the voltage differences generally increase, often reaching sever If you connect the devices by a cable shield, for example, 100-mA currents may flow. Even higher voltages and resulting current flows can result if you connect one of the devices to a nonpower ground, such as the earth grc lightning safety with a cable-TV or a satellite-broadcast receiver.

Note: *Never lift or disconnect safety grounds; it's not only illegal, but also dangerous.* You use a ground adap provide a safety ground for three-conductor power cords with two-prong outlets, not to defeat the safety grour three-prong outlet provides. Defeating a safety ground could allow lethal voltages to appear on all equipment ir interconnected sys- tem should a power-line fault develop in the lifted device.

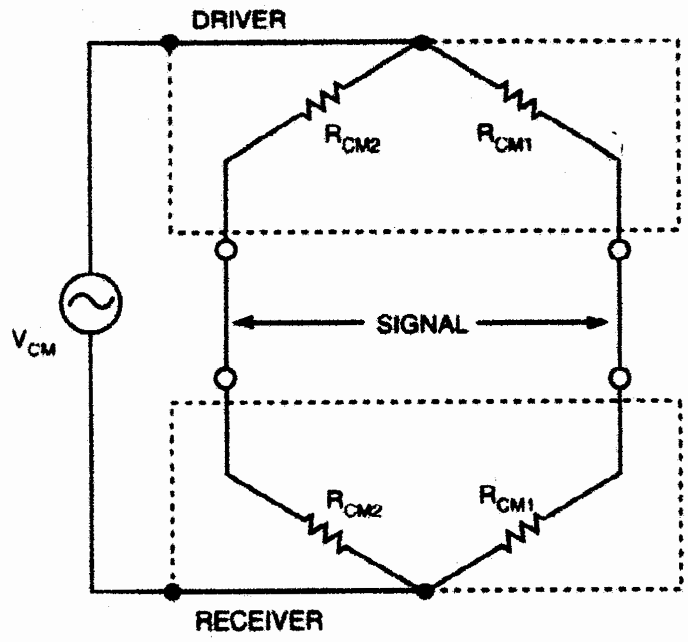


Fig. 2

By recasting the topology, you can see how a Wheatstone bridge results from the impedances to ground driver and receiver.

"Floating" devices use two-wire power cords, and each chassis assumes an open-circuit voltage as high as 120\ with respect to safety ground. If you externally ground any accessible point, including signal connectors, curren limited to about 1 mA. This current can cause you an unpleasant but harmless shock. If you leave it ungroundec parasitic power-line current flows only in any cables you use to connect the devices. You do not eliminate-only reroute-the parasitic current flow. In unbalanced interfaces, even a few microamperes of interchassis current ca significantly degrade dynamic range. At low frequencies, floating devices are effectively high-impedance curren sources with short-circuit currents as high as 1 mA and open-circuit voltages as high as 120V.

In larger systems, the flow of power-line parasitic currents becomes more complicated. In any case, assume tha significant currents flow in any interchassis connections and that significant voltages exist between the local-de\ grounds.

Unbalanced, or single-ended, interfaces are common, presumably because they are inexpensive and often perfor acceptably well in small systems. They prevail in consumer audio systems, most video and RF systems, many da systems, and, unfortunately, in most electronic instruments.

All unbalanced interfaces suffer from common-impedance coupling, in which the grounded conductor of the interconnecting cable, as well as the contact resistance of any connectors, becomes the offending common impe (**Figure 1**). Because the cable shield is effectively connecting the device chassis, the shield has either noisy inter current - from floating devices - flowing through it or noisy interchassis voltage – from grounded devices – imp across it. The resulting noise voltage across the shield directly adds to the signal. Consider20ft of cable, having ohm/ft which a 500-uA interchassis current flows. This resistance shield resistance, connecting two floating dev between adds 250-uV noise, which is only 61 dB lower than a 300-mV cosumer-audio reference signal.

You can reduce this noise coupling by shortening the cable, which reduces both the resistance and the inductan the shield. Using cable with a heavier gauge shield improves matters at low frequencies but has little effect at hi frequencies. At power frequencies, wire impedance roughly equals dc resistance, which decreases by a factor of every 10 AWG decrease in gauge. For example, if you replace a #12 with a #2 AWG – measuring 0.26 in. in di – you reduce hum by about 20 dB. However, at RF, inductance determines wire impedance. Diameter has little on inductance, which is proportional to length. For example, 8 ft of #10 AWG has an impedance of about 22 ol MHz –

