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TOWARDS A GENERALISATION OF ERROR CORRECTION AMPLIFIERS

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0. Introduction

The "current dumping" amplifier was introduced at the 50th Convention of the AES (1) in 1975 by Peter Walker and Michael Albinson and represented a milestone in the evolution of analogue amplifers. Until that time, most transistor power amplifiers had been variations of directly biased output stages operating in either class A or class AB, together with overall negative feedback, to achieve acceptable linearity, although there were already several innovative developments, such as the work of Blomley (2)

Naturally, tube technology was well developed at this time, where device characteristics dictate rather different system topologies since tubes are essentially high voltage, low current devices and there is no "PNP" equivalent. Also, error reduction schemes were established for use with tube electronics well before the Walker amplifier. For example, a patent by Llewellyn (3) describes a method of error feedback distortion reduction, though simultaneous feedforward correction was not cited. Feedforward error correction however was originally described by the Black patent (4) and although it is used in high frequency circuits (5), feedforward had not, prior to 1975, found application as a correction procedure in audio power amplifiers. However, local feedforward within an overall feedback loop had been successfully applied in the tube circuitry of AR (6), where the cathodes of the output tubes effectively feed across the primary to secondary of the output transformer and directly couple to the load impedence, although no attempt was made to seek a topology capable of a balance condition equivalent to the Walker circuit (1).

Since 1975, there has been extended debate as to the virtues and fundamental principles of current dumping. Some researchers have proposed a balanced bridge analogue (7) to explain the distortion null, while others such as Nigel Allinson (8) have correctly recognised the combination of both feedback and feedforward within a common structure. While yet a further school (9, 10) has attempted to deny the existance of the mechanism of feedforward distortion correction preferring what appears to be an impractical overall feedback loop (canonic form), that neglects the elegance of the original concept, which exhibits both a natural empathy with real device characteristics and a true error null.

In this paper, we re-affirm the existance of the "current dumping" principle and extend the comparative discussions by demonstrating equivalance with a more general composite error feedback, error feedforward model. It is not intended that this model invalidates other critical observations, rather that it complements them by taking an alternative stance, aimed primarily at linking earlier work on error feedback and error feedforward (11, 12, 13).

We then conclude the comparative discussion by proposing a structure common to control engineering, analogue computing and transient analysis from which many of the present day distortion correction systems can be derived. This re-inforces the foundation of error correction and allows a vehicle for identifying new topologies that exhibit broad-band distortion correction. To demonstrate error correction, we use the transfer error function (14) as an indicator of both system performance and a means of identifying and classifying sets of system balance equation to achieve distortion nulling. This error function as well as expressing the system error, also allows the sensitivity of the balance equation (5) to be directly evaluated.

1. Primitive model of error feedback/feedforward

Error correction, rather than error reduction, implies there is a balance equation (or equations), which under optimal alignment exhibits a broad band distortion null rather than just a reduction in distorion. The apparent implication is that the canonic equivalent feedback (only) loop must allow infinte loop gain over a broad band of interest. However, for simple feedback, this theoretic requirement is impossible, an observation recently emphasised by Lipshitz and Vanderkooy (15), a paper forming a useful complementary discussion.

To achieve a theoretic broad-band distortion null at least one feedforward path that extends beyond the feedback loop is mandatory, to compensate for the limitations of any practical feedback loop that can be devised. We concur with Allison (8), Vanderkooy and Lipshitz (11) that this compenstion is a fundamental requirement, any system attempting to eliminate the feedforward path yet attain zero distortion is impractical. Of course, in making this observation we do not deny the low distortion achievments possible with feedback, we are considering the limiting case: what is achievable with feedback alone can, in principle, be enhanced with the inclusion of feedforward. However, in practical systems, device characteristics may well negate the performance advantages offered by a particular system philosphy, where there has already been much debate (16, 17, 18).

Let us commence by re-examining the error correction strategy proposed in an earlier paper (12), which offered both error feedback and error feedforward and which in combination enable a true distortion cancellation of the error arising from the non-linear output cell, N.

In this primitive model illustrated in Fig 1-1, the overall voltage gain for a given set of parameters {N, a, b} is A, where the corresponding target gain $A_t = 1$. Also, at any instant, the non-linear output cell is assumed to have an incremental gain N, where in this example for zero distortion $N \rightarrow 1$, which is compatible with the adoption of $A_t = 1$.

