Topological Analysis of Consumer Audio Electronics: Another Approach to Show that Modern Audio Electronics Are Acoustically Transparent 4053 (C-2)

David A. Rich and Peter Aczel
The Audio Critic
Quakertown, PA 18951, USA

Presented at the 99th Convention 1995 October 6–9 New York

AES

This preprint has been reproduced from the author's advance manuscript, without editing, corrections or consideration by the Review Board. The AES takes no responsibility for the contents.

Additional preprints may be obtained by sending request and remittance to the Audio Engineering Society, 60 East 42nd St., New York, New York 10165-2520, USA.

AN AUDIO ENGINEERING SOCIETY PREPRINT

Topological Analysis of Consumer Audio Electronics: Another Approach to Show that Modern Audio Electronics Are Acoustically Transparent

DAVID A. RICH, *AES Member,* **AND PETER ACZEL,** *AES Life Member*

The Audio Critic, 1380 Masi Road, Quakertown, PA 18951, USA

ABSTRACT

The circuit topologies of audio electronics that have been judged to have good sound quality in open-loop subjective tests are examined to identify any consistently recurring attribute of the designs that might be responsible for their claimed sonic superiority. No such topological feature was identified, thus confirming the results of controlled listening tests which show these devices to have no sonic signature.

0 INTRODUCTION

Controlled listening tests have consistently shown that electrical components will be audibly indistinguishable if they have: (1) flat frequency response, (2) noise and distortion levels below audible thresholds, (3) high input impedance and low output impedance [1]. Despite this, many audiophiles still hold that audible differences exist. Audiophiles present a number of spurious reasons for their belief that controlled listening tests fail to reveal the differences in the components. No amount of reasoning, by professionals in the field, have dissuaded audiophiles in the belief that the sonic differences exist. One reason for the audiophiles' continued belief in sonic differences is that they are exposed to a significant amount of technical or semitechnical information about why the components sound better, from the manufacturers, dealers, and the press. The designers claim that some undiscovered "X factor"—that is, undiscovered except by the designer—must be considered to achieve good sound quality [2].

In this paper we examine the circuit topologies of audio equipment. Our goal is to see, by examining the circuit topologies, if some new design approach is being used that would not have been used if the design had been developed simply to achieve good results in traditional bench measurements. We concentrated the analysis on equipment that had received favorable reviews as a result of open-loop listening tests. It should be noted that open-loop listening tests of the same piece of equipment often lead to very different conclusions and reviews. Those who support the open-loop listening method can only cite loudspeaker reviews as consistent [3].

None of the audio equipment examined in this paper has been designed to change the sound intentionally by introducing frequency-response errors or nonlinear distortion. At first blush it would appear obvious that a designer would not purposely color the sound of his design, but it has been observed that it becomes is easier to sell products in the high-end market if they do change the sound. The single-ended class A 20-watt triode amplifier craze is an example of this. Audiophiles have a tendency to assume that if a product sounds different it must be better [2].

Our surveys of the designs said to "sound good" have shown little commonality among the topologies [4], [5], [6]. The surveys yielded a comparison of basic topologies, shown for preamplifiers in Table 1 and for power amplifiers in Table 2. A quick scan will show how random the design topologies are. If a designer had discovered a unique topology that sounded better but did not measure better, we would expect that topology to spread to other companies through "reverse engineering" of the product with the superior sound.

It will be shown in the analysis below that none of the design techniques found in the examined audio components can be expected do anything that would affect the input/output transfer characteristic of the electronics in a way that could affect the sound quality of the electronics—unless they also resulted in a change in the measured performance. The results of the study of the circuits in this paper will thus confirm the results of the controlled double-blind tests which have shown that no sonic differences exist in audio components that measure well.

1 A REVIEW OF THE STANDARD PRACTICE

Although nothing truly unique can be found in the design of audio components, a number of circuit-design techniques can be seen in audio equipment that are distinct from common design practice. Before examining these differences, we shall first review the most common topology used for amplification of audio signals when good performance on standard measurements is the goal of the design.

The basic topology [7] is shown in Fig. 1. This topology is commonly used in audio-band integrated operational amplifiers. More complicated designs with three voltage-gain stages (NE5534 for example) are often used to achieve very low noise floors or operation at low supply voltages. The lack of a good *pnp* device in the standard IC process also results in deviation from this standard topology. When the tight matching properties of the transistors on a die in an integrated circuit are exploited, modifications of this topology can result in very low noise and 16-bit settling [8] with a single gain stage. It should be noted that this design by Scott Wurcer also requires a process that has high-performance *pnp* devices. The topology in Fig. 1 is also used in high-power audio power amplifiers, where the design may be fully discrete or a combination of an integrated voltage-gain stage with discrete output transistors.

Q_1 and Q_2 form a standard differential pair. Q_3 and Q_4 form an active load for this stage. The gain of this stage is set by the g_m of Q_1, the output impedance of transistors Q_2 and Q_4, and the load from the next stage. Since the output resistance of Q_2 and Q_4 will be large, the gain of the circuit will be large at low frequencies, provided the circuit is not loaded by the next stage. Q_5 and current source I_2 form a follower stage that keeps the input impedance of the second stage from loading the first stage. In power amplifiers the follower also isolates the input stage from the nonlinear base-emitter junction capacitance of the second stage. This nonlinear capacitance, if not isolated, can lead to distortion, given the large voltage swings at the output of the second stage [9]. Cascoding the second gain stage will also eliminate this source of distortion by keeping the V_{CE} of Q_6 relatively constant. The differential pair in a power amplifier may also be cascoded in order to keep the V_{CE} (or V_{DS} if FETS are used) of Q_1 and Q_2 at a low enough voltage so that high-speed, low-noise devices can be used.

Transistors Q_3 and Q_4, in addition to providing an active load, also perform the differential-to-single-ended conversion for the first stage. This significantly improves the CMRR (common-mode rejection ratio) for the first stage in comparison with a stage that does not have this circuit [7]. High CMRR in a differential stage, in addition to requiring a good differential-to-single-ended converter, also requires that the tail have a high output resistance [7]. For this reason a current source is used for the tail, instead of a resistor.

High CMRR is important when an amplifier is in a noninverting feedback configuration. In this configuration the common-mode swing is equal to the input signal swing. If the differential amplifier pair transmits some common-mode signal to the second gain stage, then the second stage will be unable to distinguish between the signal resulting from the differential-mode gain of the differential pair and the signal resulting from the common-mode gain of the differential pair. This will cause the feedback loop to respond to the common-mode signal that has been amplified by the first stage. Consequently, the output will be distorted because the amplifier no longer attempts to keep the voltage across the summing junction at zero. This distortion results because the voltage across the summing junction varies as the common-mode signal at the input of the differential pair varies. Note that the small-signal common-mode gain of a differential pair usually varies across the common-mode voltage range of the amplifier, adding to the distortion effect.

Q_6 and the current source I_2 form the second gain stage. As in the first stage, the low-frequency gain of the stage will be large because the load on the stage comes from the output impedance of Q_6, the output impedance of the current source I_2, and the input impedance of Q_7. Q_7 and I_4 form a follower circuit that prevents the second gain stage from being loaded by the output stage. In power amplifiers, cascoding the second stage will reduce distortion by isolating the active element (Q_6) from the large swings at the output of the stage. Added cost and reduced maximum signal swing are the disadvantages of the cascode stage.

The output stage is a class A/B push-pull stage. It is formed with Q_8 and Q_9 and is biased by D_1 and D_2. C_1 compensates the amplifier. The capacitor is in the Miller feedback configuration. This configuration reduces the size of the capacitor, which is an advantage in an integrated design and also moves the nondominant open-loop pole of the amplifier to a higher frequency [10].

