Feedforward Error Correction in Power Amplifiers[*]

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in particular*

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The engineering community is presently putting much effort into designing low-distortion amplifiers with techniques more sophisticated than ordinary feedback. To guide this effort a detailed exposition of feedforward error nulling techniques is presented, and a recent commercial design is analyzed with an illustration of significant modifications. The paper thus consists of: 1) a brief history of error feedforward, why it was eclipsed by feedback, and why the time is now ripe to exploit its possibilities for total error nulling; 2) an analysis of Black's feedforward configuration and how it relates to more recent circuit concepts such as Macdonald's active error feedback, Sandman's error takeoff, Walker's "current dumping," and several new topologies; 3) an illustration of the only commercially available error feedforward circuit, the Quad "current dumping" amplifier; 4) significant modifications to the latter scheme using practical amplifiers, and generalizations of the bridge system incorporated in this concept; 5) an incorporation of error correction into class-D switching amplifiers with resulting relaxed design criteria.

0 INTRODUCTION

The year 1977 marked the fiftieth anniversary of the invention of negative feedback, as we know it today, by Harold S. Black in 1927. So contrary to intuition were the claims made in his patent application, and so far-reaching in their implications, that it took nine years before the patent was finally granted in 1937 [1]. It is enlightening to read the inventor's own account [2] for the historical perspective in which it places the discovery. This is something that is very difficult to appreciate nowadays, since negative feedback as an error-reduction or control mechanism is now all-pervasive and taken for granted in many areas of application. Even the more esoteric aspects of negative feedback systems, such as their stability criteria, are now well understood as a result of the work of Nyquist and Bode, among others. In fact, error *feedforward* as a distortion reduction process distinct from negative *feedback* is now virtually unknown, and it is sobering to realize that its invention in 1923 (and patenting in 1928 [3]), also by Harold Black, predates that of negative feedback by four years. What is feedforward, how does it differ from feedback, why has it been almost totally neglected, and can its use benefit the

audio industry in particular? These are some of the questions which we shall attempt to answer in the sequel. It is our belief, and we trust that we shall be able to substantiate it below, that error feedforward can bestow considerable advantages, and that the time is now ripe to begin to reap some of these benefits. To our knowledge there is at present only one feedforward audio amplifier on the market, namely, the Quad 405 "current dumping" power amplifier [4]–[6], and we shall therefore use its design principle to illustrate the latter portion of this discussion. We also show ways in which the principle can be extended to include even a feedforward pulse-width modulation (class-D) power amplifier within its ambit.

1 ERROR FEEDFORWARD VERSUS NEGATIVE FEEDBACK

The negative feedback principle is now so well known that it is sufficient to refer to Fig. 1 for its salient features. By feeding back to the input of amplifier A a fraction β of its output e_0 in antiphase to the input signal e_i, the overall signal gain of the system is reduced by a factor $(1 + G\beta)$ below the open-loop gain G of the amplifier; that is, an excess gain factor of $(1 + G\beta)$ is required in the amplifier. In return for throwing away this gain when closing the negative feedback loop, we obtain a number of powerful

advantages as well as a number of less desirable disadvantages. The advantages include signal gain stabilization at $1/\beta$ for large amounts of excess gain, output impedance reduction by the feedback factor $(1 + G\beta)$, and above all from our present point of view, a reduction in the distortion of the overall amplifier by the same factor over that of the amplifier without feedback. Unfortunately these advantages are available only in limited amounts — limited, that is, not by engineering considerations, but by theoretical considerations which are apparent from the gain formula in Fig. 1. The open-loop gain G of any amplifier is frequency dependent [that is, $G = G(\omega)$] in both magnitude and phase, and so the vanishing of the denominator, that is, $1 + G\beta = 0$, warns of the potential instability with which one is flirting when negative feedback is being used. For stable

it is *in principle* theoretically impossible to totally null the error by means of negative feedback. We should perhaps remark that we shall use the term "error" in the sense of Black [2, p. 58] and Sandman [7], [8] to include not only new signal components introduced by nonlinearities in G, but also hum, noise, and gain, frequency response, and phase errors present in the output signal. For example, as $G \rightarrow \infty$ in Fig. 1, the signal gain stabilizes at the value $1/\beta$. Any departure of the signal gain from its desired value (due, say, to variations in the value of G) could thus be considered as an error to be corrected.

In contradistinction to Fig. 1, the basic principle of error feedforward[1] is exemplified in Fig. 2.[2] Ignoring for the moment the delay lines τ_1 and τ_2, the concept is as originally described by Black [3]. The attenuated output βe_1 of the

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[1] We shall use the term "error feedforward" to differentiate feedforward error correction in Black's sense from "signal feedforward," a technique frequently used in operational amplifier design as a compensation technique to improve closed-loop stability when negative feedback is applied, but having nothing to do with error correction.

[2] Clearly, a number of distinct permutations of amplifier and sum/difference network polarities is possible in practice. Fig. 2 illustrates just one such possibility.

If $\tau_1 = \tau_2 = 0$: $\quad e_o = e_1 + e_2$

$$G_{ff} = \frac{e_o}{e_i} = G_1 + G_2 - G_1 G_2 \beta$$

$$= \frac{1}{\beta} \quad \text{if} \begin{cases} \text{either } G_1\beta = 1 \\ \text{and/or } G_2\beta = 1 \end{cases}$$

Fig. 2. Error feedforward principle, with main amplifier A_1 whose errors are nulled by the auxiliary amplifier A_2 provided $G_2\beta = 1$. The signal gain G_{ff} is then also $1/\beta$.

$$G_{fb} = \frac{e_o}{e_i} = \frac{G}{1 + G\beta} \xrightarrow{G\beta} \frac{1}{\beta}$$
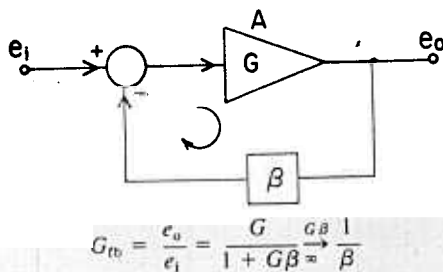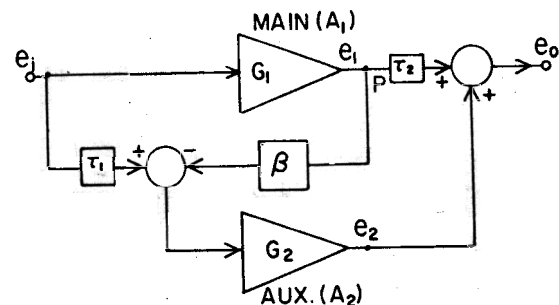
Fig. 1. Standard negative feedback amplifier with forward gain $G(\omega)$ and feedback attenuation β. The signal gain G_{fb} tends to $1/\beta$ as the loop gain $G\beta$ tends to infinity. The distortion of A is reduced by the factor $(1 + G\beta)$.

introduce the potential regeneration present in negative feedback circuits,—there is no stability criterion to be satisfied. Third, no excess gain or bandwidth is required in the main amplifier. The delay line (or phase correction network) τ_1 is designed to correct for the time delay of the main amplifier channel A_1 in order to allow proper synchronous subtraction of its attenuated output signal from the input signal. Similarly, τ_2 corrects for the time delay of the auxiliary amplifier A_2. It is important to note, however, that the absence of time-delay correction does not introduce any form of potential instability into the circuit. It merely leads to less accurate error detection and cancellation than would otherwise be possible. In this sense it is totally different from phase compensation as used in a negative feedback amplifier to enhance stability. Another important distinction between error feedforward and negative feedback is that, since the former acts additively, whereas the latter corrects by a multiplicative process, gain matching accuracy in the various paths in an error feedforward amplifier is of paramount importance. This is apparent from Fig. 2 and the overall system's gain formula. Let us suppose that at any instant there is present at the output P of amplifier A_1, in addition to the linearly amplified portion $G_1 e_1$ of the input signal e_1, a distortion component δ, say, so that $e_1 = G_1 e_1 + \delta$. The error term is attenuated by a factor β, but does not cancel in the input summing network, and so appears at the output of A_2 as $e_2 = -G_2 \beta \delta$. It thus follows that the δ terms will exactly cancel in the output summing network, and e_o will be free of distortion due to A_1, provided that $G_2\beta = 1$, a condition which we shall refer to as "output path balance." If this condition is satisfied, we are guaranteed that, within the accuracy of amplifier A_2, all errors caused by amplifier A_1 will be nulled at the output e_0. Another independent condition relates the gain of amplifier A_1 to the attenuation β. If, for example, we choose $G_1\beta = 1$, that is, $\beta = 1/G_1$, then all primary signal components will be canceled in the input summing network, and only error components will be left to pass through the auxiliary amplifier A_2. We shall call this condition "input path balance." It ensures that the auxiliary channel carries only "pure distortion" and not primary signal, and so minimizes the load on A_2. If, however, we set the input path balance somewhat differently from $G_1\beta = 1$, say to correct the overall amplifier gain to a value different from that of A_1 itself, then A_2 will carry some primary (gain error correction) signal in addition to the distortion components. Its signal-handling capability must now be proportionately greater, as it is now being used, not solely as a distortion-nulling amplifier, but also for gain renormalization. With practical output summing networks for power amplifiers there may be good reasons (as we shall see) for unbalancing the input path. Note, at this stage, that it is possible to achieve both input and output path balances simultaneously by setting $G_1 = G_2 = 1/\beta$. It should be clear that the maintenance of path balance is a crucial requirement of an error feedforward amplifier, as its error nulling accuracy is vitally dependent on proper subtraction and subsequent addition. Its overall gain stability also depends on these same conditions, as we shall see. We should finally note that the output impedance of an error feedforward amplifier is not reduced as in a negative feedback amplifier, while a measure of redundancy is introduced, in that the failure of either A_1 or A_2 does not render the amplifier inoperative, but merely lets it revert to the same performance as a single (unstabilized) amplifier of about the same gain but of reduced output capability. This "fail safe" advantage can be of considerable importance in certain applications.