Hence observing the respective coefficients {a, b} in the error feedback and feedforward paths in the scheme of Fig 1-1, the overall transfer function A for non-optimal parameter alignment is,

$$A = \frac{N - b(N-1)}{1 + a(N-1)} \qquad \text{1-1}$$

Defining the corresponding transfer error function (14, 19) E_1, as,

$$E_1 = \frac{A}{A_t} - 1 \qquad 1\text{-}2$$

and substituting for A, noting that in this case $A_t = 1$, then

$$E_1 = (N-1)\left[\frac{1-(a+b)}{1+a(N-1)}\right] \qquad 1\text{-}3$$

For this example, equation 1-3 shows that the error function E_1, tends to zero when either N=1 and/or more fundamentally, when the numerator contains the relationship,

$$a+b = 1 \qquad 1\text{-}4$$

which is the balance condition that enables a theoretic distortion null, providing equation 1-4 is maintained over an adequate bandwidth.

It is here that the need for a feedforward path that extends beyond the feedback loop is evident: the parameter 'a' is determined by the stability constraints of feedback, while the factor 'b' can be freely selected to yield a broadband distortion null as it is independent of factors affecting loop stability. We therefore conclude that both feedback and feedforward are complementary to achieving exact error correction as a = 1 cannot be attained over a broad band.

2. Conceptual equivalance of error feedback/feedforward to current dumping

Consider the re-configured schematic in Fig 2-1, noting its equivalence with Fig 1-1. This system is conceptually similar to the Walker amplifier (1) and the later derivative amplifier offered by Sansui (Super - ff) (20). At this stage this equivalance is less obvious as the bridge components of the current dumping amplifier are excluded, though the respective feedback and feedforward paths are identified; we will shortly extend the schematic to include the bridge that is more commonly associated with the Walker interpretation of current dumping.

The loop gain A_ℓ of the negative feedback path in Figs 1-1, 2-1 is,

$$A_\ell = -N\frac{a}{(1-a)} \qquad 2\text{-}1$$

where in this form, the term a/(1 -a) can represent the gain in the forward path.

If we observe the feedback loop in the original configuration of Fig. 1-1, the parameter 'a' can be selected as a first-order, low pass filter to achieve a stable loop, and is a typical for an error feedback loop (14).

$$\text{i.e. } a = \frac{1}{1 + j\omega\tau_a} \qquad 2\text{-}2$$

Observe how as $\omega \rightarrow 0$, $a \rightarrow 1$ which achieves optimum distortion correction at dc, while the time constant τ_a establishes the dominant loop break frequency, although unfortunately it also prevents optimum distortion correction by forcing $a < 1$ for $\omega > 0$.

It is common practice in feedback amplifiers, including the Walker amplifier, for the forward gain to take a form approximately to $A_0/(1 + j\omega A_0\tau_1)$, where A_0 is the dc gain and $1/(2\pi\tau_1)$ is the gain-bandwidth product (ie the frequency at which the loop gain, for a first-order system, is unity), where the form of the function is illustrated in Fig. 2-2.

Hence, noting in Fig. 2-1 that the loop gain (excluding the output stage of incremental gain N as N≈ 1) has the equivalent form a/(1-a), we can set

$$\frac{a}{1-a} = \frac{A_o}{1 + j\omega A_o\tau_1} \qquad 2\text{-}3$$

whereby,

$$a = \frac{A_o}{(1 + A_o)} \cdot \frac{1}{\left[1 + j\omega\frac{A_o\tau_1}{(1 + A_o)}\right]} \qquad 2\text{-}4$$

Hence, for the special case where $A_0 \rightarrow \infty$, then $a \rightarrow 1/(1+j\omega\tau_1)$ and $A_\ell \rightarrow -N/(j\omega\tau_1)$, a result corresponding to equation 2-2 where $\tau_a = \tau_1$.

We observe in this example that in the canonic form, the loop filter is an integrator in cascade with the non-linear output cell N, a condition representative of negative feedback power amplifiers, where providing N is well behaved at high frequency, good stability margins can be achieved.

The optimum feedforward parameter 'b' to achieve broad band error correction is determined from equations 1-4 and 2-4 as

$$b = \frac{1 + j\omega A_o\tau_1}{(1 + A_o)\left(1 + j\omega\frac{A_o\tau_1}{1 + A_o}\right)} \qquad 2\text{-}5$$

which for $A_o \to \infty$, simplifies to

$$b\Big|_{A_o \to \infty} = \frac{j\omega\tau_1}{1 + j\omega\tau_1} \qquad 2\text{-}6$$

In Fig 2-3, the Fig 2-1 structure is further modified to show how feedforward error addition can be performed by an output network L_o, R_o, while the forward gain a/(1-a) is represented by equivalent integrator transfer function (for $A_o \to \infty$) as $1/(j\omega\tau_1)$. Hence, comparing the coefficients b, (1-b) in Fig 2-1 with the L_o, R_o network of Fig 2-3, it follows that,

$$b = \frac{j\omega L_o/R_o}{1 + j\omega L_o/R_o} = \frac{j\omega\tau_1}{1 + j\omega\tau_1}$$

$$(1\text{-}b) = \frac{1}{1 + j\omega L_o/R_o} = \frac{1}{1 + j\omega\tau_1} .$$

i.e. assuming the forward gain exhibits a first-order response, the error summation exactly matches the requirement for optimum distortion correction, providing

$$\tau_1 = \frac{L_o}{R_o} \qquad 2\text{–}7$$

Before completing the discussion of conceptual equivalence of the Fig 1-1 structure with the Walker amplifier, let us investigate how the system can be adapted to include overall gain. In Fig. 2-4, the scheme of Fig. 2-1 is reconfigured to include a feedback parameter k. However, by redefining the forward gain to be a/(1-a)k, the loop gain remains unaltered as do the equations derived for optimum balance. The final stage in our comparison with the Walker amplifier can now be made.