2 HIGH-END AUDIO DESIGN PRACTICE

Fig. 2 shows an amplifier stage that is typical of the topology used in some high-end audio products. I_1 is often a simple resistor and not an active element in these designs. A topology of this type is not available as an integrated circuit, so the circuit of Fig. 2 is a discrete design. Audiophile products are usually discrete designs because designers do not have sufficient freedom to optimize the parameters they believe are important when using commercial integrated circuits. The principal advantages of discrete design from a measurement point of view are the possibility of reduced noise levels and increased output drive capability. A high-performance, small-signal level, discrete op-amp using the standard topology is discussed by Jensen in [11]. The downside of discrete design is that the circuit will be slowed down because of larger parasitics, and it may cost more. The former problem is not a significant concern at audio frequencies.

It should be noted that many high-end products may have integrated operational amplifiers in the signal path. The op-amps may often precede or follow the more exotic discrete circuitry to be examined in this paper. Since the op-amps have none of the design features that high-end designers believe are required to prevent sonic degradation, this practice appears very strange. For example, high-end designers who hold that an amplifier must use very low levels of feedback to prevent sonic degradation may use op-amps that have high feedback levels at low frequencies in the signal path [12]. Despite their deviation from the "politically correct" techniques common to high-end design that we shall look at in this section, electronics with integrated operational amplifiers in the signal chain often receive very favorable reviews.

The first thing to be observed in Fig. 2 in comparison with Fig. 1 is that the amplifier stages have much lower low-frequency gains. This is the result of emitter degeneration of the stages (R_1 and R_3) and the resistive loading of the stages' output (R_2 and R_4). Reduction of the gain of the individual stages is an attempt to reduce the global feedback of the circuit. This concept was first advanced by Otala [13], [14], although Otala's analysis has not been accepted by some peer reviewers, such as Cherry [15], Cordell [16], [17], and Jung [18]. By resistively loading the first and second gain stages, the open-loop gain, and thus the open-loop transfer function, becomes constant up to the 10 kHz to 50 kHz range, since the dominant pole will now be present at these higher frequencies. The dominant pole position can be higher because the open-loop gain is lower. But note that the gain-bandwidth product of the amplifier remains unchanged.

Since the gain of the second stage is low, Miller compensation is not often used, and the dominant pole is usually set by a capacitor at the output of the second gain stage. Sometimes no compensation capacitor is required at all, if the open-loop transfer function is at a very low level. This is because the dominant pole of the amplifier is at a low enough frequency to compensate the amplifier, and it is therefore not required to add additional capacitance.

Often the additional poles associated with the unity-gain output stage of the amplifier will prevent compensation of the amplifier without a compensation capacitor. This problem becomes more difficult if the output stage has to drive a reactive load. Some designers work around this problem by picking the feedback off the second voltage-gain stage and not the output. While this solves the problem of dealing with the finite bandwidth of the output stage and driving reactive loads, it adds a new problem because the output stage is now running open-loop. The open-loop output stage can be a source of significant distortion. Surprisingly, some designers (Threshold and Coda, for example) have been able to create power amplifiers whose output stage runs open-loop with distortion below 1%, but these amplifiers never achieve state-of-the-art distortion numbers. Running an output stage open-loop is much more common in preamplifiers, but even so it is a technique used in only a minority of the designs. Often manufactures claim their amplifiers have no global feedback when they put the feedback loop before the output stage [19].

When the feedback network has reactive components, as in an RIAA equalizer, stabilizing the amplifier becomes more challenging because the return feedback function is no longer a constant. Picking the feedback before the output stage solves this problem. With increasing frequency, the RIAA network loads the amplifier, and the open-loop gain of the amplifier decreases. The open-loop transfer function thus has the desired property of being constant in the audio band. In some design we have examined, the open-loop transfer function decreases at frequencies below 100 Hz because the open-loop gain of the amplifier is then dominated by the output impedance of Q_3 and I_2 and not the feedback network. High distortion below 100 Hz is a consequence of this.

The approach of keeping the amplifier's open-loop transfer function constant throughout the audio band increases distortion below the dominant pole frequency. It is argued by the designers of such circuits that distortion is not increased, since the gain stages have large amounts of emitter or source degeneration in an attempt to linearize each voltage-gain stage. It is argued that this linearization allows for low distortion even when the global feedback rates can be reduced. This argument is contradicted by the well-known result from feedback theory that multiple small feedback loops will not be as effective as one global loop [20]. Some designs may have no global feedback or very small amounts of feedback (6 dB or less). In these cases designers must move beyond local feedback and use more exotic methods of error cancellation [21]. Tandberg is one manufacturer that has used this approach in many products.

The feedback rate in the various electronics examined here that were said to "sound good" varied by 4 orders of magnitude at low frequencies! Sometimes an amplifier will not receive a good review if the "open-loop" (viz., subjective, no controls) tester knows he is listening to a high-feedback design, but if the reviewer is unaware the design has circuitry that uses high feedback, then the equipment may be praised for excellent sound. An example of this was cited above—the use of operational amplifiers, which can have very high feedback rates at low frequencies, in the signal path of equipment that also has low-feedback discrete stages.

Design of input stages with a very wide open-loop linear range [18], [22] is often cited as important to a high-end designer. This is an attempt to reduce *transient intermodulation distortion.* In brief, this effect occurs when an amplifier slew limits. Early work suggested that transient intermodulation distortion could be eliminated only if small level of global feedback were used [13]. Later work showed that the effect could be eliminated if the input stage linearity was made great enough so that under worst-case conditions the summing junction is never moved outside the linear range of the input stage [23]. The amplifier cannot slew limit as long the input stage remains linear. Designers will use FET devices and/or degenerate the gain device in the front-end differential pair to achieve the wide linear range.

Some researchers have suggested that this requirement results in significant overdesign and that the transient intermodulation effect cannot occur with bandlimited music signals [24]. This explains why bipolar op-amps with no degeneration of the differential stages are acoustically transparent in controlled listening tests. Also note that sophisticated tests are not required to test for transient intermodulation distortion. If an amplifier has low levels of THD at 20 kHz on full voltage swings, it is free of transient intermodulation distortion. If an in-band test is required, then some unusual three-tone intermodulation tests can be used [9].

The amount of degeneration of the input stages of amplifiers said to "sound good" varied significantly from no degeneration on a bipolar stage to orders of magnitude beyond the emitter (or source) resistance of the active device.

Fig. 3 shows an amplifier that is designed to be fully complementary from the input stage onward. This relatively popular technique has many different forms. Often R_1 and R_6 are replaced by active-element based current sources for improved CMRR. Some designs may not include an output stage as shown in Fig. 3; others may have the stage. The fully complementary design technique may be useful in reducing distortion in amplifiers that are run at low feedback levels. One reason this is helpful in low-feedback designs is that, when large amounts of local feedback are used in the second gain stage, the voltage swing at the input of the stage (the bases of Q_5 and Q_6) must be larger, since the stage's gain is reduced. As a result, the first gain stage's output swings are higher, and this stage can now contribute significant distortion. The downside of a fully complementary amplifier is increased $1/f$ noise and dc offset, as well as decreased CMRR. The decreased CMRR results because the differential-to-single-ended conversion occurs at the second gain stage between the unmatched *npn* and *pnp* devices. In the differential pair shown in Fig. 1, this conversion is done with matched *pnp* devices. Another clear disadvantage of a fully complementary circuit is increased parts count.