It is instructive at this juncture to examine in somewhat greater detail the nature of the gain stabilization and distortion reduction furnished by both error feedforward and negative feedback systems. For instance, it may not be obvious that an error feedforward amplifier even affords gain stabilization. As usual, the sensitivity of the overall gain G_t to a parameter x, denoted by $S_x^{G_t}$, is defined by

$$S_x^{G_t} = \frac{x}{G_t} \frac{\partial G_t}{\partial x}.$$

It represents the percentage change in G_t caused by a 1% change in x. In terms of these sensitivities, the actual change ΔG_{ff} in G_{ff} caused by changes ΔG_1 in the gain G_1 of amplifier A_1 and ΔG_2 in the gain G_2 of amplifier A_2 in Fig. 2 is given by

$$\frac{\Delta G_{ff}}{G_{ff}} = S_{G_1}^{G_{ff}} \cdot \frac{\Delta G_1}{G_1} + S_{G_2}^{G_{ff}} \cdot \frac{\Delta G_2}{G_2}$$

to first order. Using the overall gain formula from Fig. 2, we find (see Klaassen et al. [9]) that for the feedforward case,

$$G_{ff} = G_1 + G_2 - G_1 G_2 \beta = \frac{1}{\beta} \qquad \begin{array}{l} \text{if } G_1\beta = 1 \\ \text{and/or } G_2\beta = \end{array}$$

$$S_{G_1} \frac{G_1(1 - G_2\beta)}{G_1(1 - G_2\beta) + G_2} = \begin{cases} G_2\beta, & \text{if } G_1\beta = \\ 0, & \text{if } G_2\beta = \end{cases}$$

$$S_{G_2} \frac{(1 - G_1\beta)G_2}{G_1 + (1 - G_1\beta)G_2} = \begin{cases} 0, & \text{if } G_1\beta = \\ 1 - G_1\beta, & \text{if } G_2\beta = 1 \end{cases}$$

$$S_\beta \frac{G_1 G_2 \beta}{G_1 + G_2 - G_1 G_2 \beta} = \begin{cases} & \text{if } G_1\beta = \\ -1, & \text{if } G_1\beta = G_2\beta = \end{cases}$$

It follows that:
1) Output path balance ($G_2\beta = 1$) makes $S_{G_1}^{G_{ff}} = 0$ (that is, gain insensitive to G_1).
2) Input path balance ($G_1\beta = 1$) makes $S_{G_2}^{G_{ff}} = 0$ (that is, gain insensitive to G_2).
3) Both paths balanced makes the overall gain insensitive to both G_1 and G_2 so that, to first order, $\Delta G_{ff} = 0$. In fact, under these conditions, it can be shown that, to second order

$$\frac{\Delta G_{ff}}{G_{ff}} = -\frac{\Delta G_1}{G_1} \cdot \frac{\Delta G_2}{G_2}$$

This shows that feedforward gain-error reduction is multiplicative. For example, if both paths are balanced to 1%, the overall gain error is only 0.01%.

This should be contrasted with the situation under negative feedback (Fig. 1), where

$$G_{fb} = \frac{G}{1 + G\beta} \quad \overset{G}{\underset{\infty}{\to}} \quad \frac{1}{\beta}$$

$$S_G{}^{G_{fb}} = \frac{1}{+ G\beta} \quad \overset{G\beta}{\underset{\infty}{\to}} \quad 0$$

$$S_\beta{}^{G_{fb}} = - \frac{G\beta}{1 + G\beta} \quad \overset{G\beta}{\underset{\infty}{\to}} \quad$$

We note that as G (or $G\beta$) $\to \infty$, the limiting values of G_{fb} and the sensitivity functions are the same as those for the feedforward case. For *finite* values of G, however, these are only approximately realized, whereas for feedforward they can be exactly achieved for finite G_1, G_2.

Similar considerations apply to error reduction. From Fig. 2 it is seen that the distortion of the main amplifier (δ_{A_1}) is reduced by feedforward according to

$$\frac{\delta_{ff}}{\delta_{A_1}} = 1 - G_2\beta = 0, \quad \text{if } G_2\beta = 1$$

and so, as we have already seen, is completely nulled if the output path is balanced. Any remaining distortion due to A_2 is only second order (that is, distortion of the distortion, and so less than that of either A_1 or A_2 alone) if the input path is also balanced so that A_2 carries only the distortion signal. If the input path is unbalanced so that A_2 also carries a primary signal component, its distortion of the primary signal also appears at the output. For Fig. 1, however,

$$\frac{\delta_{fb}}{\delta_A} = \frac{1}{1 + G\beta} \quad \overset{G\beta}{\underset{\infty}{\to}} \quad 0$$

and so, although distortion is reduced as loop gain is increased, no distortion null is possible for finite loop gain.

From the foregoing it is apparent that error feedforward and negative feedback are two radically different approaches to error correction. They differ fundamentally in their manner of operation, and of the two, only error feedforward holds out the prospect of complete error nulling *in principle*. It is, in fact, a more obvious technique than negative feedback, and actually predates the latter. Why, then, is it almost unheard of? Apart from its greater circuit complexity, the reasons lie in the requirement for precise path gain balance according to the equations in Fig. 2. Ordinary unstabilized amplifiers cannot achieve the required gain constancy. In fact, Black [2] relates how their prototype error feedforward amplifier had to have its filament current and high-tension voltage constantly readjusted in order to keep the tubes' gains correct. With the invention of feedback, which does not require this kind of critical balance, error feedforward faded into the background for the moment. Feedback could achieve most of what was required at that stage. But as operating frequencies and bandwidths increased, it became increasingly difficult to produce a stable negative feedback amplifier to do the job, and error feedforward with its reduced open-loop gain—bandwidth requirement, and lack of Nyquist stability prob-

lems, came back into favor, and is now in use in VHF and microwave amplifiers [10]–[13] where feedback technology is inadequate to the task. In [13] Seidel states: "In relative bandwidth, absolute frequency, and the degree of correction attainable, feedforward has markedly exceeded the capabilities of feedback." The secret of success lies in making each of the two amplifiers A_1 and A_2 in Fig. 2 itself a negative feedback-stabilized amplifier, so that its gain will be tightly controlled, and hence eliminate the major problem faced by the feedforward principle. Lest it be thought that by making A_1 and A_2 into feedback amplifiers we are reintroducing the very stability problems which had been insurmountable barriers before, it should be noted that this is not the case, since fairly modest negative feedback will produce gain stability of sufficient accuracy to enable the overall error feedforward to vastly reduce nonlinear distortions in the complete system. The solution thus lies in a combination of the two principles, namely, error feedforward around negative feedback amplifiers.

2 NEGATIVE FEEDBACK GENERALIZATIONS

Before proceeding to a more detailed look at possible error feedforward realizations, we should mention an elaboration of the negative feedback scheme which has occasionally been confused with error feedforward. This is "active error feedback," to use the terminology of Macdonald [14], although similar schemes have also appeared much earlier. This occurs in two possible configurations, shown in Fig. 3(a) and (b). The former is what we shall call "input injected" and the latter what we shall call "output injected" active error feedback. At first sight these (particularly the latter) do look like true error feedforward, for the output signal e_0 is attenuated by β and then subtracted from the input signal e_i to produce the true "error" caused by the main amplifier A_1. This error signal is then amplified by A_2 and injected into either the input [case (a)] or the output [case (b)] of the main amplifier. This, however, is where the fundamental distinction from error feedforward is to be noted: the error is injected into the system *before* the point at which it is sensed, that is, before the β-network. This forms, in both cases, a potential regeneration loop as indicated in the figures, and as is apparent from the denominators of the gain formulas. True error feedforward senses the error before the point at which it is corrected, not after, and so avoids the instability totally. Examples[3] of input-injected active error feedback schemes include those of Baggally [15] in 1933, Macdonald [14] in 1955, Bollen [16] in 1973 and Klaassen et al.[17] in 1975. Klaassen dubbed his scheme "quasifeedforward," although, as shown above, it is in reality only negative feedback. Output-injected active error feedback circuits include those of Hähnle [18] in 1933, Ford [19] in 1945, Macdonald [14] in 1955, and McMillan's "multiple feedback" [20] in 1956. Considerable care is required in analyzing some of the more cunning schemes put forward for error reduction, but in most cases it will be found that in essence they reduce to one of the three

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[3] Please note that we do not make any pretentions of completeness in the references cited. We have cited what we believe to be a representative selection of the relevant sources.

schemes: negative feedback, active error feedback or error feedforward.