We have observed that for a large value of A_o, there is a near-optimum value for 'a' of unity, where this parameter $[A_o/(1+A_o) \to 1$ as $A_o \to \infty]$ is very insensitive to changes in A_o, ie. at low frequency almost perfect error correction is achieved by using a large loop gain. We note also that although the forward path has a low 3dB break frequency ($\omega = 1/A_o\tau_1$), when translated to parameter 'a' by equation 2-4, this frequency is multiplied by a factor $(1 + A_o)$, implying a break frequency tending to ($\omega = 1/\tau_1$). Therefore, an important circuit attribute of the Walker topology (1, 21) is that although the dc gain A_o maybe poorly defined, the terms a and τ_1, are more accurately specified, consequently, when selecting 'b' in the feedforward summing network, the balance is both predictable and stable.

In completing our comparison, we concur with Walker, by noting the forward gain requires a close adherence to an integrator transfer function. In principle, the integrator can be configured using a virtual-earth technique, where effectively the gain between X and Y, -a/(1 - a) in Fig 2-4, is formed using an amplifier incorporating a local, frequency selective feedback network, where the basic circuit is presented in Fig 2-5a. In Fig 2-5b, the more complete current dumping topology is shown, while in Fig 2-5c, the equivalent descrete circuit with associated capacitance of the Walker amplifier (1, 11) is illustrated, which can also yield an excellent approximation to the required integrator response. It follows from the circuits of Fig 2-5 a, b that the gain between nodes X and Y is,

$$\frac{-a}{(1-a)} = \frac{-1}{j\omega R_1 C_1}$$

$$\text{whereby,}\ a = \frac{1}{1 + j\omega R_1 C_1}$$

and corresponds to equation 2-4 where $\tau_1 = R_1 C_1$ and $A_o \rightarrow \infty$.

Hence using the balance condition stated in equation 1-4 and the result of equations 2-6, 2-7 the original Walker expression for balance follows,

$$\text{i.e. } R_1 C_1 = L_o/R_o \qquad \text{2-8}$$

This Section has shown, that at a conceptual level, "current dumping" can be usefully compared with a composite error feedforward/error feedback structure. More important however, is the requirement for an error feedforward path that extends beyond the main feedback path to realise theoretic broad band distortion cancellation. The Walker amplifier would appear to be the first power amplifier to exploit this critical and fundamental requirement. Essentially it places the majority of the error reduction within the feedback path, which reduces sensitivity to imbalance and then simultaneously uses error feedforward for fine tuning the alignment of the balance condition at high frequency, and thus achieves true error correction - also of importance is that the error feedforward path does not require gain, allowing a passive summation network to be used (see Section 4 for further discussion). Consequently, the non-linear and time dispersive errors within the feedback loop that arise from output stage non-linearity are negated by feedforward.

3. Error feedforward/feedback correction with arbitrary gain

The correction topology decribed in Section 1 is restricted to a non-linear output cell that has a nominal gain of unity, where the balance condition in this primal example established an overall gain also of unity. However, in this Section, we generalise the error feedforward/feedback structure to a system with arbitary gain paramenters, where in general N > 1.

Since an accurate definition of overall gain is required, it is necessary to imbed a reference amplifier as shown in Fig 3-1, whose voltage gain R is, ideally equal to the overall target gain A_t. This strategy enables an output stage of incremental gain N, where N > 1, to be combined with error feedback/feedforward to fine tune the overall voltage gain, minimise distortion and thus achieve a better approximation to the target gain A_t.