While a fully complementary amplifier may be overly complex and may not always yield the best performance, one aspect of its design can lower distortion over the standard topology. This aspect of the design is the push-pull second stage. This is especially true in power amplifiers, where a large voltage swing occurs at the output of the second gain stage. Achieving a push-pull second gain stage while retaining the active load on the differential pair for high first-stage low-frequency gain is not a trivial design problem. A circuit which does this was published by Cordell [21]. A common-mode feedback circuit biases the active loads of the first stage. High CMRR is maintained by using differential pairs for both the first and second stage. Differential-to-single-ended conversion occurs in the second stage with an *npn*-based current mirror. The combination of the above with cascodes in the second stage and a buffer stage between the first and second stages results in an amplifier with remarkably low distortion levels. Despite its many advantages, the circuit has never found a commercial realization—perhaps because it measures too well to "sound good" to the indoctrinated open-loop listener.

While complementary symmetry has been quite common in high-end electronics said to "sound good," the latest trend is to single-ended designs. Even complete power amplifiers have been designed with single-ended output stages. These designs are receiving favorable reviews now that "single-ended" has become the voguish buzzword among open-loop listeners.

Another common trend in high-end design is the use of class A output stages. Audiophile folklore has always held the view that class A amplifiers "sound better." This can be carried as far as biasing the output stage of a power amplifier into class A. Clearly, crossover distortion is possible in class A/B stages, but as Sandstrøm has pointed out this can be minimized [25] in good designs, which show very low amounts of distortion at all signal levels. Cherry [26] has also identified a potential source of distortion in class A/B amplifiers, but as his paper explains this source can be easily dealt with.

Some power amplifiers are stated to be class A by manufacturers so that audiophiles will think they "sound good," even when they are in reality class A/B amplifiers with high quiescent currents levels. Other amplifiers use dynamic biasing circuits that keep the output device which is not driving the load biased to a small constant quiescent current. This technically fits the definition of a class A amplifier, but since the output devices still experience wide variations in current flow, the problems identified by Sandstrøm still apply. This dynamic biasing approach has been adopted in nonaudio applications that must run at low power-supply voltages [27]. In these applications an emitter-follower based output stage will not work, and the dynamic biasing circuit has been shown to be a good way to bias a common-emitter output stage. The designers of these dynamically biased output stages still refer to them as class A/B and never class A. Audiophiles are not aware of such distinctions and in open-loop listening tests they find amplifiers labeled class A to have excellent sonic qualities in comparison with class A/B amplifiers—even if the so-called class A amplifier passes into class A/B above a certain signal level or if the amplifier labeled class A uses dynamic biasing.

If a high-end designer is using an op-amp, he may put a load resistor from the output to the negative supply rail to cause a large dc current to flow. This dc current forces the *npn* output transistor on for the full swing of the output, yielding class A operation [28]. Again the operative word is *may* (in the sentence before the last), since this practice is used in some designs but missing in many others. One design I have examined got this backwards by placing the resistor to the positive supply rail, forcing the slow lateral *pnp* transistor on instead. Despite this the amplifier has received good reviews. (The old football-team adage "what the ref don't see don't bother the ref" appears to apply to high-end audio reviewers as well.)

There are other design trends in output stages which can be seen in high-end designs. In low-level signal stages, current-limiting circuits are not often used. Resistors in series with the output provide the current limiting. In power amplifiers, current limiters may still not be used, with designers relying on rail fuses to protect the amplifier. If current limiting is used, it will not be the simple one-transistor foldback design. It can be shown, using some novel test procedures, that an improperly designed current-protection circuit can activate prematurely into real loudspeaker loads [29], [30]; thus, relatively sophisticated protection circuits are required. The effect of carefully designed current-protection circuits can be measured by checking to see if the voltage output is reduced when driving reactive loads. Often these tests must be done on a dynamic basis because the amplifier may have inadequate heat sinks for steady-state testing [6]. Table 2 shows, among many other things, the variety of protection schemes used in power amplifiers. It should be noted that many amplifiers said to "sound good" in open-loop listening test have protection circuitry which operates poorly.

Some designers state that their products "sound good" because of extensive use of field effect transistors. Designers' typical explanation for this is that FETs perform more like tubes. More scientific explanation for using these devices includes the fact that FETs do not require a dc current at the gate, that FETs increase the input stage's linear dynamic range (a requirement for some designers, as explained above), and that FETs are more robust into overload and short-circuit conditions (an important attribute if the amplifier has limited protection circuits, as explained above). Some IC manufactures often encourage the use of BiFET op-amps for better sound quality, using the explanation that they have a distortion characteristic more like tubes [31]. This begs the question that if the distortion numbers are very low, why does it matter what the characteristics of the distortion are?

The well-known downside of FET devices is that they have lower g_m for a given bias current. Thus a source follower will have higher distortion levels than a bipolar device if the bias currents are constant. *1/f* noise can also be a problem in an input stage. Crossover distortion can be a significant problem with MOSFETs, not easily dealt with by means of global negative feedback [21]. A complete discussion of the tradeoffs between FETs and bipolar devices in the output stages of power amps is beyond the scope of this paper, as these represent real engineering tradeoffs not based on anecdotal sonic considerations. See [6] for more details.

Radically overdesigned power supplies are sometimes used in high-end audio equipment, especially for low-level electronics. It is not uncommon to see power transformers and rectifiers much larger than required to drive the power supplies. Perhaps the added size and weight of the components cause open-loop listeners to think it "sounds better."

Multiple stages of regulation are often used [32]. Sometimes this is carried to the point where each active amplifier has its own local regulator. Discrete regulators are often used instead of cheaper monolithic devices. Again, no consistent design practice is observable, since inexpensive monolithic regulators often drive complex, discrete, active electronics in some designs. On the other hand, very complex discrete regulators often are used to drive low-cost op-amps. This approach is common in high-end Japanese designs. One interesting feature in the high-end Japanese designs is the use of a complete push-pull output stage in the regulator. It is unclear why a positive regulator should ever be required to sink current. One assumes that it has a role in reducing transient noise signals on the supply line.

Sometimes high-end discrete regulator stages use no global feedback. This approach, perhaps an attempt to mimic the low-feedback design of the active stages, results in a less capable regulator with much higher output impedance. Some high-end designs will be dual-mono right to the power cord. Other designers will use supply rails for both channels derived from a single voltage regulator.

Perhaps the greatest deviation in power-supply design occurs in power amplifiers. Some amplifiers will have complete regulation of all active elements, including the output stage. Holding the output-stage voltage rails constant is counterproductive if a purely engineering analysis is applied. A much more logical approach would be to dynamically vary the power supply voltage to the output stage so that the V_{BC} (or V_{DS}) of the device could be held constant. This would linearize the output stage and allow for the use of faster devices with lower V_{BCO} (or $V_{DS(\max)}$). Despite the technical quicksand that power amplifiers with regulated output stages stand on, the open-loop sonic descriptions have often been very favorable.

Most power amplifiers do not have regulated output-stage supply rails, but some have regulated rails for the voltage-gain stages. This approach improves distortion performance, reduces crosstalk if the power-amp channels share the same power supply, and increases immunity to power-line noise. The downside of regulating the voltage-gain stages is a significant increase in complexity because the regulated voltage must be higher than the unregulated voltage applied to the output stage if the available output-voltage swing is not to be limited by the regulated supply. This requires additional transformer windings for the regulated power supply. One interesting way to have voltage rails at a lower potential than the output-stage rails, without losing headroom, is to design the output stage with a small amount of gain. This approach, which is used by Bryston, trades power-supply complexity for output-stage complexity. Again, so-called "good-sounding" power amplifiers use no consistent power-supply design, as can be seen in Table 2.