Although the two active error feedback schemes are not error feedforward, and do not admit of true error nulling even in principle, in spite of their use of two amplifiers, they are not quite the same as ordinary negative feedback, and so merit some further comments. First, apparently neither scheme need result in any gain reduction in the main channel A_1. This statement is, however, somewhat misleading, since both schemes *do* require excess gain (namely, G_2), but place it in the subsidiary channel. If analyzed on the basis of total available gain, which is really a fairer basis for comparison, they both do exhibit the same gain reduction as negative feedback. In fact, if we sever the upper input line (marked with a cross) in Fig. 3(a), it reduces precisely to Fig. 1 with a total gain $G = G_1G_2$. The effect of the upper input line is to introduce an extra (generally insignificant) term into the numerator of the gain formula in Fig. 3(a). The stability margin is, however, in both cases determined solely by the total loop gain $G\beta = G_1G_2\beta$, which also determines gain stabilization, distortion reduction and output impedance changes. As regards the scheme of Fig. 3(b), this is mainly of use where A_1 is a low-gain stage (such as, for example, an emitter follower) so that most of the gain lies in the auxiliary amplifier A_2 (that is, $G_2 >> G_1$). Its improvements are due solely to the loop gain $G_2\beta$; the gain
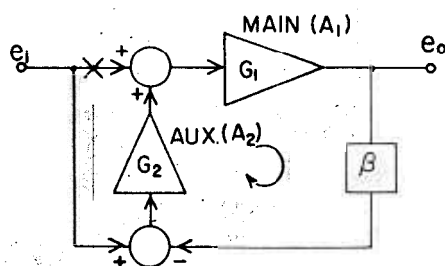
$$\frac{e_o}{e_i} = \frac{G_1(1 + G_2)}{1 + G_1G_2\beta} = \frac{1}{\beta} \quad \text{if } G_1\beta =$$

$$\underset{\infty}{\overset{G_2}{\rightarrow}} \frac{1}{\beta}$$

(a)
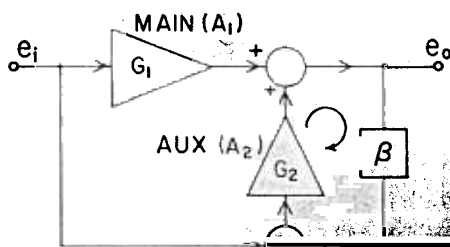
$$\frac{e_o}{e_i} = \frac{G_1 + G_2}{1 + G_2\beta}$$

$$\underset{\infty}{\overset{G_2}{\rightarrow}} \frac{1}{\beta}$$

(b)

Fig. 3. Active error feedback in its two forms. In each case the input/output error is derived, amplified by A_2, and injected into either the input (a) or the output (b) of the main amplifier A_1. The signal gain stabilizes at $1/\beta$ as $G_2 \rightarrow \infty$.

of A_1 does not contribute here. Otherwise the same comments as above apply here as well.

We should mention that a number of elaborate power output stages [21]–[23] have appeared recently, incorporating two separate amplifiers in their circuit topology, and thus looking superficially as if they might indeed be some form of error feedforward. In spite of claims to the contrary [21], they turn out instead to be clever applications of negative feedback. When analyzing such topologies, which may present themselves as the limiting case of a genuine feedforward configuration, one must beware of falsely ascribing to the limiting circuit those properties possessed by the general case.

3 ERROR FEEDFORWARD REALIZATIONS

At last we come to the true error feedforward realizations of Black's principle [3] for audio amplifiers, as opposed to VHF repeater amplifiers. We will cite Macalpine [24] in 1936, Ketchledge [25] in 1956, Sandman [7], [8] in 1973/1974, Klaassen et al. [9] in 1975, Walker and Albinson [4]–[6] in 1975/1976, the circuits in [26], [27], and Stocchino [28] in 1978. Of these, only Klaassen and Stocchino incorporate some form of time-delay compensation. There also exist some error feedforward realizations in which the author *has* the option of nulling the distortion due to the main amplifier but chooses for other reasons not to do so. These include McMillan [20] in 1956 and Davis' "total differential feedback" [29] in 1958. The primary differences between these audio realizations and those of Seidel et al. [10]–[13] are that, whereas in the VHF regime it is feasible to use true biconjugate networks (three winding transformers or Wheatstone bridges, for example) for error determination and nulling, for audio amplifiers of the highest quality, and especially for high-power amplifiers, this is no longer practical, and alternative arrangements must be used. Another simplification generally adequate, especially with a high-speed auxiliary amplifier A_2, is the elimination of time-delay correction networks τ_1 and τ_2 from the circuit of Fig. 2. [For if A_2 is sufficiently fast relative to the required (signal + error) – bandwidth, τ_2 can be eliminated; and we are always at liberty to redefine the "error" to include the time delay of amplifier A_1, so that we can also omit τ_1, the penalty, of course, being a greater demand on A_2.]

The omission of a true biconjugate output summing network is, as we have indicated above, generally necessitated by practical considerations when high-quality error feedforward audio (power) amplifiers are considered. This has two most important consequences, which can be seen by reference to Fig. 4(a) and (b), which shows two possible realizations of the Fig. 2 concept. Fig. 4(a) incorporates series addition of the two output signals and hence requires the inconvenience of a floating load, while Fig. 4(b) shows a passive (say, resistive) parallel output summing network in use. Examples of both output configurations will be found in the references cited. In both examples the lack of conjugacy between the outputs results first in mutual interaction between A_1 and A_2, and second in the formation of an undesirable negative feedback loop (as indicated) around

A_2 with its potential regeneration. Both circuits rely on the low output impedances of A_1 and A_2 to help prevent problems from these causes. In particular, the lower the output impedance of amplifier A_1, the greater the loop attenuation around A_2, and hence the less the feedback problem becomes. In practice, both the feedback loop and the mutual amplifier loading can be problems. In the series configuration of Fig. 4(a) each amplifier must sink the output current from the other, since both carry the full load current. This means that the auxiliary amplifier must have full output current sinking capability even if it is only really being used to correct for small errors caused by A_1. This is a major drawback of the series type of output summing realization. In the parallel type of output summing configuration shown in Fig. 4(b) one would generally have $Z_4 << Z_L$ and $Z_4 << Z_3$ (that is, $Z_3 \| Z_4 << Z_L$) for a practical power amplifier of low output impedance in which A_1 furnishes the bulk of the load current. The low output impedance of A_1 again reduces the significance of the regeneration problem, while a deliberate unbalance of the input path (by choosing $G_1\beta \neq 1$) and so allowing not just error signal but also some primary signal component through A_2, can virtually eliminate the current sinking requirements which A_2 would otherwise need to satisfy. It would thus appear that the Fig. 4(b) arrangement is the more practical for an audio power am-

plifier. It is, in fact, the basis of the first modern error feedforward audio amplifier to reach the market, namely, the Quad 405 "current dumping" amplifier described by Walker and Albinson in [4]–[6], and about which we shall have more to say in the sequel.[4] The output path balance condition $G_2\beta = 1$ which guarantees complete nulling of the distortion generated by A_1 now assumes a somewhat modified form because of the attenuation and lack of conjugacy of the output summing network. For, assuming that A_1 and A_2 both have zero output impedance, the output signal e_0 is related to e_1 and e_2 by

$$e_0 = \frac{Z_L}{Z_L + Z_3 Z_4/(Z_3 + Z_4)} \cdot \frac{Z_3 e_1 + Z_4 e_2}{Z_3 + Z_4} \propto Z_3 e_1 + Z_4 e_2$$

and so, to combine correctly to null the distortion at the output, it is necessary that e_2 and e_1 be in the ratio $Z_3 : Z_4$, that is,

$$\frac{e_2}{e_1} = \frac{Z_3}{Z_4}.$$

The output path balance condition $G_2\beta = 1$ thus now becomes

$$\frac{G_2 \beta e_1}{e_1} = \frac{Z_3}{Z_4}$$

that is,

$$G_2\beta = \frac{Z_3}{Z_4} \qquad (1)$$

while the condition $G_1\beta = 1$ for input path balance, because of Eq. (1), is equivalent to

$$\frac{G_2}{G_1} = \frac{G_2\beta}{G_1\beta} = G_2\beta = \frac{Z_3}{Z_4}$$

that is,

$$\frac{G_2}{G_1} = \frac{Z_3}{Z_4}. \qquad (2)$$

If we unbalance the input path to reduce the current sinking requirement of A_2, a good choice is $e_2 = e_1$ in Fig. 4(b), for then under no load ($Z_L = \infty$) A_2 sinks no current, while, provided $Z_3 \| Z_4 << Z_L$, the current sinking demand upon A_2 is greatly reduced. Referring to Fig. 4(b) we see that this requires that

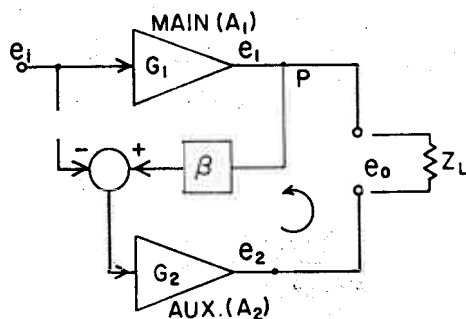$$e_2 = G_2\left(\frac{e_1}{G_1} - \beta e_1\right) = e_1$$

that is,

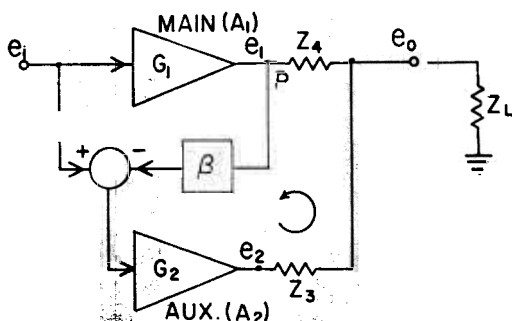$$\frac{G_2}{G_1} = G_2\beta + 1.$$

Using Eq. (1) this becomes

$$\frac{G_2}{G_1} = \frac{Z_3}{Z_4} + 1 \qquad (3)$$

$$e_0 = e_1 - e_2$$

$$\frac{e_0}{e_1} = G_1 + G_2 - G_1 G_2 \beta = \frac{1}{\beta} \quad \text{if} \begin{cases} \text{either } G_1\beta = 1 \\ \text{and/or } G_2\beta = \end{cases}$$

(a)

$$e_0 \propto Z_3 e_1 + Z_4 e_2$$

$$\frac{e_0}{e_1} \propto Z_3 G_1 + Z_4(G_2 - G_1 G_2 \beta) \propto \frac{1}{\beta} \quad \text{if} \begin{cases} \text{either } G_1\beta = 1 \\ \text{and/or } G_2\beta = Z_3/Z_4 \end{cases}$$

(b)

Fig. 4. Two practical realizations of the error feedforward idea of Fig. 2. The series type of output summing network (a) requires full output current sinking capability from the auxiliary amplifier, while the parallel network (b) can be arranged to considerably reduce the output requirements of amplifier A_2.