With reference to Fig 3-1, the overall voltage gain A_h of the high gain system is,

$$A_h = \frac{N - b(N-R)}{1 + a(N-R)} \qquad 3\text{-}1$$

and the corresponding error function E_h is,

$$E_h = \frac{A_h}{A_t} - 1 \qquad 3\text{-}2$$

$$\text{i.e. } E_h = (N-A_t)\left[\frac{1 - (b + aA_t)\dfrac{(N-R)}{(N-A_t)}}{1 + a(N-R)}\right] \qquad 3\text{-}3$$

Examination of equation 3-3 shows that the error function is zero for the two sets of conditions,

$$\begin{cases} R = A_t & 3\text{-}4 \\ aA_t + b = 1 & 3\text{-}5 \end{cases}$$

$$\text{and/or, } N = A_t \qquad 3\text{-}6$$

Following a similar procedure to Section 1, we reconfigure Fig 3-1 to the structures shown in Fig 3-2 a,b. It is at once apparent that the first system is conceptually similar to the Walker (1) and Sansui (20) correction schemes, while the second is similar to Sandman's (22) error pick-off system. The disadvantage of the second system is of course the requirement for an extra amplifier with gain in the error feedforward path that is a function of the choice of coefficients in Fig 3-2(a), where for example: if $N = A_t$, a = 0.04, b = 0.20 then $aA_t = 0.80$ and $bR = bA_t = 4.00$, assuming $R = A_t$ and $aA_t + b = 1$.

4. Towards a more general error feedforward/feedback topology

A more general appraisal of distortion correction systems reveals that some systems can be decribed as an extension of the constant voltage class of filter structure, where under appropriate alignment, limited non-linearity within the filter topology is permissible. In fact, the Walker amplifier (1), Sandman's error pick-off (22) and feedback/feedforward topologies (8, 12) which all accommodate broad band distortion cancellation can be directly linked to the structure in Fig 4-1.

Although in Fig 4-1, the amplifier A_1 ... A_n are shown as generalised transfer function, they in turn can be decomposed into local feedback/feedforward structures in a similar way to that of Fig 4-1. For example, in the comparison with the Walker amplifier discussed in Section 1, local feedback around the forward amplifier yielded an approximation to an integrator response such that when the transfer function defining impedance of Fig 2-5a, b, c were included, a close equivalance to the bridge of the current dumping amplifier was observed.

We conclude this paper by demonstrating a procedure for establishing conditions of distortion cancellation for the more general feedback/feedforward topology of Fig 4-1.

The overall transfer function G_n for the n-stage structure of Fig 4-1 is given by,

$$G_n = \frac{b_o + b_1A_1 + b_2A_1A_2 + \ldots}{1 + a_1A_1 + a_2A_1A_2 + \ldots} \qquad 4\text{-}1$$

where assuming a target transfer function G_{tn} and defining an error function E_{Gn} similar to that in equation 1-2,

$$\text{then, } E_{Gn} = \frac{G_n}{G_{tn}} - 1$$

$$\text{i.e.} \quad E_{Gn} = \frac{(b_o - G_{tn}) + (b_1 - a_1G_{tn})A_1 + (b_2 - a_2G_{tn})A_1\ A_2 + \ldots}{G_{tn}(1 + a_1A_1 + a_2A_1A_2 + \ldots)} \qquad 4\text{-}2$$

Using equation 4-2, we can now procede to invent a range of systems for which $E_{Gn} = 0$:

Example of error correction systems

(i) Transfer function independent of $A_1, A_2, \ldots A_n$.

Observation of equation 4-2 reveals a set of (n +1) balance conditions which forces $E_{Gn} = 0$ under optimum alignment such that each of the (n + 1) equation does not include the amplifier gains $A_1, A_2, \ldots A_n$.

$$\left.\begin{array}{l} \text{ie. } b_o = G_{tn} \\ \left[b_r = a_rG_{tn}\right]_{r=1}^{n} \end{array}\right] \qquad 4\text{-}3$$

The error function E_{Gn} of equation 4-2 also reveals that providing $\{A_1, A_2, \ldots A_n\} >> 1$, that significant reduction in sensitivity to balance misalignment can be attained. However, we note with caution that for $G_{tn} > 1$ then $b_o > 1$, which requires gain in the first feedforward path. Of course, we could choose the coefficients

$$\{a_r, b_r\}_{r=1}^{n} = 0$$

though for an n-stage feedback amplifier there is advantage in using several local feedback paths.

(ii) Transfer function independant of A_2, A_3 ... A_n..

The problem in example (i) of $b_o > 1$ for $G_t > 1$ can be circumvented by regrouping the left hand terms in the numerator of equation 4-2,

i.e. write:

$$E_{Gn} = \frac{A_1\left[b_1 - \left\{G_m\left(a_1 + \frac{1}{A_1}\right) - \frac{b_o}{A_1}\right\}\right] + (b_2 - a_2G_m)A_1A_2 + \ldots}{G_m(1 + a_1A_1 + a_2A_1A_2 + \ldots)}$$

whereby the balance equations become,

$$\left.\begin{array}{l} b_1 = G_m\left(a_1 + \frac{1}{A_1}\right) - \frac{b_o}{A_1} \\ \text{and}\left[b_r = a_rG_m\right]_{r=2}^{n} \end{array}\right] \quad 4\text{-}4$$

i.e. by including A_1 in the balance relationship, b_o can now be selected independently, allowing $b_o < 1$ or even $b_o = 0$ if desired.