Designers often make a big deal about the effect passive components can have on sound quality. Teflon PC boards, silver wire, and very expensive bulk-metal resistors are just a few of the strange things that may be found in high-end audio designs, but most of the emphasis appears to be on capacitors.

Fig. 4 shows the locations where capacitors can be used in an active amplifier. C_1 rejects any dc voltage present at the input of the circuit. In addition, it ensures that any dc bias current required by the stage will not be sourced from the circuit preceding it. This prevents clicks and pops during the operation of passive potentiometers and switches, such as could result if dc current flowed in these components. R_1 provides any bias current required by the stage and sets the dc reference voltage for the stage. The use of a FET at the circuit's input will also eliminate any bias currents. C_1 is thus often eliminated if a FET input is used. C_3 prevents any dc offset present at the output of the stage from passing to the next stage.

While C_3 blocks any dc that could be passed to the next stage, it does not prevent headroom loss in high-gain circuits due to the presence of a large dc offset at the active circuit's output. Placing C_2 in the feedback loop solves this problem, since it reduces the dc gain of the stage to unity and the dc offset is not amplified. This approach often reduces the offset to less than 10 mV, and it thus becomes possible to eliminate C_3.

Manufacturers will often claim a stage is direct-coupled if C_1 and C_3 are not present, even if C_2 is. They claim the circuit "sounds better" because the capacitors have been eliminated from the signal path, and these amplifiers do very well in open-loop listening tests. This is not logical because, if any signal degradation results from the presence of a capacitor, it can be caused by C_2 as easily as by the other capacitors. C_1 and C_3 are outside the closed-loop amplifier, so any nonlinearity directly affects the output, but any nonlinearity in C_2 will also directly affect the output, since C_2 is in the feedback loop, not the forward loop, and thus it is not linearized by the feedback amplifier.

In almost all designs, R_2 and R_3 will be significantly smaller than R_1 and R_4. This results in a requirement that C_2 must be significantly larger than C_1 or C_3 if the poles that result from each of the capacitors are to be in the same place. Often designers will not place electrolytic capacitors in the signal path [33]. As a result, parts cost for capacitors can become very high. Sometimes designers will use film capacitors for C_1 and C_3 because these capacitances are not so large as to make the cost and size of a film capacitor unacceptable, but for C_2 an electrolytic will be used. This approach makes no engineering sense, since C_2 is physically larger than C_1 or C_3 and it carries a larger displacement current; thus any distortion mechanism due to capacitors will dominantly come from C_2.

Some designers eliminate coupling capacitors completely. The dc offset is reduced with trim pots or active circuitry in the feedback path [34]. The active circuitry is often called a dc servo. The dc servo is widely used in nonaudio electronics, where establishing the required dc level of the circuit is difficult or where a complete monolithic implementation of a system with no external components is required. There are no purely engineering reasons to use this complex scheme in the audio applications under discussion here.

One example of such a circuit is shown in Fig. 5. The added operational amplifier, C_4, C_5, R_5, and R_6 form a noninverting integrator. R_7 sums the servo into the active stage. Note that the presence of R_7 will change the closed-loop gain of the active stage, but this change is easily calculated. In the audio band, the signal at the output of the integrator is significantly attenuated and the integrator is effectively out of the circuit, although some audio designers who do not use dc servos argue that the presence of the integrator somehow still affects the sound. At dc, the integrator forces the gain stage's output to become equal to the integrator's dc input offset. Very low-offset op-amps that are not designed for audio applications can be used for the integrator op-amp because it is not in the signal path at audio frequencies; however, for some reason known only to the designer, the integrator stage is sometimes formed not with a precision op-amp but with discrete components.

The disadvantage of the dc servo is added complexity; furthermore, a failure of the servo or the amplifier can be catastrophic, with the possibility that the full power-supply rail will be present at the circuit's output. In preamps that use capacitors for dc blocking at the output such a situation is not possible. Ideally, a dc detection circuit should be placed at the output of any unit using a dc servo, but this is rarely done. In the event of a component failure, the dc detector would mute the output.

3 CONCLUSION

The examination of the topologies of audio equipment said to "sound good" has shown little commonality in the designs. Some designs use no feedback, others a small amount, and yet others a large amount at low frequencies. Some designers include the output stage in the feedback loop; others do not. Some designers will use no capacitors in the signal path; others will use only expensive film caps, while still others use less expensive electrolytics. Some designers will use complex power-supply regulators, while others will use no regulation at all in power amplifiers. Some designers will work mostly with FETs; others use only bipolar devices. Some designers use fully complementary circuits, while others use only single-ended circuits. Some designers use ICs in the signal path or for voltage regulation; others will use only discrete designs. Some designers may even mix design styles within a given unit.

The random nature of the designs strongly suggests that no "X factor" parameter is being optimized. Instead, we can assume that the designer, using open-loop listening tests, has convinced himself that the changes he is making to the circuitry are affecting the sound. In this process—design, listen open-loop, design—the circuit designer has no checks and balances to guide him in his work. Open-loop listening is to a very great extent subject to the biases of the listener, and a designer wanting to prove that his new idea is better-sounding is clearly biased. Closed-loop listening tests would show if a change were truly happening when a circuit change was implemented, but designers are unhappy when a new circuit idea is shown by such tests to be of no consequence. They thus try to dismiss the controlled test results, instead of facing the reality that electronics exhibiting proper measurements are sonically transparent.

This can often work in reverse. A designer might not use in his circuit an expensive component that would result in a measurable change in the device's performance because he has convinced himself through open-loop listening tests that the better component produces no sonic change. That is probably the reason why a lot of very expensive high-end equipment uses inexpensive D/A converters or digital interpolation filters. Considerably less expensive mainstream components, often said to sound less good in open-loop listening tests, use much better-performing parts.

The results of the study of the circuits in this paper confirm the results of the controlled double-blind test which have shown that no sonic differences exist in audio electronics that measure well. When double-blind listening test are performed, random answers occur to the question "Which component sounds better?". When circuits that are claimed to "sound better" are analyzed, random design techniques are noted. Both analysis techniques, approaching the subject from opposite ends, converge to the same conclusion: audio electronics that measure properly will sound acoustically transparent. No "X factor" exists. Designers are wasting their time developing audio equipment using the "design, listen, design" approach because they are not using controlled techniques in the "listen" part of the process. If controlled techniques were used, the designers would discover that audio design is no different form other electronic design. It is done with a set of specifications, with paper and pencil, with computer analysis programs, and with laboratory measurements.

It should also be noted that audio—high-end audio in particular—appears to be the only technological discipline suffering from the peculiar attitudinal syndrome discussed above. You will not find automotive engineers, for example, claiming that one brand of spark plug (or ignition wiring or distributor) with exactly the same specifications and measured performance as another "feels better" when driving, and certainly not that it makes the car go faster!

We are faced with so many real problems in audio design. It is time for the designers of audio electronics to recognize that they are not accomplishing anything and move on to the solution of those real problems.

4 REFERENCES

[1] Clark, D. "High-Resolution Subjective Testing Using a Double-Blind Comparator." *Journal of the Audio Engineering Society* 30 (May 1982): 330–38.

[2] Rich, D. A. "Scientific Audio Design versus Consumer Belief Systems: An Introduction." Proceedings of the *DSPx '95* Technical Program, San Jose, CA (15–18 May 1995).