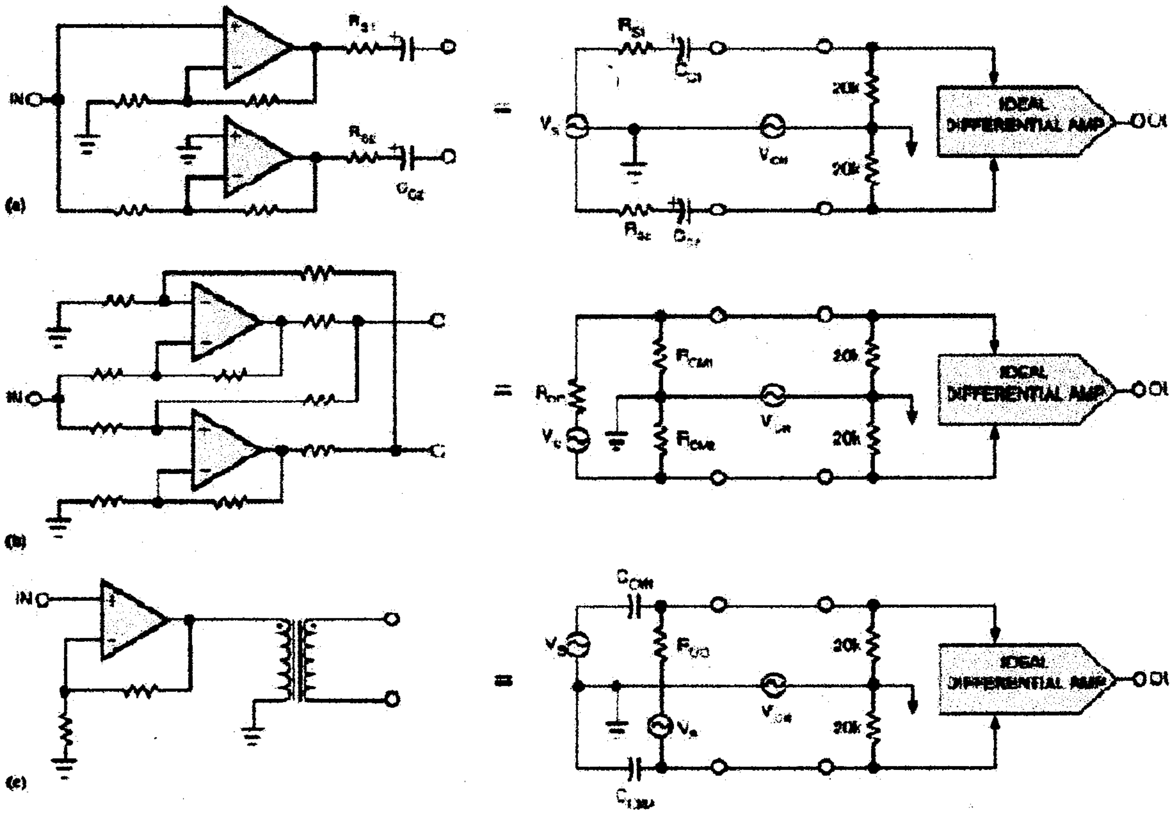


Fig. 3

Your design can have grounded (a), active-floating (b), or transformer-floating drivers (c), as their schei and equivalent circuits indicate.

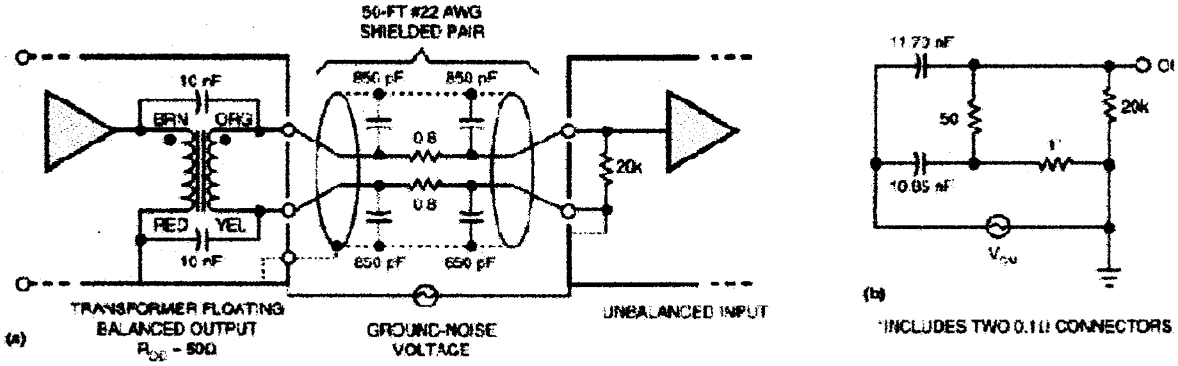


Fig. 4

The transformer-floating driver yields significant attenuation of 60-Hz ground noise (a), as this analysis equivalent circuit (b) shows.

the AM broadcast band. Replacing this #10 with #0000 AWG – measuring about 1/2 in. in diameter – reduces t impedance only to about 18 ohm.

Because the shield impedance rises with frequency and the power-line noise is essentially capacitively coupled, common-impedance coupling is generally efficient at coupling power-line "trash" at 100 kHz to 10 MHz. Most devices' performance suffers when you couple such noise to their inputs. For example, audio systems sometime demodulate this conducted RFI – producing clicks, pops, or buzzes. More often, the RFI results in subtle intermodulation distortions, in which listeners describe the reproduced audio as having a veiled or grainy qualit **(Reference 2)**.

Another way to reduce impedance is to decrease interchassis current flow through the shield. If either or both d are floating, you can reduce parasitic power-line-to-chassis capacitances by specifying a low-capacitance split-b power transformer, for example, in the floating devices. If both devices are grounded, reducing interchassis vol powering both devices from the same outlet may help. But a transformer or another ground-isolation device in signal path – effectively making the interface differential – may be required to substantially reduce noise-current the shield. In audio systems using unbalanced interfaces, 3-ft-long cables often cause hum, and 20-ft-long cable existence of two or more grounded devices in the system practically guarantees hum. In these cases, only a grou isolator can reduce parasitic shield currents enough to make hum inaudible. For high-dynamic-range systems wi signal frequencies lower than 10 MHz, unbalanced interfaces pose enormous practical problems in controlling common-impedance coupling.

Balanced interfaces have pitfalls, too

The use of balanced line drivers, balanced twisted-pair cables, and balanced line receivers is a long-standing pra professional-audio and many other systems. In theory, these balanced or differential interfaces are the perfect sc to the interchassis-ground-noise problem. However, many misconceptions exist about several important details reducing the theory to practice, often causing poor system performance **(Reference 3)**.