[4] The name "current dumping" amplifier arises from the role of the main amplifier A_1 whose function can be seen as that of dumping current into the load, any errors which it makes in the process being corrected by the auxiliary amplifier A_2. The apparently strange numbering of the impedances Z_3 and Z_4 is to ensure consistency with our later notation and that of [5].

which should be compared with the simple balance condition of Eq. (2).

At this stage a most important point should be mentioned. The error feedforward scheme is designed so that, provided output path balance is maintained, the errors of A_1 (both nonlinear distortion and gain errors) are totally canceled at the amplifier output. For example, even if A_1 becomes inoperative, the whole amplifier continues to operate with the same overall gain, but of course with reduced output drive capability. What this means is that the signal e_1 at the output P of A_1 can be *completely arbitrary* without affecting the output e_o. In practice, to reduce the load on A_2, we would obviously choose to make e_1 as close to the desired signal as possible, but the fact that the error-sensing point P is a "node" at which any injected signal is totally nulled from the output e_0 is most significant. For it means that the main amplifier A_1 *need* not be driven from the input signal e_1, but can be driven from any convenient point in the circuit. Fig. 5(a) illustrates the independence of the output upon A_1 by showing the latter replaced by an arbitrary signal generator e_1 injecting its output into point P: the output signal is totally independent of e_1 provided that the output path is balanced. Suppose that A_1 is a unity-gain emitter follower type output stage. Then it would make sense to drive it from the output of the auxiliary amplifier A_2, as illustrated in Fig. 5(b). This arrangement *automatically* ensures that the minimum current sinking condition $e_2 = e_1$ for A_2 is satisfied. The configuration of Fig. 5(b) is that of "current dumping." We note that the "dumper" with its low output impedance now aggravates the regeneration problem around A_2 by bridging the loop attenuation in Z_3 and Z_4 and introducing the dumper phase shift into the loop. This means that great care is needed in execution in order to ensure stability — current dumping involves a complicated balance of feedback and feedforward. It should also be noted that in the current dumping configuration only the dumper stage is enclosed within the error feedforward path. Since the auxiliary amplifier A_2 now carries a full-level primary signal component, its errors appear directly at the output and are not reduced in order, as occurs with simple feedforward when A_2 carries only error signal. In consequence, the performance of A_2 places a limit on the overall performance of such an amplifier. (This is also true of the circuit proposed by Ketchledge [25].)

Since we have now developed the error feedforward principle so that the feedforward ancestry of current dumping is clear, it is worthwhile to complete the full derivation of the actual circuitry used in this scheme. All that is required is a rearrangement of the input circuitry of Fig. 5(b) to incorporate an operational amplifier as A_2. This is shown in Fig. 5(c), where it will be noted that if Z_1 and Z_2 are chosen such that

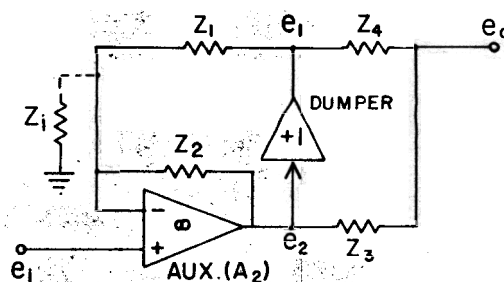$$\frac{Z_2}{Z_1} = \frac{Z_3}{Z_4} , \qquad\qquad (4)$$

that is, so that the four-component passive bridge formed by Z_1, Z_2, Z_3, and Z_4 is balanced, then the output path balance Eq. (1) is satisfied. This follows because the auxiliary path gain from e_1 to e_2 is $-G_2\beta = -Z_2/Z_1$. This ensures complete distortion nulling for the dumper, and minimal
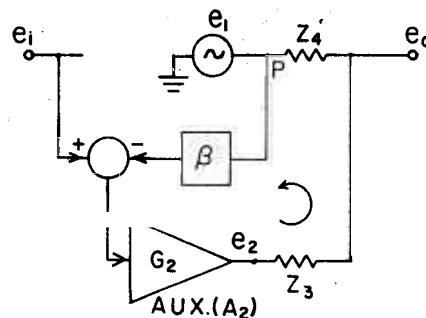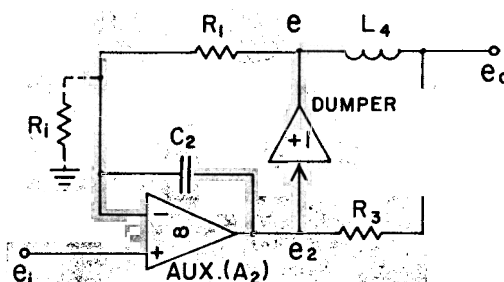

If $G_2\beta = \dfrac{Z_3}{Z_4} : \dfrac{e_o}{e_1} \propto Z_4 G_2 = \dfrac{Z_3}{\beta}$ independent of e

(a)
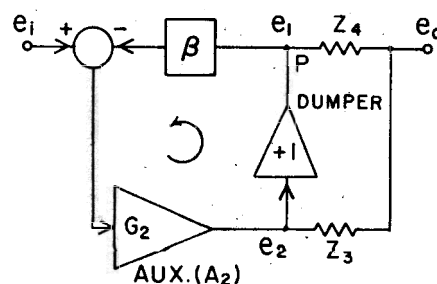
If $G_2\beta = \dfrac{Z_3}{Z_4} : \dfrac{e_o}{e_1} \propto Z_4 G_2 = \dfrac{Z_3}{\beta}$ independent of e_1

(b)

If $\dfrac{Z_2}{Z_1} = \dfrac{Z_3}{Z_4}$, e_o is independent of e_1

(c)

If $R_1 R_3 = \dfrac{L_4}{C_2}$, e_o is independent of e_1

(d)

Fig. 5. Conceptual development of "current dumping" from the error feedforward arrangement of Fig. 4(b). The independence of the output e_o upon e_1 (a) enables us to rearrange the main amplifier drive (b). Replacing A_1 with an operational amplifier leads to a four-component bridge (c) whose elements can be chosen to be complex (d).

power demand from the error-correcting auxiliary amplifier. Finally, noting that the Z's need not be real in order for bridge balance to be achieved, and since a practical power amplifier should have as low as possible an output impedance, we replace Z_4 with an inductor L_4 and hence let Z_2 become a capacitor C_2, thus obtaining the actual current dumping configuration shown in Fig. 5(d). (See [4]–[6].[5])

It is interesting to note that following the publication [5] of the current dumping circuit, much confusion arose concerning its true *modus operandi*, claims being made by some that it was not true error feedforward but simply a cunning application of negative feedback [30]–[33], while others correctly understood its operation [34]–[44]. It is hoped that the above derivation exhibits clearly its genuine feedforward nature.

As a final comment at this stage we remark that the main amplifier A_1 of an error feedforward circuit can be of entirely arbitrary design. We have already seen how an emitter follower can be accommodated in this position. However, we can be much more general than this. For example, A_1 could be a pulse-width modulation (class-D) power amplifier. Such schemes hold out considerable promise for high-efficiency very-high-quality high-power audio amplifiers. We shall have more to say on this concept later. A related idea has been suggested by Divan and Ghate [45] to construct a very-high-performance switching regulator.

4 ERROR TAKE-OFF AND CURRENT DUMPING

A useful artifice for extracting the true error signal (representing both gain error and nonlinear distortion) of the main amplifier A_1, if it is an operational amplifier, should also be mentioned here. This scheme is illustrated in principle in Fig. 6(a) and has been dubbed "error take-off" by Sandman [8], although it has also been used previously by McMillan [20] and Ketchledge [25]. If A_1 is perfect, its inverting input Q is a true virtual ground. Any signal appearing at Q represents an attenuated version $\beta\delta$ of the error component δ of e_1 appearing at P, given by

$$\beta\delta = \frac{Z_i}{Z_i + Z_1}\,\delta$$

Hence Q is a suitable error take-off point for driving the auxiliary amplifier A_2 (which must now be inverting), and for error nulling it is clear that the balance condition, Eq. (1), becomes

$$G_2 \cdot \frac{Z_i}{Z_i + Z_1} = \frac{Z_3}{Z_4} \tag{5}$$

If only nonlinear errors and distortions, but not gain errors, are required to be fed to A_2 (that is, if we wish to maintain input path balance), the error take-off point can be moved to a suitably chosen tapping along Z_1 at which point all primary signal components cancel.