[3] Harley, R. "The Rôle of Critical Listening in Evaluating Audio Equipment Quality." 91st Convention of the AES, New York, NY (4–8 October 1991): Preprint 3176, p. 4, 1st paragraph and footnote 1.

[4] Rich, D. A. "The Present State of CD Player Technology: Who Is Doing It Right?" *The Audio Critic* 15 (Spring through Winter 1990–91): 8–41.

[5] Rich, D. A. "Reasonably Priced Preamplifiers for the Reasonable Audiophile." *The Audio Critic* 18 (Spring/Summer 1992): 12–38.

[6] Rich, D. A. "A New Look at Medium- and High-Priced Power Amplifiers." *The Audio Critic* 20 (Late Summer 1993): 14–37.

[7] Gray, P. and R. Meyer. *Analysis and Design of Analog Integrated Circuits.* 2nd ed. John Wiley & Sons, 1984.

[8] Wurcer, S. "An Operational Amplifier Architecture with a Single Gain Stage and Distortion Cancellation." 92nd Convention of the AES, Vienna, Austria (24–27 March 1992): Preprint 3231.

[9] Borbely, E. "A Multi-Tone Intermodulation Meter." (Parts I, II, and III; Part III titled "Measuring Nonlinear Distortions.") *The Audio Amateur* 20.2 (April 1989): 7–15; 20.3 (August 1989): 13–29; 21.3 (August 1990): 8–25.

[10] Eschauzier, G. H. and J. Huijsing. *Frequency Compensation Techniques for Low-Power Operational Amplifiers.* Kluwer Academic Publishers, 1995.

[11] Jensen, D. "JE-990 Discrete Operational Amplifier." *Journal of the Audio Engineering Society* 28 (January/February 1980): 26–34.

[12] Rich, D. A. Reviews of the Harman Kardon HD7725 and Krell Studio. *The Audio Critic* 21 (Spring 1994): 46–50.

[13] Otala, M. "Transient Distortion in Transistorized Audio Power Amplifiers." *IEEE Transactions on Audio and Electroacoustics* AU-18 (September 1970): 234–39.

[14] Otala, M. "Feedback-Generated Phase Nonlinearity in Audio Amplifiers." 64th Convention of the AES, London, England (25–28

February 1980): Preprint 1576.

[15] Cherry, E. M. "Feedback, Sensitivity, and Stability of Audio Power Amplifiers." *Journal of the Audio Engineering Society* 30 (May 1982): 282–94; 31 (November 1983): 854–57.

[16] Cordell, R. R. "Another View of TIM." (Parts I and II.) *Audio* 64.2 (February 1980): 38–49; 64.3 (March 1980): 39–42.

[17] Cordell, R. R. "Phase Intermodulation Distortion Instrumentation and Measurements." *Journal of the Audio Engineering Society* 31 (March 1983): 114–24.

[18] Jung, W., M. Stephens, and C. Todd. "An Overview of SID and TIM." (Parts I, II, and III.) *Audio* 63.6–8 (June, July, and August 1979).

[19] Dalzell, L. "On Amplifiers and Feedback." *The Audio Amateur* 26.1 (January 1995): 20–25.

[20] DiStefano, J., A. Stubberud, and I. Williams. *Feedback and Control Systems.* 2nd ed. Schaums's Outline Series, McGraw-Hill, 1990: 221.

[21] Cordell, R. R. "A MOSFET Power Amplifier with Error Correction." *Journal of the Audio Engineering Society* 32 (January/February 1984): 2–17

[22] Jung, W. G. *Audio IC Op-Amp Applications.* 3rd ed. Howard W. Sams & Co., 1986.

[23] Leach, W. M. "An Amplifier Input Stage Design Criterion for the Suppression of Dynamic Distortions." *Journal of the Audio Engineering Society* 29 (April 1981): 249–51.

[24] Cherry, E. M. and K. P. Dabke. "Transient Intermodulation Distortion—Part 2: Soft Nonlinearity." *Journal of the Audio Engineering Society* 34 (January/February 1986): 19–35.

[25] Sandstrøm, T. "Distortion in Class AB Power Amplifiers." *Journal of the Audio Engineering Society* 31 (November 1983): 858–61.

[26] Cherry, E. M. "A New Distortion Mechanism in Class B Amplifiers." *Journal of the Audio Engineering Society* 29 (May 1981): 327–28.

[27] Sakurai, S. and M. Ismail. *Low-Voltage Operational Amplifiers.* Kluwer Academic Publishers, 1995.

[28] Jung, W. G. "The Magnavox 16-Bit Series: Making Good Players Better." *Audio* 71.6 (June 1987): 75–80.

[29] Otala, M. and P. Huttunen. "Peak Current Requirement of

Commercial Loudspeaker Systems." *Journal of the Audio Engineering Society* 35 (June 1987): 455–62.

[30] Baxandall, P. J. "A Technique for Displaying the Current and Voltage Output Capability of Amplifiers and Relating This to the Demands of Loudspeakers." *Journal of the Audio Engineering Society* 36 (January/February 1988): 3–17.

[31] *Burr-Brown IC Data Book.* Vol. 33c. OPA2604. Burr-Brown Research Corp., 1992: page 2.221.

[32] Jung, W. G. "Regulators for High-Performance Audio." (Parts 1 and 2.) *The Audio Amateur* 26.1 (January 1995): 8 and 26.2 (April 1995): 8.

[33] Jung, W. G. and R. N. Marsh. "Picking Capacitors." *Audio* 64.2–3 (February and March 1980).

[34] Clark, B. "DC Servo Loop Design for Audio Amplifiers." *The Audio Amateur* 13 (August 1982): 14.

[35] Snyder, W. *Class A Audio Design.* New England Analog, 1990.

[36] Bongiorno, J. W. "Ampzilla III." *The Audio Amateur* 15.4 (September 1984): 7–19.

[37] Borbely, E. "Third-Generation MOSFETs: The Servo 100." (Part I.) *The Audio Amateur* 15.1 (March 1984): 7–18.

[38] Didden, J. M. "New Power Amp Alternatives: A 200 Watt Version." (Parts I and II.) *The Audio Amateur* 14.4 (November 1983): 7–15; 14.5 (January 1984): 30–39.