(iii) Sandman's error-pick off distortion correction (22).

The system of error correction first attributed to Sandman (22) can be observed conceptually in the scheme of Fig 4-1 by simplifying the structure as follows:

Let,

$$b_2 = b_3 = \ldots = b_n = 0$$

$$a_2 = a_3 = \ldots = a_n = 0$$

This simplification shown in Fig 4-2 reveals a single loop negative feedback amplifier where the input error voltage is amplified by b_o and then summed with a weighted contribution from the output of amplifier A_1. The overall transfer function of the Sandman scheme, G_s, and corresponding error function, E_s, follow from equation 4-1, 4-2 as,

$$G_t = \frac{b_o + b_1 A_1}{1 + a_1 A_1} \qquad 4\text{-}5$$

$$\text{and,}\quad E_t = \frac{(b_o - G_{t1}) + (b_1 - a_1 G_{t1})A_1}{G_t(1 + a_1 A_1)} \qquad 4\text{-}6$$

where zero distortion in amplifier A_1 is achieved when

$$\left.\begin{aligned} b_o &= G_{t1} \\ b_1 &= a_1 G_{t1} \end{aligned}\right] \qquad 4\text{-}7$$

i.e. b_o determines the target gain G_{t1} and the condition for minimum distortion is set by $a_1 b_o = b_1$.

The principle disadvantage of this scheme is again the requirement of $b_o > 1$ for $G_{t1} > 1$, though since $G_{t1} < A_1$, the error amplifier b_o can be designed to exhibit a correspondingly lower distortion contribution than A_1, thus a useful performance enhancement is possible.

(iv) Current dumping error correction (1, 16)

Following the discussion of Section 1, we observe that the conceptual topology of the current dumping amplifier (1, 16) follows directly from Fig 4-1 when

$$b_o = 0,\ a_1 = 0$$
$$a_3 = a_4 = \ldots = a_n = 0$$
$$b_3 = b_4 = \ldots = b_n = 0$$

and the non-linear output cell N is represented within Fig 4-1 by the amplifier A_2. The transfer function G_w and corresponding error function E_w of the Walker current dumping topology then take the general form

$$G_w = \frac{b_1 A_1 + b_2 A_1 A_2}{1 + a_2 A_1 A_2} \qquad 4\text{-}8$$

$$E_w = \frac{(b_1 A_1 - G_{t2}) + (b_2 - a_2 G_{t2})A_1 A_2}{G_{t2}(1 + a_2 A_1 A_2)} \qquad 4\text{-}9$$

In this particular example, the target transfer function G_{t2} can be derived from the overall transfer function G_w by putting $A_2 = 1$, a condition representative of zero distortion in the unity-gain output stage forming the current dumpers.

$$\text{i.e. } G_{t2} = G_w \Bigg|_{A_2 = 1} = \frac{(b_1 + b_2)A_1}{1 + a_2 A_1} \qquad 4.10$$

hence subsituting G_{t2} in the expression for E_w, from equation 4-9, and noting with passive error summation that $b_1 + b_2 = 1$,

$$E_w = \frac{(a_2 b_1 A_1 - b_2)(1 - A_2)}{1 + a_2 A_1 A_2} \qquad 4\text{-}11$$

i.e. zero distortion results when either $A_2 = 1$ (the optimum output stage gain in this example) or, more fundamentally,

$$\frac{b_2}{b_1} = a_2 A_1 \qquad 4\text{-}12$$

which is an alternative form of the balance condition for the current dumping amplifier, that is readily observed by reference to the analysis of Section 1.

In fact, the expression in equation 4-11 for E_w, succinctly describes most of the principle attributes of the basic Walker amplifier ie.

(a) zero distortion when $A_2 \rightarrow 1$

(b) further distortion reduction even with a finite loop gain, by using error feedforward.

(c) significant reduction of error by using negative feedback where the term $(1 + a_2 A_1 A_2)$ desensitives the balance condition to misalignment for $a_2 A_1 A_2 >> 1$.

(d) a true broad-band error null capability under optimal alignment of the balance condition.

(e) $\{a_2, b_1, b_2\} < 1$ if desired allowing passive error summation, though the parameters are in practice frequency dependant (1, 8, 16).

(v) Focus function error correction

Within the multi-loop feedback/feedforward structure of Fig 4-1, error correction can in principle be targeted individually onto any of the n amplifier stages A_1 to A_n to desensitise the overall transfer function to changes in incremental gain of the selected amplifier. This ability to focus the error correction we define a focus function.