In audio and most other systems, the purpose of the interface is to transfer maximum signal voltage. This methc "voltage matching," requires a low output impedance at the driver and a high input impedance at the receiver. I confuse this approach with impedance matching, in which driver-output and receiver-input impedances are equa maximum *power* is transferred. High-frequency interfaces may need to use this approach, although it wastes hal driver voltage, because of transmission-line effects.

Although you can define "balanced lines" in many ways, all such definitions require that the two lines have equa impedances to a reference point, usually ground. This property enables rejection of ground noise between the d and the receiver. In the rearranged schematic of the basic balanced interface, the impedances to ground of the d and receiver plus a differential responding amplifier form a Wheatstone bridge (Figure 2). Thus, if the bridge is perfectly ratio-matched or nulled, the differential amplifier sees identical ground noise at its two inputs. Under t conditions, an ideal differential amplifier would have zero output, and the interface would have infinite commor rejection. If the bridge is not perfectly nulled, some of the ground noise is converted to differential signal.

For example, a ratio mismatch of 1% results in a CMRR of only 40 dB. (CMRR is the ratio of differential or no mode signal gain to common-mode gain.) Achieving high CMRR requires extreme precision. It's important to r the bridge-nulling impedances are the *common-mode impedances* of the driver and receiver. Their differential impedances simply appear across the line and do not accomplish the nulling. Bear this in mind when you consid nulling behavior of the bridge.

A Wheatstone bridge is most sensitive to small fractional impedance changes in one of its arms when all arms h same impedance. It is least sensitive when upper and lower arms have widely differing impedances – for exampl upper arms approach zero impedance or lower arms approach infinite impedance. You must change these impe in pairs because the impedance ratios of the two sides must match for the bridge to null. Therefore, you can *mi* the sensitivity of a balanced system to small impedance imbalances by making common-mode impedances as lo possible at one end of the line and as high as possible at the other. Fortunately, this condition is consistent with driver output and high receiver input that normal-mode impedances required for efficient signal-voltage transfe the interface.

It's important to realize that a balanced interface is a subsystem. Everything that connects to the balanced line n rigorously maintain its impedance balance for you to achieve maximum interface CMRR. This balance applies t line driver, the interconnecting cable, and the line receiver. Maintaining this balance is especially important if yo to freely interconnect various devices and interchange cables, as is usually the case. Practical balanced line drive cables, and receivers have finite and imperfectly matched impedances that play a critical role in limiting the CM performance of balanced interfaces.

Rejecting common-mode noise does *not* require symmetrical signal swings on the balanced lines. High-CMRR l interfaces can have the signal on either line or symmetrically on both lines. The presence or absence of a normal signal has nothing to do with rejection of common-mode noise. You need to consider signal symmetry only in t context of cable shields or crosstalk.

Practical line drivers do not have zero normal- and common-mode output impedances. Balanced audio systems have grounded, active-floating, and transformer-floating drivers (**Figure 3**). Each has a typical normal-mode (si output impedance of 50 to 600 Ohm. For audio systems, you should carefully consider how a balanced driver b when it drives an unbalanced input – where one output line is grounded. Any line driver should incorporate cur limiting or thermal shutdown to prevent damage or failure from driving a grounded line.

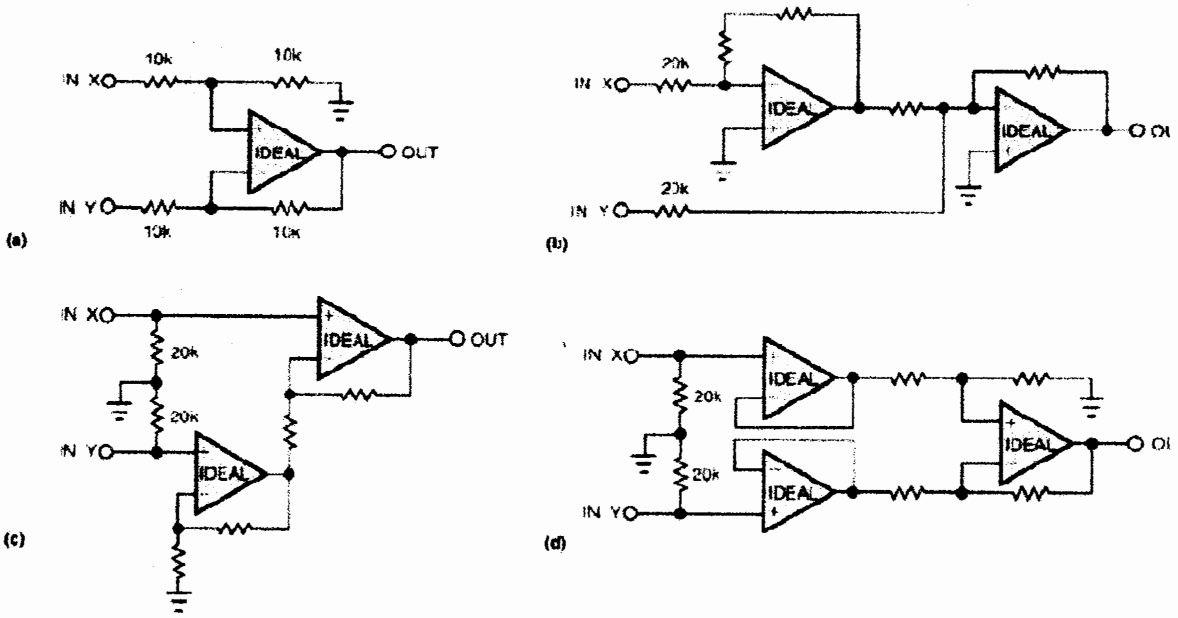
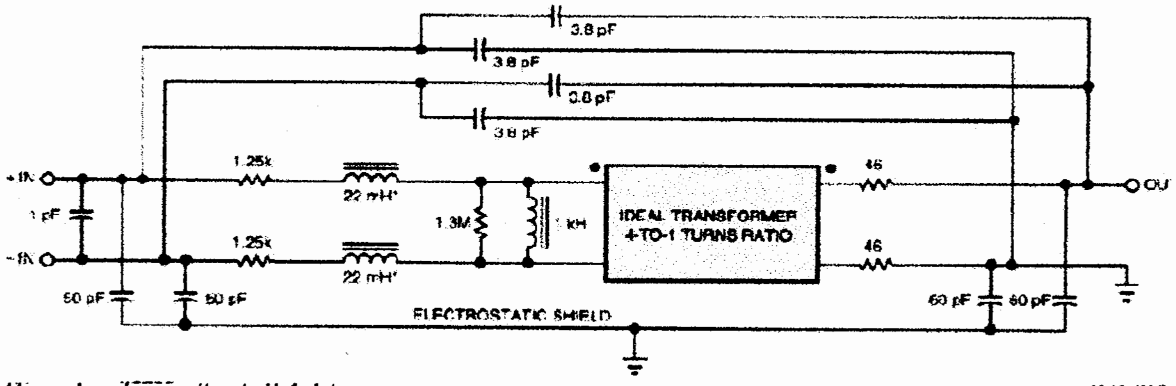