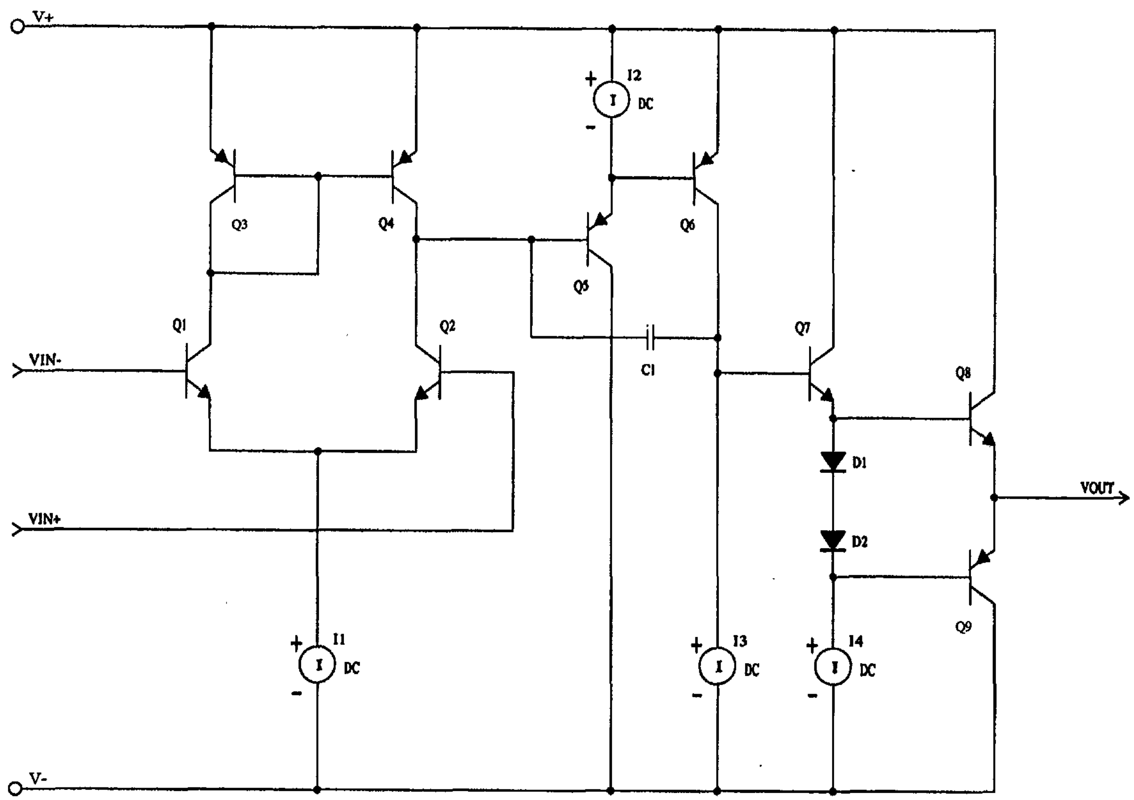

Fig. 1. Basic topology used in audio-band integrated op-amps.

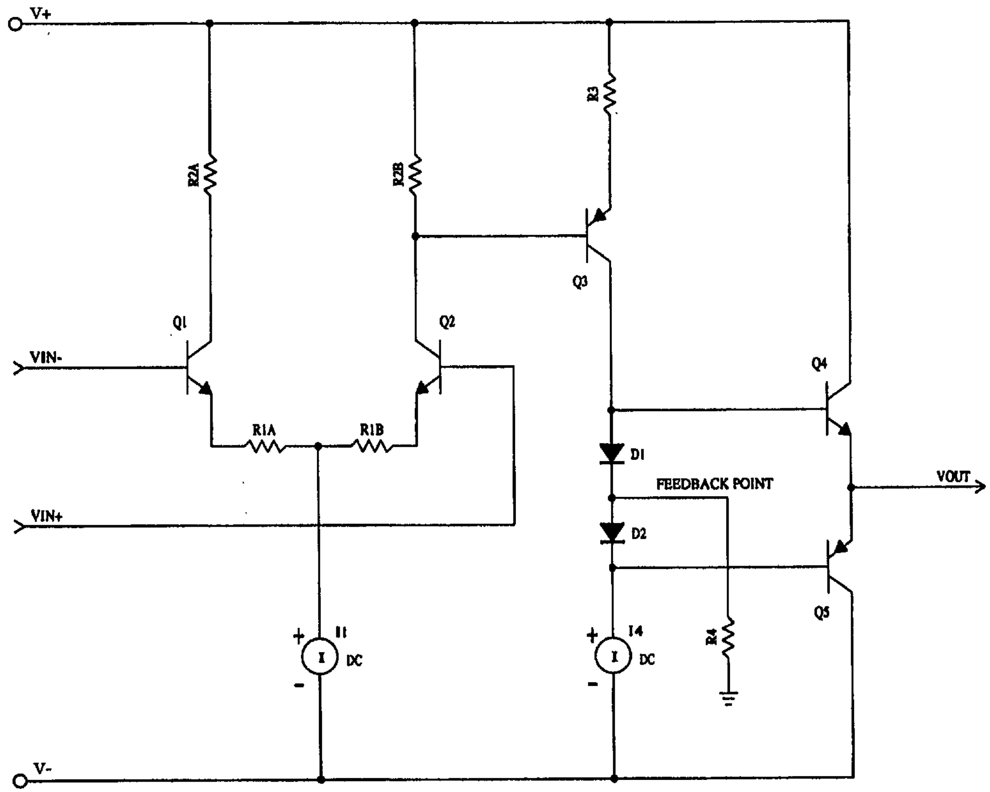

Fig. 2. Typical topology of an amplifier stage in a "high-end" product.

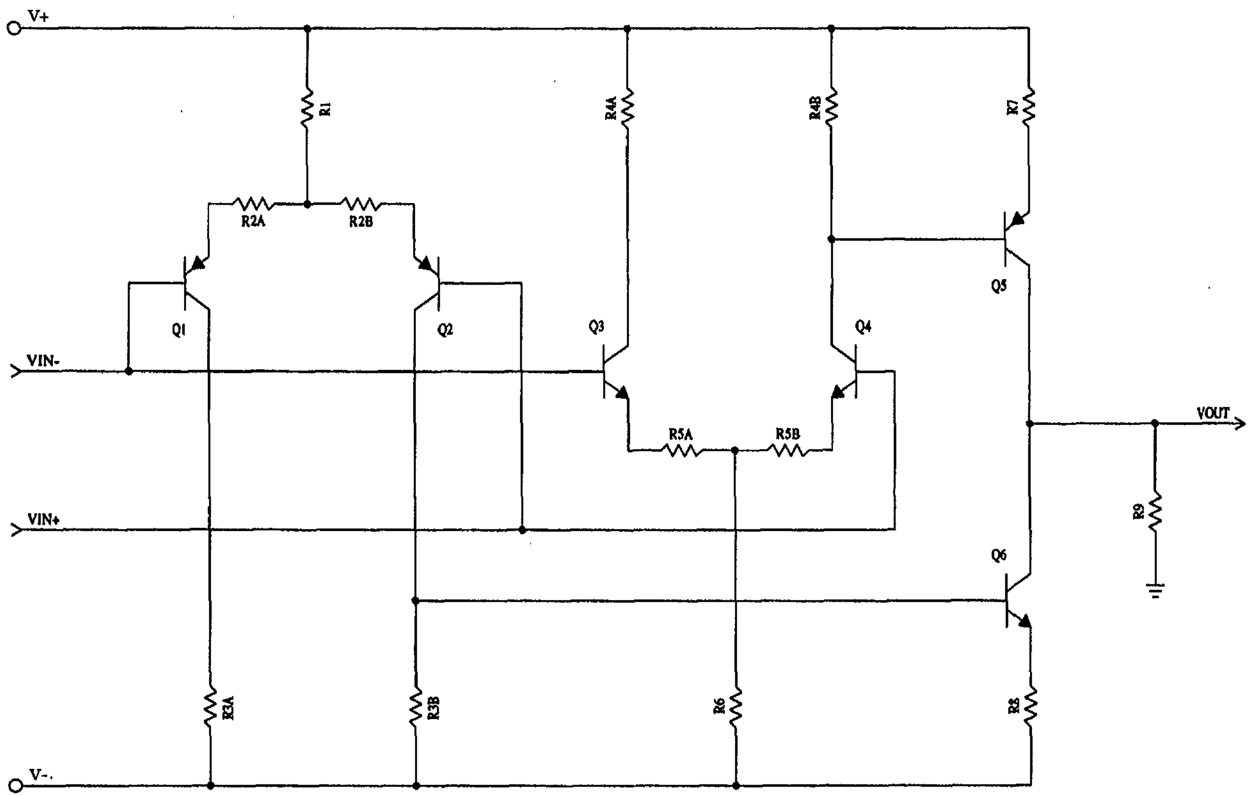

Fig. 3. Amplifier with fully complementary topology from the input stage onward.