This error take-off configuration can be iterated by mak-

ing A_2 itself an inverting operational amplifier whose error is sensed in turn at its inverting input, and corrected by a further auxiliary amplifier (A_3), whose output drives the load through a further output summing impedance. In this way, feedforward error correction of arbitrarily high order can be achieved. Such [20] and Sandman [8], and also mentioned by
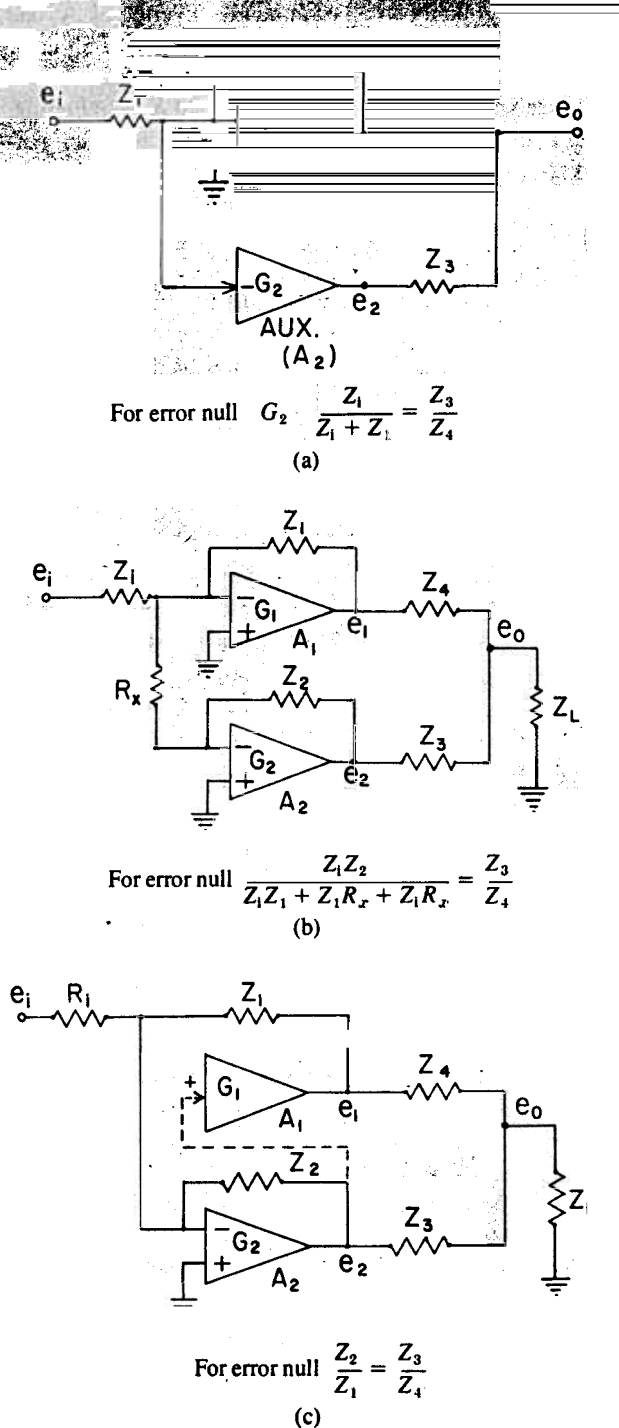

For error null $G_2 \dfrac{Z_i}{Z_i + Z_1} = \dfrac{Z_3}{Z_4}$

(a)

For error null $\dfrac{Z_1 Z_2}{Z_1 Z_1 + Z_1 R_x + Z_i R_x} = \dfrac{Z_3}{Z_4}$

(b)

For error null $\dfrac{Z_2}{Z_1} = \dfrac{Z_3}{Z_4}$

(c)

Fig. 6. (a) Sandman's "error take-off" principle, in which any errors (including gain errors as well as nonlinear distortions) caused by A_1 appear at its virtual ground summing point Q and are thus available for amplification by the auxiliary amplifier A_2 and subsequent subtraction from the output signal. (b) The circuit rearranged to make use of an operational amplifier for A_2. (c) Showing how "current dumping" is derived from (b). The resistor R_x has been removed, and A_1 can derive its input signal from the output of A_2. If A_2 has a large gain then there will be no distortion in e_0 if $Z_1 Z_3 = Z_2 Z_4$.

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[5] The impedances Z_i and R_i shown dashed in Fig. 5(c) and (d), respectively, enable overall gain adjustment to be made. Alternatively (see later) the noninverting input could be grounded and e_1 fed to the inverting input through Z_i (or R_i).

[25].

Fig. 6(b) shows the error take-off scheme derived from Sandman [8] using the iterative technique, but omitting the third amplifier A_3. The intent here is to have a signal e_1 which is $-(Z_1/Z_i)e_i$, but errors in A_1 will prevent exact attainment of this goal because of both nonlinear error and gain insufficiency. The error will appear at the input to A_1, and this presumably small error can be inverted by A_2 and added to the output at the correct level to cancel all error (both gain and nonlinearity) appearing at the load.

To analyze the circuit, let us assume that A_1 is imperfect

where δ is the total gain error, having both linear and nonlinear components. We assume for simplicity of analysis that A_2 has infinite gain. It is easy to show that e_2 is given by

$$e_o = -\frac{Z_2}{R_x}\left\{\frac{Z_1 R_x e_i + Z_1 R_x e_i}{Z_1 Z_1 + Z_1 R_x + R_x Z_1}\right\}$$

$$\frac{Z_2}{R_x} \cdot \frac{Z_1 R_x}{Z_1 Z_1 + Z_1 R_x + R_x Z_1}\delta \qquad (7)$$

so that e_2 contains only error. The output voltage e_o can be written conveniently in terms of e_i and δ, using the above equations for e_1 and e_2:

$$e_0 = \frac{Z_L \| Z_3}{Z_4 + Z_L \| Z_3}e_1 + \frac{Z_L \| Z_4}{Z_3 + Z_L \| Z_4}e_2$$

$$\left(\frac{Z_L \| Z_3}{Z_4 + Z_L \| Z_3}\right)\left(-\frac{Z_1}{Z_i}e_1 + \delta\right)$$

$$+ \left(\frac{Z_L \| Z_4}{Z_3 + Z_L \| Z_4}\right)\left(-\frac{Z_2}{R_x}\frac{Z_1 R_x}{Z_1 Z_1 + Z_1 R_x + R_x Z_1}\right)\delta$$

In order that the error terms in δ vanish we require that

$$\frac{Z_3}{Z_4} = \frac{Z_1 Z_2}{Z_1 Z_1 + Z_1 R_x + Z_1 R_x} \qquad (8)$$

which we regard as a balance condition to remove distortion from Z_L. Note that Z_L does not occur in the balance condition, as expected when the errors at the load are nulled to zero. We note also that the voltage is reduced from the intended $-(Z_1/Z_i)e_i$ by the passive attenuator formed by Z_3 and Z_4, regarding e_2 as a ground point. The output impedance of the amplifier is the parallel combination of Z_3 and Z_4 even though amplifier A_1 may have a large internal source impedance. Eq. (8) is the counterpart of Eq. (5) for the Fig. 6(b) realization.

In practice we would regard A_1 as the main power amplifier and A_2 as the error corrector. It would be logical to choose Z_4 quite small relative to Z_L, and Z_3 should be considerably greater than Z_4.

We now show how current dumping can be related to Sandman's error take-off of Fig. 6(b). Note that the amplifier A_1 does not really need an input. An incorrect signal for e_1 makes δ larger in Eq. (6), and A_2 simply has to work

harder to null this greater "error" by A_1. Thus A_1 can derive its input from any convenient signal that might reduce the load on A_2, for example. Additionally we note that R_x can be made zero with impunity, and the balance condition for zero distortion becomes, instead of Eq. (8),

$$\frac{Z_3}{Z_4} = \frac{Z_2}{Z_1} \qquad (9)$$

a condition of identical form to that of current dumping [Eq (4)]. Fig. 6(c) shows this arrangement. The input of A_1 may be left unspecified. In current dumping, the "dumper" A_1 is a high-power emitter follower of nearly unity gain, and to minimize the load on A_2, the input of the emitter follower can be driven from the error correction amplifier A_2, as shown by a dashed line.

The components Z_1, Z_2, Z_3, and Z_4 form a kind of bridge, and the balance condition Eq. (9) can also be written as

$$Z_1 Z_3 = Z_2 Z_4 \qquad (10)$$

Note that amplifier A_1 of Fig. 6(c) enters the circuit at what we shall define as a nodal point. Any disturbance applied to the terminal labeled e_1 does not appear at the load. This in no way implies that e_1 should be zero during load excursions. On the contrary, e_1 should be such that it supplies most of the load power through Z_4, so that the load on A_2 is minimized. It simply means that if A_1 supplies a distorted version of the intended signal as e_1, then the amplifier A_2 will supply any necessary error current, but naturally it would be best if this error signal were kept small.