Fig. 5

By using one, two, or three op amps, you can build a simple single op amp (a), a current-mode dual op a (b), a voltage- mode dual op amp (c), or an "instrumentation," or triple, op amp (d).



NOTES:

* INDUCTANCES VARY INVERSELY WITH FREQUENCY, INCREASING AT APPROXIMATELY 3 db PER OCTAVE DOWN TO APPROXIMATELY 1 Hz.

VALUE INDICATED AS 1 kH APPLIES AT 20 Hz, AND VALUE INDICATED AS 22 mH APPLIES AT 100 kHz.
VALUES SHOWN ARE TYPICAL.

Fig. 6

The detailed model of an audio-input transformer lets you analyze ground noise.

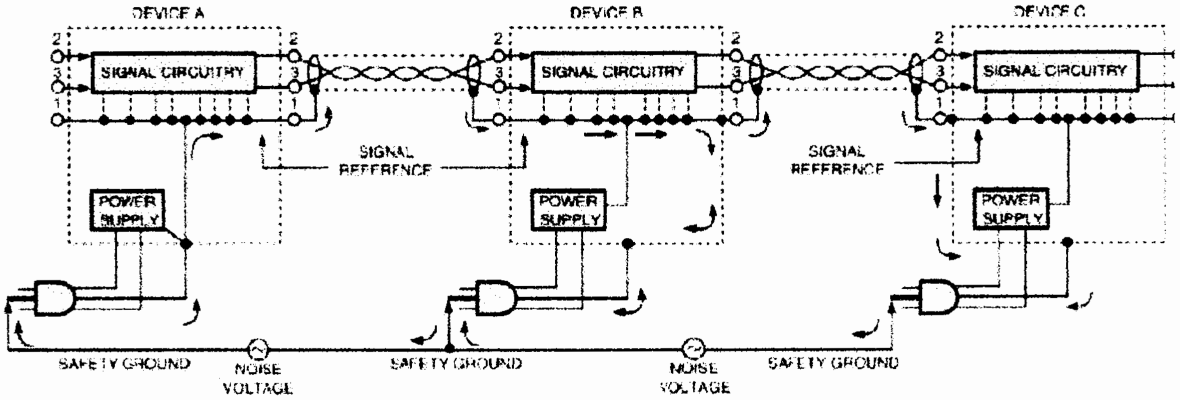


Fig. 7

Noise currents have many devious paths from shields to ground; each can cause circuit-performance pr

The grounded driver uses two antiphase voltage sources, each referenced to driver ground. Common-mode out impedances are RS1 and RS2, and normal-mode output impedance, ROD equals the sum of RS1 plus RS2. Bec RS1 and RS2 are typically 20 to 100 Ohm each with tolerances of +/-1 to +/-10%, their imbalance can be 0.4 to Ohm, resulting in an interface CMRR of 60 to 94 dB with the ideal receiver. If you use no output-coupling cap the CMRR is constant over frequency. Using electrolytic coupling capacitors with their +20% or worse toleran substantially degrade low-frequency CMRR. If you ground either output, the driver forces abnormally high – ar possibly distorted – signal current into the distant grounded input. Because this current must return to the drive may flow in an undefined return path and generate system crosstalk.

The active-floating driver uses a basic circuit comprising two op amps cross-coupled with both negative and po feedback to emulate a floating voltage source. Rod is typically 50 to 100 Ohm. Trimming the feedback resistors increases RCM1 and RCM2, and this trimming also affects the output-signal symmetry. Although manufacturer directly specify RCM1 and RCM2 one manufacturer of an IC version specifies output common-mode rejection signal symmetry under specified conditions, which allows you to determine RCM1 and RCM2 values with a cir simulator (**Reference 4**). A simulation obtained values of 5.3 and 58.5 kOhm, respectively, for a simulated part had slightly better than typical specifications. This driver produced a 57-dB interface CMRR that is constant ov frequency. Although it attempts to emulate a transformer, this driver can become unstable or oscillate when it d remotely grounded cable. To guarantee stability, you must ground the output line at the driver itself (**Reference 5**). In this configuration, the driver becomes an unbalanced output, subject to all its common-impedance-coupli problems.