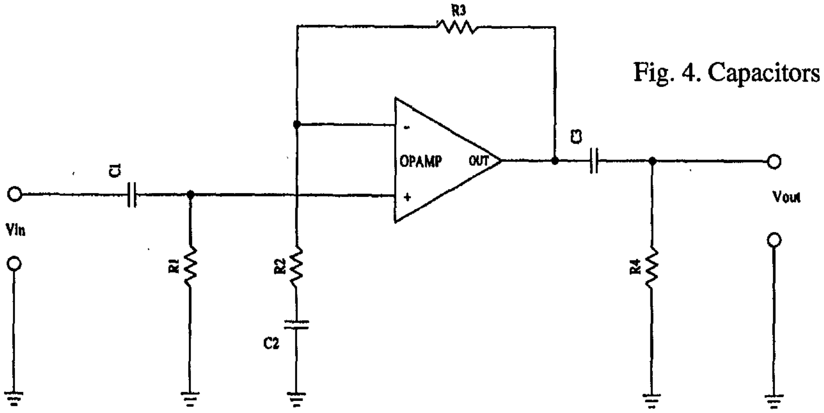

Fig. 4. Capacitors in an active amplifier stage.

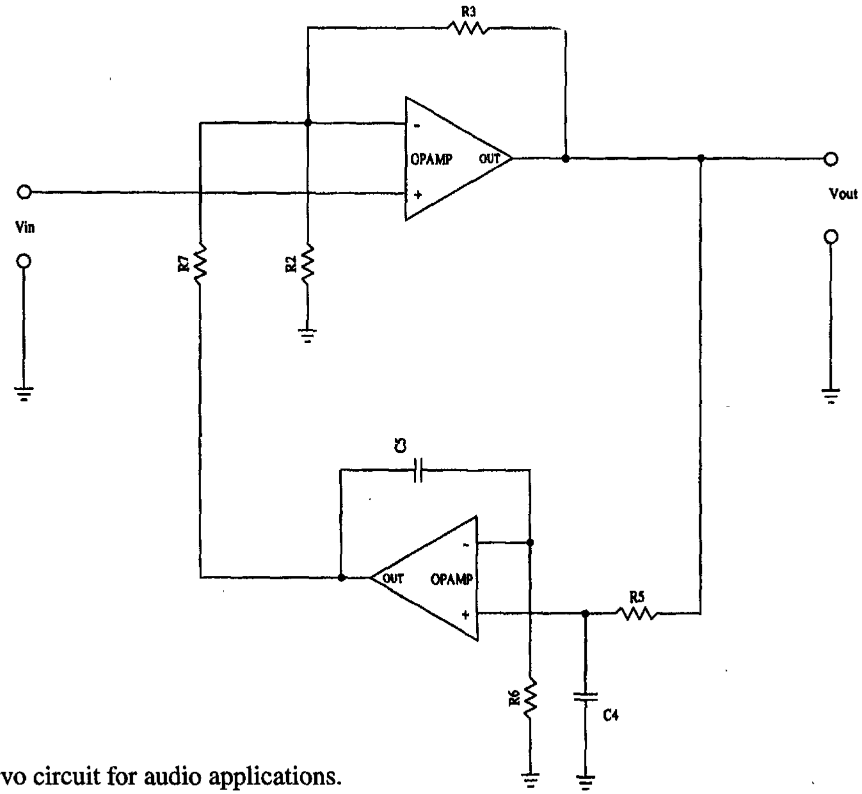

Fig. 5. Typical dc servo circuit for audio applications.

Table 1. Comparison of significant preamplifier topologies. (Model numbers and prices may not be entirely up-to-date but are still representative.)

	LINE AMPLIFIER								
	Differential Pair					*Follower Stage*	*Second Gain Stage*		
MODEL	*Active Element*	*Comple-mentary Symmetry*	*Biased by Current Source*	*Cascode Stage*	*Load*		*Active Element*	*Cascode Stage*	*Load*
Acurus P10 (phono) $395.00	NA	NA	NA	NA	NA	NA	NA	NA	NA
Adcom GFP-565 $799.95	NA (IC)	NA (IC)	NA (IC)	NA (IC)	NA (IC)	NA (IC)	NA (IC)	NA (IC)	NA (IC)
Aragon 18k (line) $995.00	Bipolar	Yes	Yes	No	Resistor	No	Bipolar	No	Active
B&K PRO10-MC $698.00	JFET	No	Yes	Yes	Resistor	No	Bipolar	No	Active
Borbely Audio[1] (partial kit form only)	JFET[2]	Yes	NA	Yes	Resistor	No	Bipolar	Yes	Active
Boulder "Ultimate" $5299.00	[JE-990-based]	[JE-990-based]	[JE-990-based]	[JE-990-based]	[JE-990-based]	[JE-990-based]	[JE-990-based]	[JE-990-based]	[JE-990-based]
Bryston BP20 $1395.00	Bipolar	Yes	No	No	Resistor	No	Bipolar	No	Active
Citation 21 $629.00	Bipolar	Yes	No	No	Resistor	Yes	Bipolar	No	Resistor
Coda 01 $2500.00	JFET	No	Yes	No	Resistor	No	MOSFET	Yes	Resistor
JE-990 (basic discrete op-amp)	Bipolar	No	Yes	No	Active	Yes	Bipolar	No	Active
Krell KRC-2 $3700.00	Bipolar[4]	Yes	Yes	No	Resistor	No	Bipolar	No	Resistor
New England Audio (partial kit form [35])	Bipolar	Yes	Yes	No	Resistor	No	Bipolar	No	Active
PS Audio 5.5 $1195.00	JFET	No	Yes	No	Resistor	No	Bipolar	No	Resistor
Rotel RHA-10 $1799.90	Bipolar	Yes	Yes	No	Resistor	No	Bipolar	No	Active
Jeff Rowland "Coherence One" $4600.00	JFET	Yes	Yes	Yes (Folded)	Resistor	No	NA	NA	NA
Sumo Athena II $828.00	Bipolar	Yes	Yes	No	Resistor	No	Bipolar	No	Resistor
Tandberg TCA-3018A $2299.00	Bipolar[6]	Yes	NA	No	Active	No	Bipolar	Yes	Resistor
Threshold FET ten/e $5700.00 (hl + pc)	JFET	No	Yes	No	Active	No	MOSFET	Yes	Active

[1]Dolan PM1 has similar phono amplifier
[2]Common-source input stage
[3]Output stage not in feedback loop
[4]Open-loop JFET/Bipolar composite follower is before this stage
[5]Bipolar predriver stage
[6]Input connected to emitter follower which drives common-emitter input stage
[7]No global NFB used; Hawksford distortion-correction circuit used at output stage