5 ANALYSIS OF CURRENT DUMPING AND SEVERAL MODIFICATIONS

The only commercially available error feedforward audio power amplifier to date is the Quad 405, produced by the Acoustical Manufacturing Co. of Huntingdon, England. A simplified diagram of its operation is shown in Fig. 7. The components labeled Z_1, Z_2, Z_3, and Z_4 are chosen to satisfy the condition

$$Z_2 Z_4 = Z_1 Z_3$$

which is exactly analogous to Eq. (10) which applied to Fig. 6(c). An earlier publication of ours [42] has analyzed this
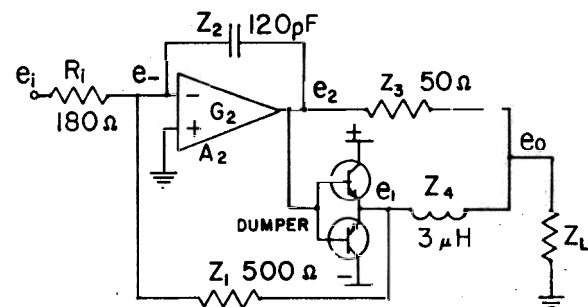

Fig. 7. A simplified diagram of the Quad 405 error feedforward amplifier. The input labeled e_1 could be grounded and the input signal applied to the noninverting input of A_2. If A_2 has a large gain, then for $Z_1 Z_3 = Z_2 Z_4$, e_o will be free of distortion. The output impedance of the amplifier is Z_3 in parallel with Z_4.

$$\frac{e_1}{Z_1} + \frac{e_1}{Z_1} + \frac{e_2}{Z_2} = 0 \qquad (12)$$

The voltage e_0 can be worked out in terms of e_2 and e_1 as follows:

$$\frac{e_2 - e_0}{Z_3} + \frac{e_1 - e_0}{Z_4} + \frac{-e_0}{Z_L} = 0 \qquad (13)$$

If we eliminate e_2 from Eqs. (12) and (13), we obtain

$$- \frac{e_1 Z_2}{Z_1 Z_3} - \frac{e_1 Z_2}{Z_1 Z_3} - \frac{e_0}{Z_3} + \frac{e_1}{Z_4} - \frac{e_0}{Z_4} - \frac{e_0}{Z_L} = 0. \qquad (14)$$

The dependence of e_0 on e_1 will be zero if the terms in e_1 vanish. This gives the condition $Z_2 Z_4 = Z_1 Z_3$, as shown in Eq. (11).

Fig. 7 shows a capacitor for Z_2 and an inductor for Z_4. This satisfies Eq. (11) for all frequencies. This ingenious idea allows almost the full output of the dumper to be applied to the load at low frequencies. An error correction signal through Z_3 need occur only at the crossover transitions of the power emitter follower "dumper."[6] In hindsight we note that in Sandman's error take-off scheme of Fig. 6(b) the balance condition Eq. (8) also admits of a solution such that Z_4 is an inductor and Z_2 is a capacitor.

If the gain of A_2 is not infinite, but given by G_2, and the output impedance is still low, then e_- cannot be neglected. Eq. (12) must be replaced by

$$\frac{e_1 - e_-}{Z_1} + \frac{e_1 - e_-}{Z_1} + \frac{e_2 - e_-}{Z_2} = 0$$

and e_2 is related to e_- by

$$-G_2 e_-$$

This together with Eq. (13) gives, after some algebra, the following balance condition:

$$Z_3 - \frac{Z_4}{Z_1} \cdot \frac{G_2}{1/Z_1 + 1/Z_1 + (1 + G_2)/Z_2} = 0 \qquad (15a)$$

or

$$\frac{Z_4}{Z_1} \cdot \frac{1}{1/G_2 Z_1 + 1/G_2 Z_1 + 1/Z_2 + 1/G_2 Z_2} = 0. \qquad (15b)$$

The second form shows clearly that as $G_2 \rightarrow \infty$, this condition is identical to Eq. (11).

Clearly it is not feasible simply to make Z_2 a capacitor and Z_4 an inductor, unless G_2 is very large. In the Quad 405 an analysis of the circuit shows that G_2 may have a gain–bandwidth that is very large, probably greater than 100

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[6] It should be noted that there is nothing inherent in the current dumping concept which requires the dumper stage to be biased off, so that it operates in class-C. If the dumper is operated with a quiescent current, it considerably eases the task of A_2 and improves ultimate performance.

such design is not easy and may lead to instability unless great care is taken to ensure proper compensation. For this reason we wanted to see if a solution to the balance condition Eq. (15) was possible if G_2 was chosen to be a normal single-pole compensated amplifier, such that its transfer function is approximately

$$G_2 = \frac{\omega_1}{s} \qquad (16)$$

where ω_1 is the unity gain frequency. Indeed a solution is possible. In Eq. (15a) we note that $(1 + G_2)/Z_2$ can be made purely real if Z_2 is a series resistor and capacitor, say R_2 and C_2. Then

$$Z_2 =$$

and this is purely real if

$$\omega_1 R_2 C_2 = \qquad (17)$$

We note also that if Z_i, Z_1, and Z_3 are real (resistors), then Z_4 can still be an inductor because of the numerator G_2 in Eq. (15a). The normal balance condition using a capacitor C_2 for Z_2, an inductor L_4 for Z_4, and resistors R_i, R_1, and R_3 for Z_i, Z_1, and Z_3, respectively, is

$$\frac{L_4}{C_2} = R_1 R_3 \qquad (18)$$

valid when G_2 is infinite. But for a single-pole compensated G_2 of the form given by Eq. (16), the balance condition Eq. (15a) results in two equations, one given by Eq. (17), and the other by

$$\frac{L_4}{C_2} = R_1 R_3 \left(1 + \frac{R_2}{R_i} + \frac{R_2}{R_1} \right) \qquad (19)$$

Eq. (19) is a significant modification to Eq. (18), especially if the unity gain frequency of A_2 is low, so that R_2 is not negligible relative to R_i and R_1. Using the component values of Fig. 7 and assuming a modest unity gain frequency of 10 MHz, it is easy to show that the bracket $(1 + R_2/R_i + R_2/R_1)$ has the value 2.002, and that the resistor R_2 to be added in series with C_2 has a value $132.6\,\Omega$. These are very significant changes from the original balance condition Eq. (18), and we felt that a verification of the foregoing theory was in order.

Fig. 8 shows the model circuit that we used to verify the basic features of the balance conditions above. The compensation capacitor C_c was chosen to give an appropriate gain–bandwidth, and R_2 and R_3 could be varied to minimize the distortion residual at the load as monitored by an accurately trimmed twin-T filter and an oscilloscope. A third variable resistor R_p was included to balance out the series resistance of L_4. Fig. 9 shows oscilloscope representations of best balance and the effect of unbalancing R_3, R_2, and R_p, respectively. An unbalance of R_3 results in a significant error residual created by the dumper transition. Setting R_2 to zero generates narrow error spikes due to the high-frequency gain behavior of A_2. A large change in R_p is necessary to make the error shown in the fourth trace. Omission of R_p is allowed if the series resistance of L_4 is

very low. It was found that with minimal compensation capacitance and $R_2 = 0$, the optimum value of R_3 was close to the calculated value of Eq. (18). But the error residual was still significant, although it represented a distortion of perhaps 0.01%, because the spikes are narrow. This incidentally is impressive performance for a complementary emitter follower output stage *with the two bases joined together*.[7]

However, when R_2 and R_3 were adjusted for minimum residual, even though the operational amplifier was compensated with a unity-gain bandwidth of 10 MHz, the error spikes were much less, giving a total harmonic distortion of about 0.001%. The values of R_2 and R_3 were in reasonable agreement with Eqs. (17) and (19). These aspects of current dumping are important, for they mean that total nulling of the error signal is possible for amplifiers A_2 having normal integrating characteristics.

An interesting aspect of dynamic balance is shown in Fig. 10. The upper trace shows the best achievable balance using the circuit of Fig. 8 with a 20-μH inductor made of many parallel strands of fine wire wound on a large former. The middle trace shows the best balance when the inductor is wound with 1.5-mm diameter wire on a small former. The error pulses are due to eddy currents in the thick wires. Such an inductor has different values at low frequencies and high frequencies. The time duration of the error pulses agrees qualitatively with an estimate based on the reciprocal of the frequency for which the skin depth of copper is 0.75 mm. It is clear that the detailed electrical nature of the components must be well controlled in current dumping when error signals of large bandwidth are allowed to occur.

One point which should be noted is that in Figs. 7 and 8 the basic configuration is inverting, and the input impedance that the source sees is approximately R_1, a relatively low value resistance. But of course the point labeled e_i could be grounded, and the signal could be introduced at the noninverting input of A_2. Such is the case for the actual Quad 405 amplifier. The analysis for such a case is slightly more complex, but it can be shown that the same balance condition Eq. (11) or Eq. (15) applies in either case. Without loss of generality we will use either configuration in our analyses.

A remaining problem in our analysis is the assumption that the amplifier A_2 has zero output impedance, although its gain is finite. A zero output impedance implies an infinite internal gain with negative feedback applied. In practice the output impedance of A_2 may be quite low. For example, the open-loop output impedance of the NE 5534 operational amplifier in Fig. 8 is about 50 Ω. If we retrace the theory leading to the balance condition Eq. (15) but include the output impedance Z_0 of the amplifier A_2, the balance condition is found to be (after considerable algebra)

[7] See Footnote 6.