The transformer-floating driver uses a transformer whose primary is driven by a single-ended voltage source. T winding resistance is typically 35 to 100 Ohm, and the two inter-winding capacitances are 7 to 20 nF each and matched to within 2% for typical bifilar-wound transformers. The interface CMRR is 110 to 120 dB at 20 Hz, decreasing at 6 dB per octave – because imbalances are capacitive – to 85 to 95 dB at 500 Hz or higher. Becau ungrounded, the transformer secondary can reference the signal to a remote input ground. The 60-Hz ground n attenuated by more than 70 dB in a typical situation (**Figure 4**). Because the noise is coupled through CCM1 attenuation decreases with frequency.

Capacitances, especially those to the shield in shielded cables, can seriously affect balanced-interface performan related issue is the long-standing debate in the audio industry about which end of a balanced cable should have i shield grounded (see **sidebar** "Which end to ground?").

Balanced interfaces can be immune to cable-induced noise if the two conductors have identical exposure to the magnetic or electrostatic fields. Tight twisting of the signal conductors tends to average the exposures, especial fields from a distant source. When the magnetic-field source is close, "star-quad" cable construction – which pa two twisted pairs to further reduce pickup cross-section – can add about 40 dB of immunity. Remember that pc and connectors are magnetically vulnerable because they create untwisted loops. Foil or braid cable shielding pr against electrostatic, not magnetic, fields. If you ground the shield at both ends, any current flow in the "drain" foil-shielded types can induce normal-mode noise. A braided shield is preferable in such cases.

The line receiver is the most important part of a balanced interface. As the Wheatstone-bridge-equivalent circui the common-mode input impedances of the line receiver, and not its circuit topology, determine the circuit's performance in real-world interfaces. One basic type of differential amplifier, an active circuit, consists of op an precision-resistor networks and performs algebraic subtraction of the two input signals. The other basic type, a transformer, is an inherently differential device that also provides electrical isolation between input and output s

You can realize active differential amplifiers with a number of well-known topologies if all the op amps are idea resistors values are exact. Eliminating these sources of error, the CMRR performance of all the circuits is identi all have 20-kOhm common-mode input impedance (**Figure 5**). Although many devices are differential in charac all can solve the basic instrumentation problem (**Reference 6**). All of these circuits have CMRR that is sensitive source-impedance imbalances. Because audio-signal sources routinely have imbalances of 0.2 to 20 Ohm, these balanced inputs rarely deliver their advertised CMRR. You should also be aware of some other limitations of th active circuits (see **sidebar,** "Active circuits have limitations, too"). Alternatively, look at a new design for a ba line receiver (see **sidebar** "A new 'active transformer' for audio can help").

Because the primary of a transformer floats, the primary inherently responds differentially, and any amplifier pre by a transformer becomes a differential amplifier. The transformer requires no trimming, and its CMRR is stable life. Figure 6 is a circuit-simulator model of a high-performance audio-input transformer. In such a model, the 5 primary-to-Faraday-shield capacitances determine the common-mode input impedances. These impedances are 50 MOhm at 60 Hz and 1 MOhm at 3 kHz, making the transformers relatively insensitive to driver impedance imbalances. For inputs that are fed by sources that must be interchangeable or have unknown impedance imbala you should consider transformers because of this insensitivity. Although audio and other low-frequency transfoı are bulky and expensive, they have several other advantages. For example, they can transform or match line imp to the optimum source impedance for the subsequent amplifier by the square of the transformer's turns ratio, th maximizing SNR and dynamic range. A transformer-coupled, balanced input stage operating from +/-15V rails easily attain 140 dB of dynamic range.

Transformers also have inherent RF common-mode attenuation. Because CMRR compares normal-mode with common-mode responses, it's generally not a useful measure of this attenuation. Typical normal-mode – 3-dB bandwidth is 100 to 200 kHz, but common-mode attenuation is more than 30 dB from 200 kHz to 10 MHz. Ar advantage of transformers is that their maximum common-mode input voltage is limited only by internal insulati is typically several hundred volts.

You should evaluate the performance of a balanced input such that test results become a reasonable predictor o performance in a real-world system. Many common test procedures miss this important point. For example, IE(Standard 268-3 for sound-system equipment specifies a CMRR test that tweaks the generator impedances to pr maximum reading. In real systems, this approach simply isn't practical, and you would have to make this

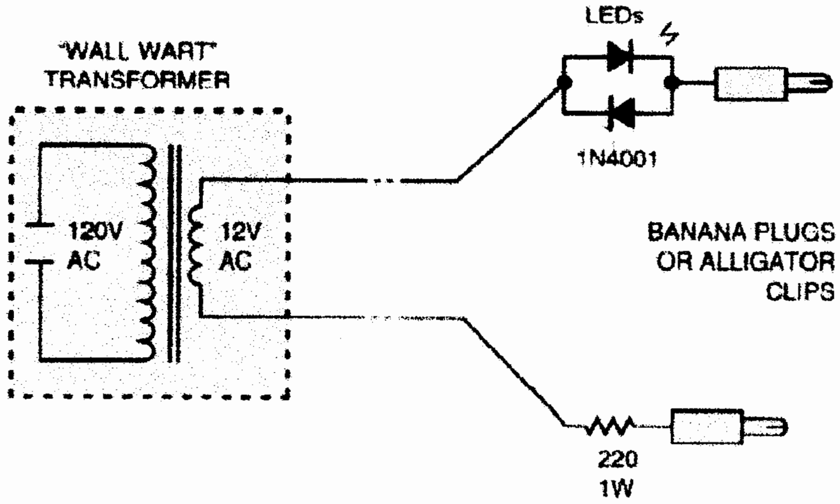


Fig. 8

By injecting a small amount of line current into coupling paths, you can observe changes in noise floor a thus, potential system problems.

adjustment whenever you reconfigure the system. Some engineers use an even more absurd test – simply shorti input terminals to each other and connecting a signal generator between the inputs and ground – to determine C This test assumes zero source imbalance, a situation you are unlikely to encounter, even in a nonreconfigurable No wonder so many engineers think balanced interfaces are not worth the effort to design.