----------> Push-Pull	Output Stage	<--------Coupling Capacitors-------> C_1	C_2	C_3	Total No. of Transistors	Power Supply Voltage	<--------------PHONO AMPLIFIER-------------> 50 Hz Pole	500 Hz Zero	2122 Hz Pole	Notes
NA	NA	NA	NA	NA	NA	NA	See A	See A	See A	See E
NA (IC)	NA (IC)	No	No	No/Film	NA (IC)	±18	See A	See A	See A	See C/D/F
Yes	MOSFET	No	No	Film	8	±20	NA	NA	NA	
Yes	Bipolar	DC servo	DC servo	DC servo	13 + 1 IC	±30	See A	See A	See A	See E
Yes	MOSFET	DC servo	DC servo	DC servo	10 + 3 ICs	±24	Active Stage 2	Active Stage 2	Passive Aft. stage 1	Stage 1 for gain only
[JE-990-based]	[JE-990-based]	DC servo	DC servo	DC servo	?	±25	See B	See B	See B	See C/D Stage 2 for gain only
Yes	Bipolar	Film	Electro-lytic	Film	10	±24	Active Stage 1	Active Stage 1	Active Stage 2	Stage 2 is inverting
Yes	None	Electro-lytic	Electro-lytic	Electro-lytic	8	±23	See B	See B	See B	See C/D/F Stage 2 is unity-gain
Yes	Bipolar[3]	No	No	No	12	±30	Active Stage 2	Active Stage 2	Active Stage 1	See C/D
No	Bipolar	NA	NA	NA	9	Varies	NA	NA	NA	
Yes	Bipolar[5]	DC servo	DC servo	DC servo	13	±20	NA	NA	NA	
Yes	MOSFET w/predr.[5]	DC servo	DC servo	DC servo	16	±25	Passive Aft. stage 1	Passive Aft. stage 1	Passive Aft. stage 2	3-stage design
No	MOSFET	Film	No	'Lytic + film byp.	10	±30	Passive Aft. stage 1	Passive Aft. stage 1	Passive Aft. stage 1	Active stages for gain only
Yes	Bipolar	'Lytic	No	'Lytic	10	±24	NA	NA	NA	
NA	Bipolar	DC servo	DC servo	DC servo	?	±20	Passive Aft. stage 2	Passive Aft. stage 2	G_m cell, no global FB Stage 1	3-stage design (stg. 2 buffers 1)
Yes	Bipolar	Film	Electro-lytic	NP 'lytic +film byp.	10	±35	See A	See A	See A	See C/D/F
Yes	Bipolar[5, 7]	No	No	Film	16	±22	Active: G_m cell, no global FB Stage 1	Active: G_m cell, no global FB Stage 1	RC ntwk in G_m cell Stage 1	Stage 2 buffers stage 1 & adds gain
No	Bipolar	DC servo	DC servo	DC servo	11 + 1 IC	±18	See A	See A	See A	See C/D

A. All equalization is performed in one noninverting active stage.
B. All equalization is performed in the first active stage.
C. An extra zero is added to reduce the reactive load of the network on the stage.
D. The added zero is canceled by a passive network at some point in the circuit.
E. A zero at approximately 130 kHz is not canceled.
F. The output impedance of the phono stage is high.

Table 2. Comparison of significant power amplifier topologies. (Model numbers and prices may not be entirely up-to-date but are still representative.)

MODEL	*<------Differential Pair*				*------>*	*Buffer Stage or Compound 2nd Stage*	*<------*
	Active Element	*Comple-mentary Symmetry*	*Biased by Current Source*	*Cascode Stage*	*Load*		*Active Element*
J. W. Bongiorno "Ampzilla III" [36] (no longer offered)	Bipolar	Yes	Yes	No	Resistor	No	Bipolar
Aragon 4004 MKII $1850.00	Bipolar	Yes	Yes	Yes	Resistor	Yes	Bipolar
B&K Sonata M-200 $998.00 each (mono)	Bipolar	No	Yes	Yes	Active	Yes	Bipolar
Borbely Audio [37] (kit, made in Germany)	Bipolar	Yes	Yes	No	Resistor	Yes	Bipolar
Bryston 4B NRB $2195.00	Bipolar	Yes	No	No	Resistor	No	Bipolar
Citation 22 $1149.00	Bipolar	Yes	Yes	Yes	Resistor	Yes	Bipolar
Cordell (prototype) [21]	Bipolar	No	Yes	Yes	Active	Yes	Bipolar
Didden (prototype) [38]	Bipolar	No	Yes	Yes	Resistor	No	Bipolar
Hafler Series 9500 Transnova $1800.00	JFET	Yes	Self-biased	Yes	Resistor	No	Bipolar
McIntosh MC500 $6,500.00	Bipolar	Yes	Yes	No	Resistor	No	Bipolar
New England Analog (plans only [35])	Bipolar	Yes	Yes	Yes	Resistor	Yes	Bipolar
Parasound HCA-2200II $1695.00	JFET	Yes	Self-biased	Yes	Resistor	No	Bipolar
PS Audio PS 100 Delta $1195.00	JFET	No	Yes	Yes (dy-namic bias)	Resistor	Yes	Bipolar
R.E. Designs LNPA 150 $2700.00 the pair (mono)	Bipolar	Yes	No	No	Resistor	No	Bipolar
Rotel RHB-10 $2699.90	Bipolar	Yes	Yes	No	Resistor	Yes	Bipolar
Sansui Vintage AU-X911DG $1250.00	Bipolar	Yes[6]	Yes	No	Resistor	No	Bipolar

[1]Nonlinear load implements soft clipping.
[2]The collector of the noninverting side of the differential pair is terminated into the emitter of the second gain stage. This is a folded-cascode-like topology.
[3]Part of a closed-loop feedback amplifier built around the output section.
[4]The Hawksford distortion correction circuit is used at the output stage.
[5]Dynamic output-bias-current set keeps quiescent current constant under different load conditions.
[6]Diamond Differential (X-cell) configuration biased with floating voltage sources.

-----*Second Gain Stage*------------------->			*Regulated Supplies on V Gain Stages*	*Output Predriver Stage(s) and Type*	*Number and Type of Output Devices*	*<--Coupling Capacitors-->*		*Protection*
Cascode Stage	*Load*	*Push-Pull*				C_1	C_2	
No	Active	Yes	No	2 Bipolar	3 per side (balanced) Bipolar	Nonpolar electrolytic	No (DC servo)	A
No	Active	Yes	No	1 Bipolar	4 Bipolar	No	NP 'lytic + film bypass	A, B, E
No	Active	No	No	None	1 MOSFET[1]	No	No (DC servo)	?
No[2]	Active	Yes	No	1 MOSFET	2 MOSFET	Film	No (DC servo)	A
No	Active	Yes	Yes	3 Bipolar[3]	4 Bipolar	Film	Electrolytic	B, D, I
No	Resistor	Yes	No	2 Bipolar	4 Bipolar	Electrolytic	No	B, F, I
Yes	Active	Yes	Yes	3 Bipolar	1 MOSFET[4]	Film	No	?
Yes	Resistor[1]	No	Yes (dynamic cascode on output stage)	2 Bipolar	4 Bipolar[5]	Film	No (DC servo)	H, I
Yes	Output stage loads 2nd stage	Yes	Yes	None	4 MOSFET	No	Electrolytic + film byp.	A
Yes	Active	Yes	No	2 Bipolar	10 Bipolar	Electrolytic	Electrolytic	B, D, E, F
No	Resistor	Yes	Yes	1 Bipolar	4 Bipolar	No	No (DC servo)	D, I
No	Active	Yes	Yes	1 MOSFET	6 Bipolar	No	No (DC servo)	A, B, D, E
No	Resistor	Yes	No	1 Bipolar	2 Bipolar	No	No	B, G, I
No[2]	Active	Yes	Yes (output stage also)	1 Bipolar	2 Bipolar	No	Electrolytic + film byp.	A
No	Active	Yes	No	2 Bipolar	5 Bipolar	Electrolytic	Electrolytic	A, B, D, E
No	Active	Yes	No	2 Bipolar	1 per side (balanced) Bipolar	No	No	E, F

Protection
A—DC rail fuses
B—Thermal sensing
C—Second-stage current limiting
D—Output-stage current limiting
E—DC input sensing
F—Excess input sensing
G—Output fuse in feedback loop
H—SOA monitor circuit (analog)
I—Output diodes to block inductive kick