As $Z_0 \to 0$, this balance condition reduces to Eq. (15). The increased complexity is hard to visualize. If Z_0 is a pure resistance, the only obvious solution is for G_2 and all the Z's to be real. Even then the balance condition does not take into account the nonlinear input current of the emitter follower dumper. In our earlier work [42] we showed that as long as the current gain of the dumper is constant, a balance condition exists. We felt that a better solution was possible and issued a challenge in the earlier article to produce circuits that would null the current transfer distortion as well.
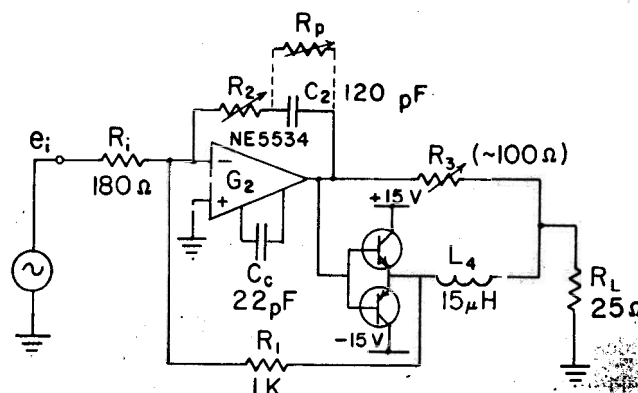

Fig. 8. A model circuit used to check the balance conditions using a normal operational amplifier for A_2, with single-pole compensation. The balance condition for R_3 can be significantly different than before, and R_2 is determined by a separate equation. In practice it was found advantageous to parallel C_2 with a resistance R_p of about 250 kΩ to balance out the series resistance of the inductor L_4.
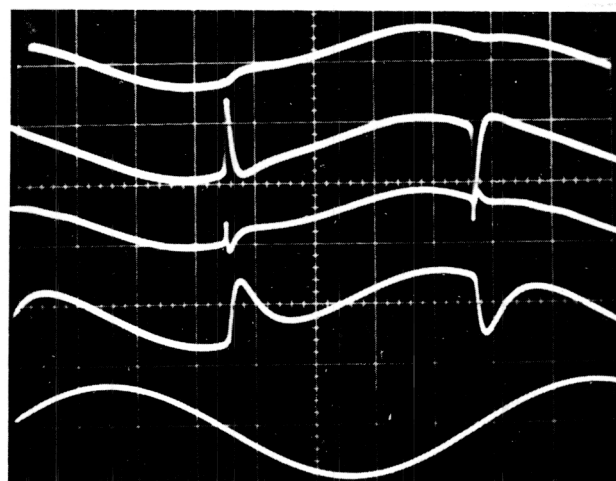

Fig. 9. An oscillogram of the performance of the model circuit of Fig. 8. The top four traces represent the output across the load resistor with the 13.2-kHz fundamental largely removed by a twin-T filter. The sensitivity is 10 mV/div. The lowest trace is the load voltage at a sensitivity of 10 V/div. The top four traces show sequentially the effect of: (1) proper balance of the model circuit, (2) an increase of R_3 by 16%. (3) setting R_2 equal to zero, (4) reducing R_p from 250 kΩ to 60 kΩ. The values of R_2 and R_3 for balance are around 150 Ω. Time scale 10 μs/div.

$$Z_3 - \frac{Z_4}{Z_1} \frac{\left(\dfrac{G_2}{Z_0} - \dfrac{1}{Z_2}\right)}{\dfrac{1}{Z_2}\left(\dfrac{G_2}{Z_0} - \dfrac{1}{Z_2}\right) + \left(\dfrac{1}{Z_0} + \dfrac{1}{Z_2} + \dfrac{1}{Z_3}\right)\left(\dfrac{1}{Z_1} + \dfrac{1}{Z_2} + \dfrac{1}{Z_i}\right)} = 0 \qquad (20)$$

This challenge was very elegantly answered by Hevreng [43], [44] whose analysis we present below. The solution also turns out to solve in a better way the problem just discussed above, namely the output impedance of the error correction amplifier. The essential point is that as more current is taken by the bases of the dumper, an extra signal must be found to cause additional error correcting action. Fig. 11(a) shows how this is achieved. The extra element Z_b included in the bridge causes an additional correction across Z_3 when the dumper draws base current. The amplifier A_2 has voltage gain G_2, output impedance Z_0, and has an output current I. The defining equations for this circuit are:

When e, e_1 and I are eliminated from these equations, the resulting equation is

$$e_i = e_o \left[\frac{\dfrac{Z_4}{Z_3} + \dfrac{Z_4}{Z_L}}{1 + \dfrac{Z_1}{Z_2} + \dfrac{Z_1}{Z_i}} + \frac{Z_0}{G_2 Z_3} \right]$$

$$+ e_2 \left[\frac{Z_0}{G_2} \left(\frac{1}{Z_3} + \frac{1}{Z_b} + \frac{1}{Z_0} \right) + \frac{Z_4}{Z_3 \left(1 + \dfrac{Z_1}{Z_2} + \dfrac{Z_1}{Z_i} \right)} \right]$$

$$+ e_b \left[\frac{\dfrac{Z_0}{G_2 Z_b}}{\left(1 + \dfrac{Z_2}{Z_1} + \dfrac{Z_2}{Z_i} \right)} \right] \qquad (21)$$

We note that both the terms in e_2 and e_b can with proper selection of components be set equal to zero. Thus there are now two balance conditions. It may have been instructive to have left e_b and e_1 instead of e_2 and e_b. But that really does not matter: the same balance conditions would have arisen. The dumper distortion can be thought of as being applied as e_1, but a version of that distortion will occur in e_2 as well. If we ignore the effect of Z_b, then $e_b = e_2$, but the null for e_2 (or e_1) still occurs. This has produced a nodal point e_1. We note that with the introduction of Z_b, a new null condition occurs and e_b becomes a nodal point as well. Any current drawn from e_b (or signal injected into e_b) will not produce a distortion in e_o, as we saw earlier for e_1.

The null condition for eliminating e_2 (or e_1) from Eq.

(21) can be written

$$\frac{Z_4}{Z_3} = \frac{Z_0}{G_2} \left(\frac{1}{Z_3} + \frac{1}{Z_b} + \frac{1}{Z_0} \right) \left(1 + \frac{Z_1}{Z_2} + \frac{Z_1}{Z_i} \right) \qquad (22)$$

and the condition to eliminate e_b is

$$Z_b = \frac{Z_0}{G_2} \left(1 + \frac{Z_2}{Z_1} + \frac{Z_2}{Z_i} \right) \qquad (23)$$

If we use Eq. (23) to simplify Eq. (22), we have

$$\frac{Z_3 + Z_b}{Z_4} + \frac{Z_3 Z_b}{Z_4 Z_0} = \frac{Z_2}{Z_1} \qquad (24)$$

This completes Hevreng's analysis. It is worth noting that if G_2 is large, Z_b tends to zero, and if Z_0 is very small Z_b also becomes small. The null given by Eq. (22) or Eq. (24) is determined by definite nonzero components as $G_2 \to \infty$, whereas in the same limit the null condition Eq. (23) vanishes as $G_2 \to \infty$ or $Z_0 \to 0$. But Hevreng's elegant solution to the current transfer distortion is important, for it shows that a linear amplifier A_2 can give precise error correction, although it has finite gain and nonzero output impedance.

The foregoing theory was tested with our model circuit. In a setup essentially that of Fig. 11(a), we found (with the dumper removed) that signals injected at e_b or e_1 did not appear in e_o when the bridge elements were adjusted in accordance with Eqs. (23) and (24). The amplifier A_2 was a differential amplifier with its own internal feedback, so that the G_2-dependence of Eq. (23) could be checked. The circuit is easily trimmed (using all resistive elements) to make the points e_b and e_1 act as nodal points. The number of nodal points can be extended indefinitely by breaking up the $Z_2 Z_b$ or the $Z_1 Z_4$ arms of the bridge into parallel units having the same impedance ratio and the same total conductance as the original arm, as drawn in Fig. 11(b). The coefficients α or β could be complex numbers, so that the different parallel arms do not even need to have similar components.
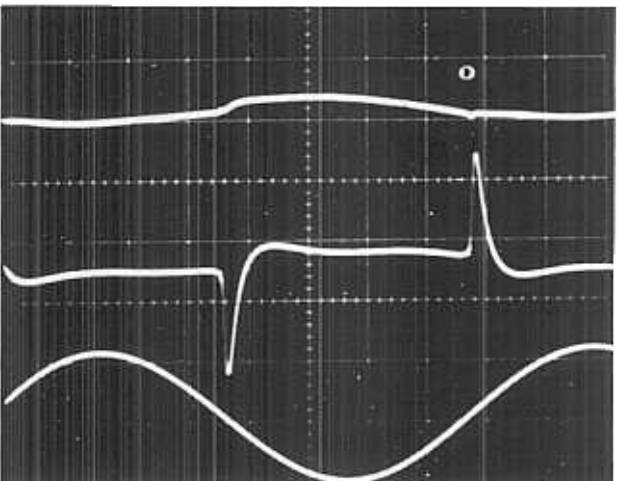

Fig. 10. An oscillogram showing the performance of the model circuit of Fig. 8. The two upper traces at 10 mV/div. are the output viewed through a twin-T filter which removes most of the 13.2-kHz fundamental signal. The lower trace is the output at 10 V/div. The upper trace is the best balance with a 20-μH inductor L_4 made of many parallel strands of fine insulated wire on a large-diameter former. The middle trace is the best balance achievable when L_4 is wound with 1.5-mm diameter wire on a small diameter former. Time scale 10 μs/div.