To illustrate the method of focusing error correction, consider an example of a $n = 3$ stage system where the second amplifier A_2 is selected for desensitisation. From the expression for the error function in equation 4-2, we may arrange the expression as follows:

$$E_{G3} = \frac{\left[(b_0 + b_1 A_1) - G_{t3}(1 + a_1 A_1)\right] + A_1 A_2 \left[(b_2 + b_3 A_3) - G_{t3}(a_2 + a_3 A_3)\right]}{G_{t3}(1 + a_1 A_1 + a_2 A_1 A_2 + a_3 A_1 A_2 A_3)} \qquad 4\text{-}13$$

Hence E_{G3} becomes zero and independant of A_2 when,

$$(b_0 + b_1 A_1) = G_{t3}(1 + a_1 A_1)$$

and,

$$(b_2 + b_3 A_3) = G_{t3}(a_2 + a_3 A_3)$$

This method can readily be extended to the general case of n amplifiers, where two balance equations again result by appropriate factorisation of the numerator of the error function.

Since in a practical amplifier, the summation coefficients of the feedforward paths are more readily implemented using passive components, we put forward the opinion and thus concur with Walker that schemes should attempt to selct 'b' coefficents less than unity such that

$$\sum_{r=0}^{n} b_r = 1 \qquad 4\text{-}14$$

where an appropriate method of passive implementation is shown in Fig 2-3. Hence if desired, this relationship can be considered as an extra constraint on the balance equation.

The error function described by equation 4-2, shows the advantage offered by multi-stage amplifier structures. Observation of the denominator reveals (to the right of the expression) high values of gain to be possible which greatly desensitize the error function to imbalance. Though a single high gain stage is permissible, multiple stages that distribute the gain and employ multiple feedback paths, allow greater degrees of freedom to attain stability under high loop gain (23) while simultaneous feedforward enables error correction to be focused on to selected critical stages to further enhance performance.

There is clearly an infinity of systems possible, where it is suggested that many of the proposals can be decomposed to a generalised feedforward/feedback structure as shown in Figure 4-1. This structure is therefore offered as a possible basis topology for error correction schemes, that do not depend upon dynamic gain modulation as corrective strategy (24).

5. Conclusion

This engineering report has attempted to establish a more general class of error correction scheme, where we demonstate a structure, that can be configured at a conceptual level to realise many so called new systems.

We agree with Lipshitz and Vanderkooy that feedback alone cannot achieve a global error null and that at least one feedforward path is fundamental to this aim. Feedforward, embedded within a feedback structure enables compensation for finite loop gains, consequently we need not seek infinite loop gain to achieve a theoretic zero distortion.

In particular, the comparison of the "current dumping" amplifier to an error feedback/feedforward structure was a catalyst in the establishment of the more general model, especially where a re-interpretation of the closed-loop gain reveals essentially identical transfer functions.

However, although conceptual models are fundamental to a proper understanding of a system, ultimately it is the circuit realisation, layout and component choice that are paramount. The elegance of a system is represented both by the elegance of the circuit topology and its practical execution. Thus, although "current dumping" and other distortion correction strategies can be compared, the ultimate judgement should be at a circuit level, together with performance evaluation of the total system.

We can of course present multiple mapping from system concepts to circuit schematic, where new systems are invented. However, we must also recognise that ultimately, all systems are derivatives of feedback and feedforward, where the more recent developments have correctly combined these techniques into a common framework, rather than depending on the special cases of just feedback or just feedforward.

Of course, the proposal represented by the more general structure in Figure 4-1, will be familiar to those experienced in analogue computing techniques and transient analysis (25) where such systems are commonly used. Amplifier designers have side stepped this knowledge to a degree and probably concentrated more on the circuit aspects than the underlying philosophy.

The structure of Fig 4-1 allows in principle, many families of error correction schemes to be invented, that exhibit true theoretic error nulls. It also enables a desensitization of the balance conditions to be achieved, where this is readily observable using error function modelling. In particular, we can theorise on new topologies using multiple stages and multiple feedback paths to achieve closed loop stability, yet combined with feedforward to enable a greater reduction in the non-linear dispersive error inherent in feedback only systems. Also of academic interest is the means of focusing error correction on selected stages within the cascade, as discussed in Section 4 - (v).

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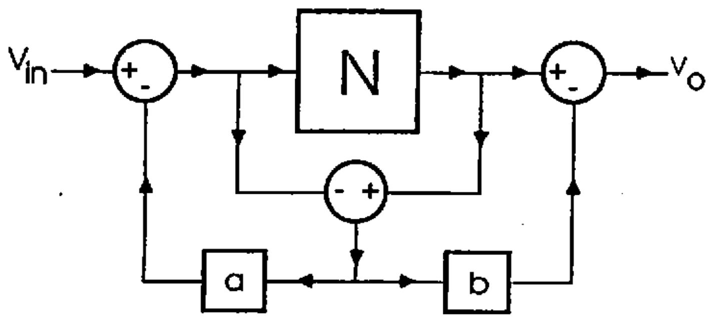


Fig. 1.1 Error feedback/feedforward distortion correction.