Common-impedance coupling can catch you, too

The noisy, parasitic power-line currents that flow between interconnected system devices can couple to the sigr *inside* each device. An Audio Engineering Society paper about this form of common-impedance coupling dubs "Pin 1 problem," because Pin 1 is the shield connection in standard XLR audio connectors (**Reference** 7). An ir survey indicates that about 50% of commercial audio equipment has a significant amount of this problem. It car either unbalanced or balanced interfaces and often interacts with other noise-coupling mechanisms to frustrate y attempts to solve system problems, although the two problems do occasionally cancel each other out. This prob most likely to reveal itself when inputs or outputs use cables with the shield connected at both ends. Were it not Pin 1 problem, this common practice would rarely significantly degrade CMRR in balanced interfaces, except w consider the Pin 1 problem.

"Pin 1" is the terminal at an equipment input or output to which the cable shield connects when you insert a mai cable connector. For standard XLR connectors, it is literally Pin 1; for 1/4-in. phone connectors, the sleeve; anc BNC and consumer RCA/IHF connectors, the outer shell. In a system hookup, noise currents can flow in these connections (**Figure** 7). In devices A and B, noise currents flow in internal-signal reference-ground conductors. currents cause voltage drops in wiring or pc-board traces and couple to the signal path, sometimes at a high gai that, for device B, even placing a so-called ground lifter on the power cord (not recommended) would not solve problem, and current would still flow between input Pin 1 and output Pin 1. Devices B and C also have another common-impedance coupling problem, because they allow noise currents coupled from the power line by the pc supply to flow in signal- reference ground.

You avoid these problems by restructuring the grounds. The noise currents still flow but not in signal-reference conductors. For example, you can use a "star" connection of shields, safety ground, power-supply common, anc chassis. This arrangement is effective and results in the lowest possible coupling. Alternatively, you can modify technique to use chassis metal as low RF-impedance connections for the shields. Before the advent of pc-board mounted connectors, the metal chassis provided these trouble-free connections. You can also use an RC shield-termination network to ground the shield at high frequencies only.

The schematic symbol for ground can be deceptive. It's tempting for an engineer to assume that the voltage at ε ground symbols is exactly the same. In reality, wires and pc-board traces having resistance and inductance conn these "ground" points to each other. Haphazard ground connections create problems. The Pin 1 problem, effect turning a shield connection into a low-impedance input, still exists because standard product-development tests reveal it. However, a simple device called a "hummer" can reveal it (**Figure 8** and **Reference 8**). This device fo about 50 mA of line current through possible common-impedance coupling paths in a device. Any change in the device's output noise floor indicates a problem.

Electronic systems are becoming more complex, the electrical environment is becoming more hostile, and custo at least in the audio industry – are expecting greater dynamic range. Engineers must recognize and solve proble may show themselves only when a user installs the product in a system. Blaming such problems on bad groundi dirty ac power is a poor excuse for failing to perform realistic product testing. The following guidelines will hel minimize your frustration and maximize performance:

- Accept the existence in the real world of power-line noise and RFI.
- Recognize the inherent limitations of unbalanced interfaces.
- Design all line drivers to have the lowest possible output impedances.
- Design balanced line receivers for the highest possible common-mode input impedances.
- Specify balanced-interface cables with minimal CMRR degradation.
- Minimize power-supply-to-chassis parasitic capacitances.
- Provide chassis ground terminal, especially for floating devices.
- *Never* disconnect a safety ground to solve a noise problem.

You can prevent most of the interface problems a user might otherwise face. If your system is burdened with th problems, system-friendly interfaces quickly become more important than the gee-whiz features of a product. W well- designed, real-world product interfaces, a plug-and-play utopia could exist.

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Author's biography

Bill Whitlock became president of Jensen Transformers Inc (Van Nuys, CA) in 1989. At Jensen, he designs and develops audio, video, and other signal-interface devices. He graduated from Pinellas County Technical Insti (Clearwater, FL) in 1965 and started his career in professional audio in 1972. In his spare time, he enjoys hik sailing, music, and restoring 1950s radios and early hi-fi gear.

WHICH END TO GROUND?

When designing an analog interface, you have to decide to ground the shielded cable at the driver end, the rec end, or both. Consider a scenario using a common audio-industry shielded twisted-pair cable. This cable has a common-mode (both inner conductors to shield) capacitance of about 67 pF/ft. Capacitance measurements on t samples from different manufacturers show an imbalance of about **4%** between the signal conductors. Accordir one of the manufacturers, normal manufacturing tolerances cause this imbalance, and you can expect similar iml in any commercial cable. If you ground the shield at the receiver end only, these 4% mismatched capacitances a much as 20% mismatched driver common-mode output impedances form a pair of lowpass filters for common-r noise (**Figure A**). Any mismatch between the two filtérs results in common-mode conversion, degrading interfa CMRR. However, if you ground the cable shield only at the driver end, common-mode noise does not appear a the cable capacitances, and this approach forms no lowpass filters (**Figure B**).