A number of realizations of the balance conditions Eqs. (23) and (24) exist, among them those which have a capacitor for Z_2, and for Z_4 an inductor in series with a resistor. It is not known whether a realization exists for amplifier gain G_2 of the form given by Eq. (16)

The practicality of using Z_b is not clear. In a realistic circuit one would try to make Z_0 small and G_2 large. Then the error caused by the neglect of Z_b may be insignificant.

6 ERROR FEEDFORWARD IN CLASS-D AMPLIFIERS

The switching (or class-D) amplifier has inherent limitations that make error feedforward attractive. The combined properties of high efficiency and low distortion are difficult to achieve simultaneously, since a very high switching frequency which reduces distortion usually has lower efficiency. The application of error feedforward allows the design of a high-efficiency brute-force class-D amplifier

with a relatively small correction amplifier. But such designs have a few characteristics which require additional techniques peculiar to class-D amplifiers.

The use of a switching amplifier requires a reactive output filter to remove the high-frequency components, allowing only the audio to be passed, and returning switching currents to the supply for efficiency. Such a filter will have an appreciable delay, amounting to a significant fraction of a cycle at 20 kHz. Such time delay means that the error-correcting amplifier must supply large currents. For example, suppose that there is a time delay τ representing a phase shift of $\phi = \omega\tau$ at frequency ω. On a straightforward basis this means mathematically that the error waveform will have a primary signal component which is $2 \sin (\phi/2)$ times e_1. For a 15° phase shift this represents an amplitude that is 26% of the main amplifier output. To keep the demand on the error correction amplifier small, a delay line may be employed, as shown in Fig. 12(a), to ensure that the input signal is delayed so that synchronous extraction of the error takes place.

In Fig. 12(a) the LC filter in the switching amplifier has some delay as well as the delay due to the switching modulator itself, and the intent is to balance out this delay with an equivalent delay τ in the application of the input signal to
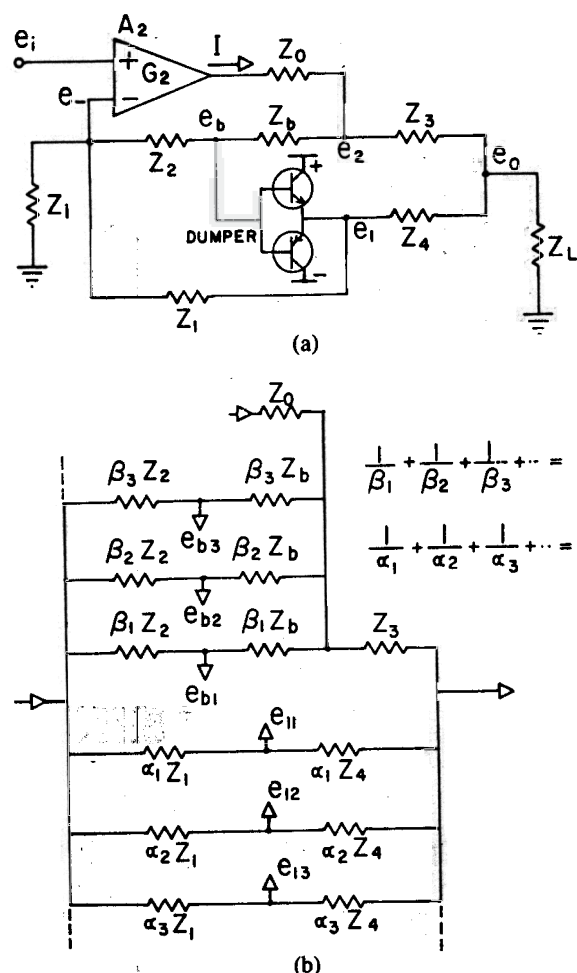

(a)

(b)

Fig. 11. (a) Hevreng's modification to current dumping which nulls out distortion due to the nonlinear current gain of the dumper using an error correction amplifier with finite output impedance Z_0. The new element Z_b causes an additional correction through Z_3 whenever the dumper draws input current. This circuit uses a noninverting configuration, but the balance conditions are the same as for the inverting version. The circuit has two "nodal" points labeled e_b and e_1: any external signals applied to these points do not appear at the output e_0. (b) Showing how the number of "nodal" points in (a) can be increased indefinitely. The α and β coefficients (which could be complex) are such that the total conductance and division ratio of the branches are the same as for the original branch.
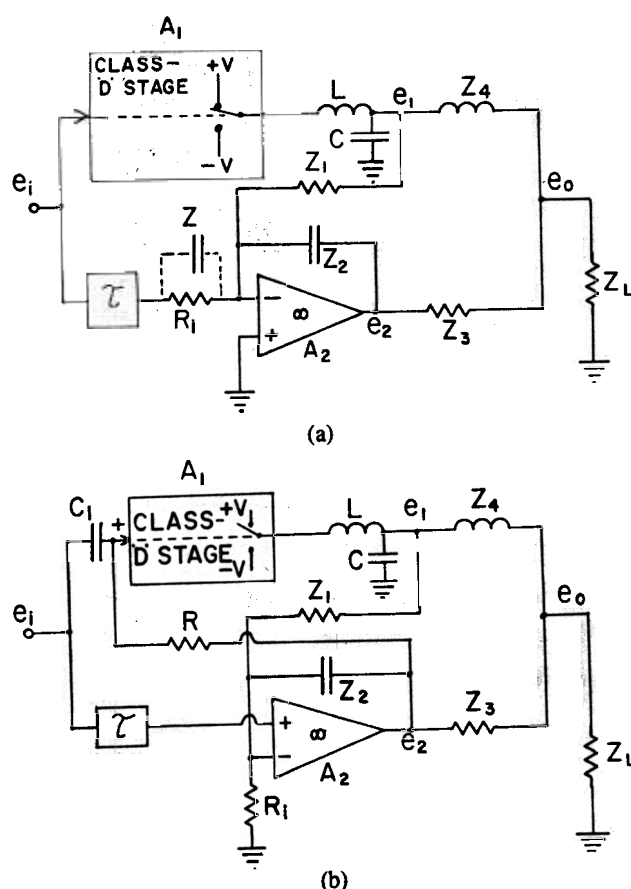

(a)

(b)

Fig. 12. (a) An illustration of how error feedforward might be applied to a class-D switching amplifier. The output LC filter is regarded as part of A_1, the class-D stage. The delay line τ is added so that the delay in e_1 is roughly compensated, hence allowing a smaller error signal. Z is explained in the text. (b) Class-D error feedforward with compensation for possible dc offsets in e_1. R prevents low-frequency saturation of A_2, while signals are passed by C_1. Careful design is necessary in this configuration to prevent instability.

the error amplifier A_2. If e_1 has no error, then we would expect e_2 to be zero provided Z_1/R_1 has been chosen to be equal to the gain of the class-D stage; however, this would allow a large current to flow through Z_3. A more sensible choice is to arrange $e_2 \approx e_1$, so that the current load on A_2 can be low. The dashed component labeled Z will make $e_2 = e_1$ if $Z_1/R_1 = Z_2/Z$. Note that R_1 and Z are not necessary for error correction; they are chosen to reduce the current load on A_2.

A remaining problem in the circuit of Fig. 12(a) can occur if Z_4 is an inductor as drawn. Suppose that the class-D stage allows a small dc error offset in e_1. The error correction will attempt to remove this at e_0, and so the current in Z_4 will increase uniformly with time, being supplied by A_2 through Z_3. Hence e_2 will increase with time until A_2 saturates. In experiments with a model class-D stage we have noticed this problem not just for dc offsets, but for high amplitudes at low audio frequencies for which Z_4 represents a small impedance. The circuit of Fig. 7 eliminates this problem by making A_1 an emitter follower and deriving its input from e_2. This solution is not practical in the present scheme of Fig. 12(a), because the time delay of the class-D output filter will certainly cause instabilities. A partial solution to the problem is illustrated in Fig. 12(b) in which the high-frequency input signals are capacitor coupled directly to the class-D stage, whereas saturation of A_2 is averted by feeding the low-frequency signals from a properly scaled e_2. In this way the stability problem can be overcome. Our model circuit incorporated all the features shown in Fig. 12(b), and the results were encouraging, although features such as the stability problem require more theoretical and experimental work.

A class-D amplifier with error feedforward would combine the qualities of high efficiency and low distortion. An additional point is that normal class-D amplifiers have an output filter whose impedance variations, though small, are not as low as a regular class-AB amplifier with feedback. With error feedforward, the output impedance is Z_3 and Z_4 in parallel, which can be quite low.

A significant advantage of error feedforward is the possibility of complete removal of switching frequency leakage. Although the class-D stage with its filter will still produce some residual RF output, the error correction loop can remove this totally in principle (see also [45]). However, one would expect the amplifier A_2 and the bridge components to become complex in their behavior at high frequencies, and attention to detail regarding high-frequency balance is essential, as we have found with our model.

7 CONCLUSION

The thrust of our paper is to inform the audio community of the history of error feedforward, its recent reemergence, and the analysis of the circuit topology of a current market product. The advantage of error feedforward in class-D amplifiers is obvious, but there are several peculiarities of this approach to switching amplifiers. We regard error feedforward as an alternative to negative feedback with some definite advantages. We do not consider it a cure-all for audio amplifiers, many of which we believe have reached audible perfection. The pursuit of ultra-low distor-

tion for its own sake is not meaningful without reference to the hearing process.

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