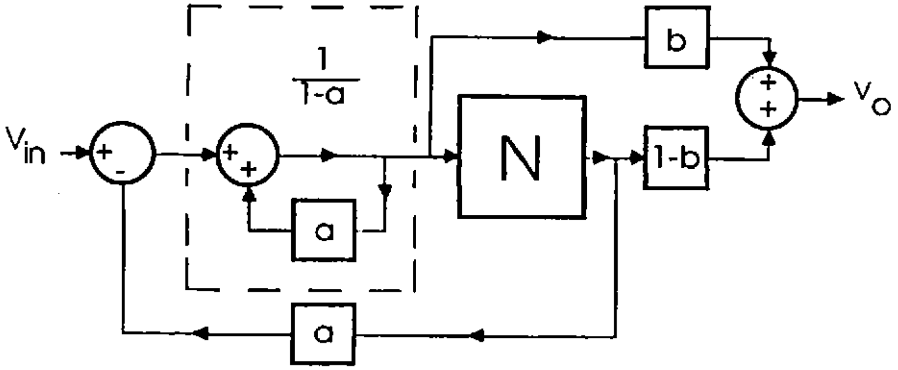


Fig . 2.1 Reconfiguration of feedback/feedforward correction scheme.

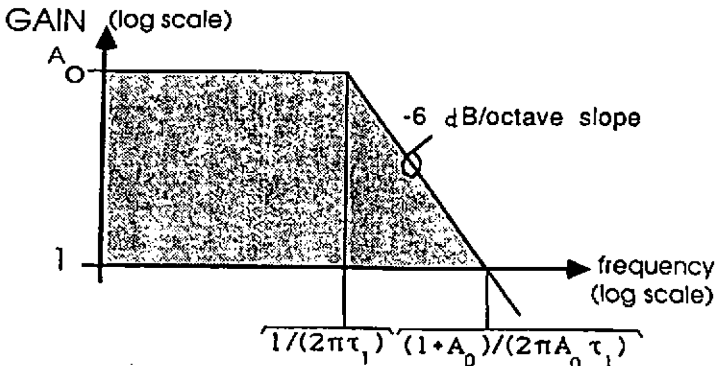


Fig. 2.2 Typical forward gain Bode plot of feedback amplifier.

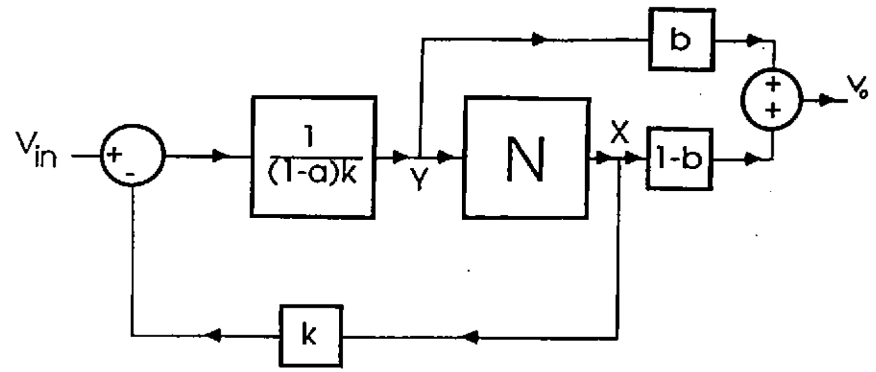


Fig. 2.4 Error correction amplifier with voltage gain.

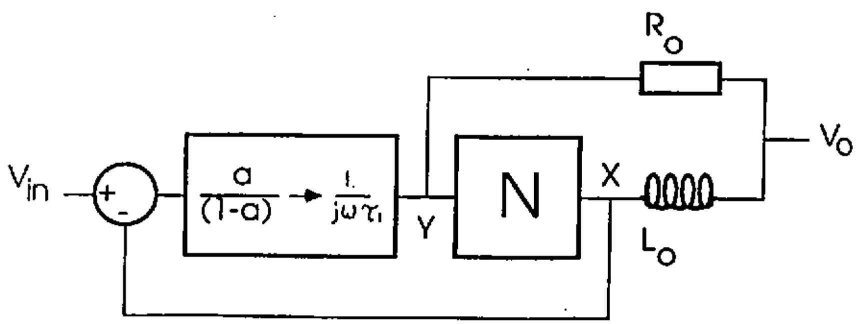


Fig. 2.3 Error summation using passive R_o, L_o network.