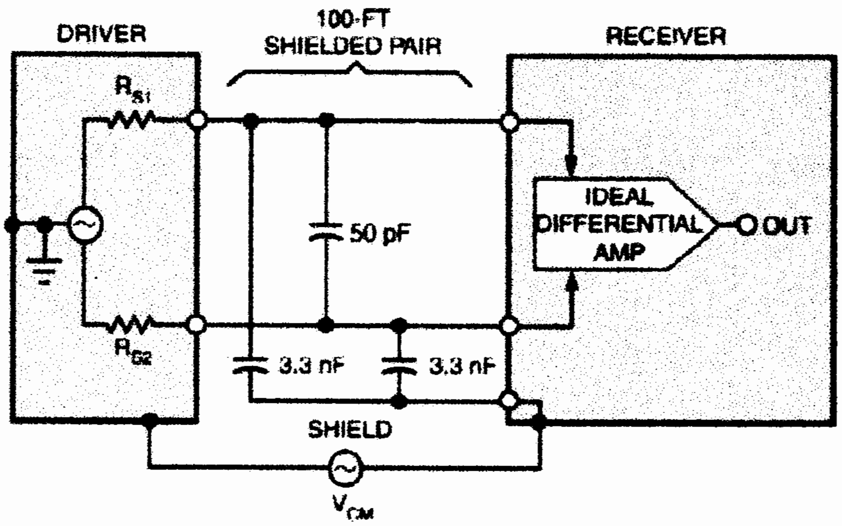


Figure A

Grounding the shield at the receiver end results in this equivalent circuit.

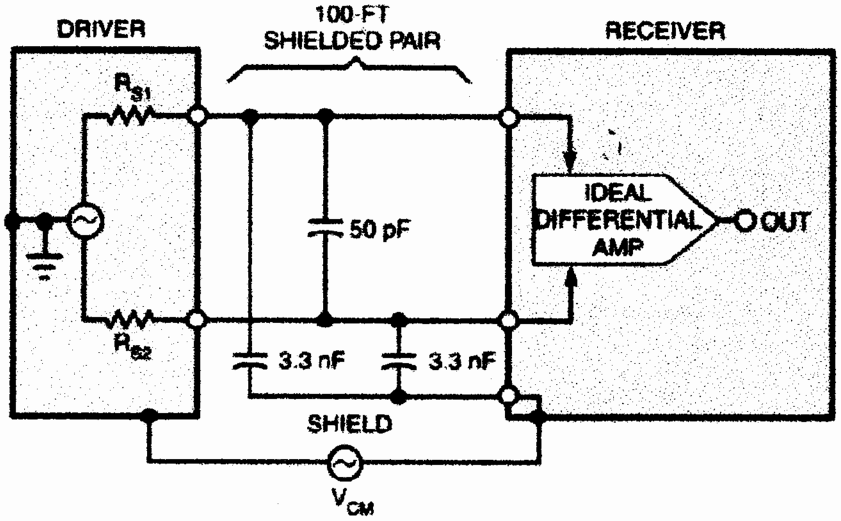


Figure B

Grounding the shield at the driver end results in a subtly different model from the one in Figure A.

If the driver produces perfectly symmetrical signal swings on the two lines and the cable capacitances are exactl matched, no signal current flows in the shield. In reality, though, signal currents flow in the shield, whether beca capacitive imbalances or because of signal asymmetry, and this flow increases with frequency. Because the curr must flow back to the driver, a shield connection there makes the path direct. If you ground the shield at the rec end only, the current may take an undefined path, causing system crosstalk, instability, or oscillation.

Grounding the shields at both ends creates some interesting trade-offs and potential problems. The problems of receiver-end grounding occur – but to a smaller extent. In poorly grounded systems or those with inputs having RFI immunity, there may be advantages. If no other grounding path exists between the chassis of two devices, u the cable shield to connect them reduces the common-mode ground noise between them, even though this appr degrades system CMRR. It would be preferable to use some means, such as power safety ground, to connect th Floating devices, such as those with two-prong ac plugs, are the most offensive in this regard, creating large co mode voltages unless grounded. In some situations, grounding at both ends may have an **RFI** advantage. If the floats at the receiver end, the cable itself can become a whip antenna, creating large RF common-mode voltages input. AM radio is the usual offender in these cases. However, a simple R-C network of a 500 resistor in series 10-nF capacitor between the shield and physical ground point other effectively terminates the shield for frequen higher than approximately 300 kHz, spoiling the efficiency of the antenna and keeping the circuit open at lower frequencies **(Reference A).**

Reference

A. Morrison, R, *Grounding and Shielding Techniques in Instrumentation,* Third Edition, John Wiley *R* Sons, 1 pg. 86.

ACTIVE CIRCUITS HAVE LIMITATIONS, TOO

Active circuits have some limitations. The single-op-amp and current-mode dual-op-amp circuits must trade common-mode input impedance for increased thermal noise caused by higher value resistors. For example, if yc double resistor values, thus decreasing CMRR sensitivity to source imbalances, noise increases by 3 dB. Using electrolytic capacitors for coupling at any of the inputs may degrade low-frequency CMRR because of the capa loose tolerances and poor aging characteristics. Adding capacitors from each input to ground to suppress RFI c common-mode input impedances to be unbalanced unless you carefully match these impedances. Because they lower common-mode input impedances at high frequencies, these capacitors degrade CMRR from source imbal

Common-mode input-voltage range is a few volts less than the power-rail value in most circuits. At high sign levels, this range can approach zero because the limit actually applies to the sum of the peak normal-mode and common-mode voltages at each input (**References A and B**). This situation can be a problem in electrically hos environments.

References

A. Graeme, J, Applications of Operational Amplifiers – Third Generation Techniques, McGraw-Hill, 1973, pgs 57.

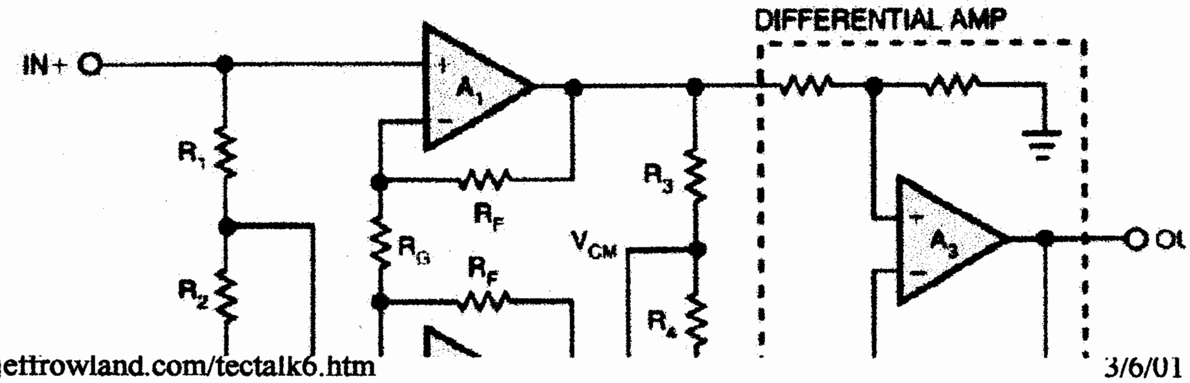
B. Perkins, C, "To Hum or Not to Hum," Sound & Video Contractor, March 15, 1986, pg 42.

A NEW "ACTIVE TRANSFORMER" FOR AUDIO CAN HELP

Line receivers with high common-mode input impedances clearly produce higher CMRR in real-world balanc interfaces. General-purpose balanced audio receivers must satisfy a number of conditions to be practical. For ex the circuit cannot rely on a dc path to ground from the driver. The driver output might be a floating-transforme secondary or floating coupling capacitors, or the output might simply be unplugged. Because these outputs may leakage currents and active inputs must have dc-bias current paths, the common-mode input impedances of acti receivers have historically been much lower than those of a transformer.

A new, patented active-input circuit uses bootstrapping to raise ac common-mode impedances to tens of megac and maintains a low-resistance dc path at the inputs. The circuit is only slightly more complex than traditional c requires no additional tightly matched components, and enables effective and novel RFI suppression. Its in-syst CMRR performance rivals that of the finest transformers, yet its signal path has response extending to dc.

The circuit is built around a conventional instrumentation amplifier (**Figure A**). Al-though you can independent bootstrap each input, this approach requires two tightly matched electrolytic capacitors to maintain balance. Ho regardless of any differential gain set by Rf and Rg, the common-mode gain of A1 and A2 is unity. This circuit simultaneously bootstrap R1 and R2 with the buffered common-mode output of A1 and A2 and a single capacit is not in the differential-signal path. This technique not only improves performance but gives the IC version of t circuit a parts count of two.



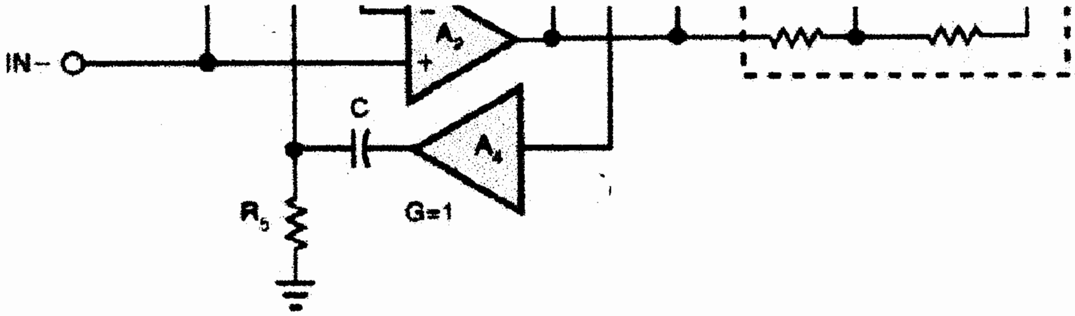


Figure A

This circuit yields high ac common-mode impedances yet leaves a low-resistance path to ground for dc.

At dc, common-mode input impedance is simply R1 or R2+R5. If common-mode gain, G, is unity, the effective of R1 and R2 approach infinity for frequencies much higher than the cutoff for the highpass filter formed by C a The following describes the common-mode input impedance, Zi, at any frequency, f, and gain, G (through A1 c and A4):

$$Z_i = (R_1 + R_2)\sqrt{\frac{1+\left(\frac{f}{f_N}\right)^2}{1+(1-G)^2\left(\frac{f}{f_D}\right)^2}},$$

$$\text{where } f_N = \frac{1}{2\pi\left(\frac{R_1 \bullet R_2}{R_1 + R_2}\right)C} \quad f_D = \frac{1}{2\pi R_1 C}.$$

For example, if R1, R2 and R5 are 10 kOhm each, the dc input resistances are 20 kOhm each, providing a path amplifier bias currents as well as for any leakage currents from the signal source. At power-line frequencies, the bootstrap can increase common-mode impedances to more than 10 MOhm in practical circuits. As the equation indicates, this impedance is ultimately limited by the gain-bandwidth products of amplifiers A1, A2, and A4.

Compared with the simple differential-amplifier design, this circuit improves real-world CMRR by orders of ma and has lower thermal noise as well because you can make the differential-amp resistors much lower in value. T circuit is covered by US Patent 5,568,561 and other pending patents. Direct inquiries about the IC or licensing Corp (Marlborough, MA, www.thatcorp.com).

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