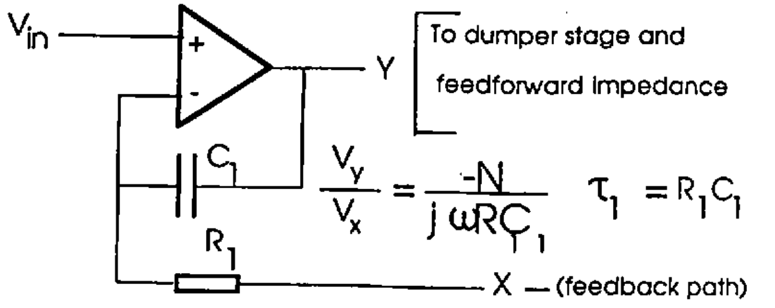


Fig. 2.5a Basic virtual-earth integrator in feedback path of current dumping amplifier.

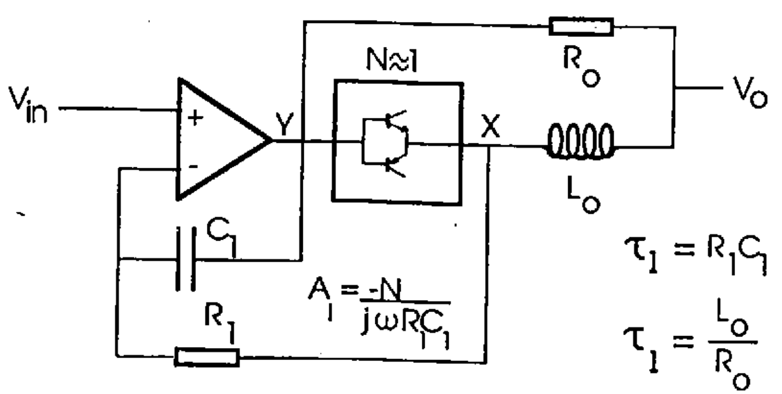
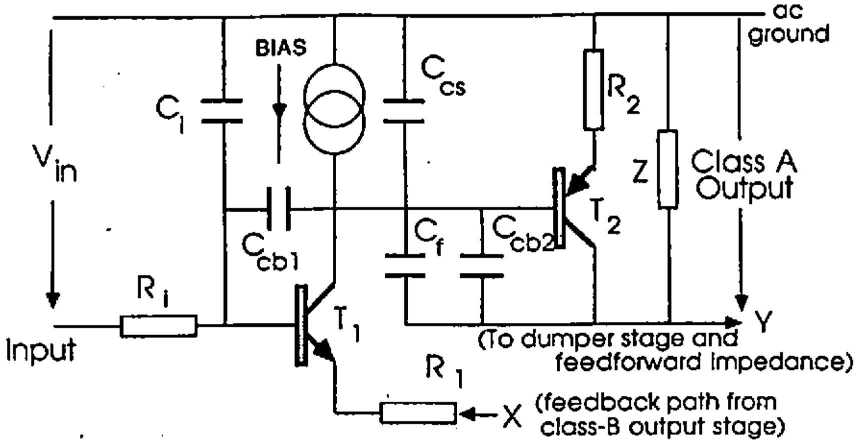


Fig 2.5b Elementary current dumping amplifier showing bridge components: R_1, C_1, loop integrator; R_0, L_0, output summation.



C_I input band limiting capacitor (in conjunction with R_i)

C_{cb1}, C_{cb2} collector-base capacitance of T_1 and T_2

C_{cs} effective capacitance of current source

C_f feedback capacitor

T_1 input transistor

T_2 composite emitter follower buffers and inverter stage

Z effective load impedance presented to the collector of T_2 due to class-B stage and biasing circuitry

Fig. 2.5c Simplified class-A amplifier topology emphasising transistor and feedback capacitors. (see ref. 1 and 16).

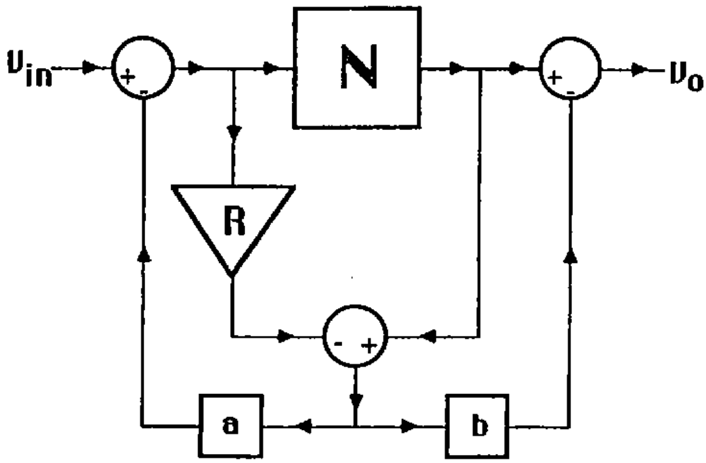


Fig. 3.1 Error feedback/feedforward with non-linear cell, exhibiting overall gain (ie. N>1).