

# Audio design leaps forward?

*Designers have long recognised the theoretical advantages of combining feed-forward error correction with feedback. But in his design for a feed-forward audio power amp Giovanni Stochino looks to have succeeded in putting theory into practice.*

Since its invention by H S Black<sup>1</sup> in the 1920s feed-forward error correction has found practical application in radio frequency and microwave amplifiers<sup>2</sup>. But it has never been used, in Black's form, in audio power amplifiers<sup>3</sup>. The reason is probably the inherent difficulty in accurately and efficiently applying Black's feed-forward principle to audio power amplifiers over the full audio frequency range.

But a newly-developed circuit technique could do just that, and, within specified limits, put Black's true feed-forward principle to work in high power audio amplifiers.

Experimental results demonstrate the effectiveness of the proposed technique, but first, a look at some of the underlying theory.

### Feed-forward or feedback?

The general input-output relationship of a power amplifier, before applying correction, can be written as  $V_p = V_i G_p + E_p$ .  $G_p$  is the volt-

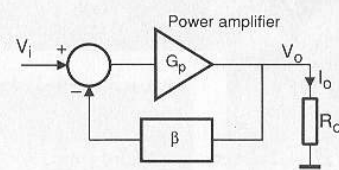
age gain, generally a function of frequency and load impedance, and  $E_p$  is the error component that includes the amplifier's non-linear distortion and noise.  $E_p$  depends on input voltage and load impedance, and on frequency.

When negative feedback is applied (Fig. 1), the input-output relationship of the corrected amplifier becomes  $V_o = A_{cl} V_i + E_{fb}$ .  $A_{cl}$  is the closed loop voltage gain, substantially defined by the feedback network, and  $E_{fb}$  is the residual error component after feedback correction.

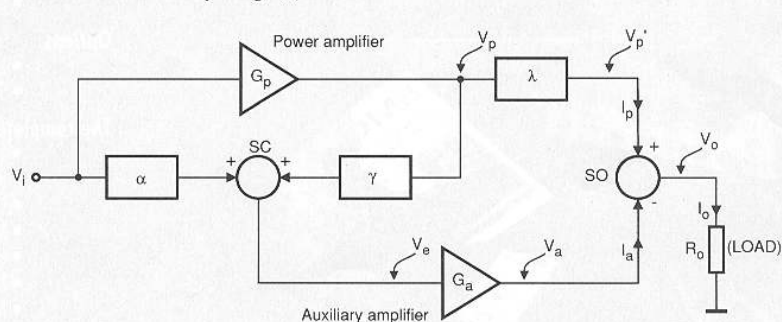
Analysis shows that distortion can never be completely nulled by negative feedback – though feedback is effective in reducing distortion as long as there is enough gain within the feedback loop.

Feed-forward is based on a different mechanism of error correction. The basic scheme (Fig. 2) incorporates a criterion network ( $\alpha$ ,  $\gamma$  and SC) to determine, isolate and extract power amplifier error; an auxiliary amplifier AA (low power requirement, low distortion

*Fig. 1. Basic elements of a feedback amplifier. Analysis shows that distortion can never be completely nulled with this configuration.*



*Fig. 2. Principles behind feed-forward. In this configuration, power amplifier error is extracted, determined and isolated via a criterion network comprising  $\alpha$ ,  $\gamma$  and SC.*



and low noise compared with the power amplifier, PA) to provide a buffered copy of  $E_p$ ; and output summing network SO. In SO the error component of PA and its copy available at the output of AA cancel out to provide a distortion-free output voltage on load  $R_o$ . Phase-amplitude equaliser network  $\lambda$  is added to the basic scheme to improve the error-correction mechanism at high frequency.

The scheme should include a few delay lines to compensate for amplifier propagation delay and connections. But their influence is negligible in the audio frequency range.

Simple analysis of the diagram gives:

$$V_0 = V_p' - V_a$$

$$= V_i G_p \lambda - G_a V_i (\alpha + \gamma G_p) + E_p (\lambda - \gamma G_a) + E_a$$

where  $E_a$  is the error component (distortion plus noise) produced by the auxiliary amplifier. Proper operation of the feed-forward technique requires that  $E_a \ll E_p$  so the effective output error is  $E_{ff} = E_a + E_p/\rho$ . Term  $\rho = 1/(\lambda - \gamma G_a)$  can be defined as the distortion rejection factor of the feed-forward amplifier and describes the effectiveness of feed-forward in removing distortion in the power amplifier.  $E_{ff}$  reduces to its lowest value of  $E_a$  when  $\rho = \infty$ , that is when  $\gamma G_a = \lambda$ , and shows the potential of the feed-forward mechanism to completely null distortion  $E_p$  in the power amplifier.

The further condition  $\gamma G_p = \alpha$  should be satisfied to nullify the component  $V_e^* = V_e(V_i)$  (see panel p. 822 for definition of  $V_e^*$ ) at the input of the auxiliary amplifier. This would minimise both  $E_a$  and power handling requirements for the auxiliary amplifier. The mathematics implies that when  $\gamma > 0$ ,  $G_p$  and  $\alpha$  have opposite signs.

**Feed-forward more promising?**

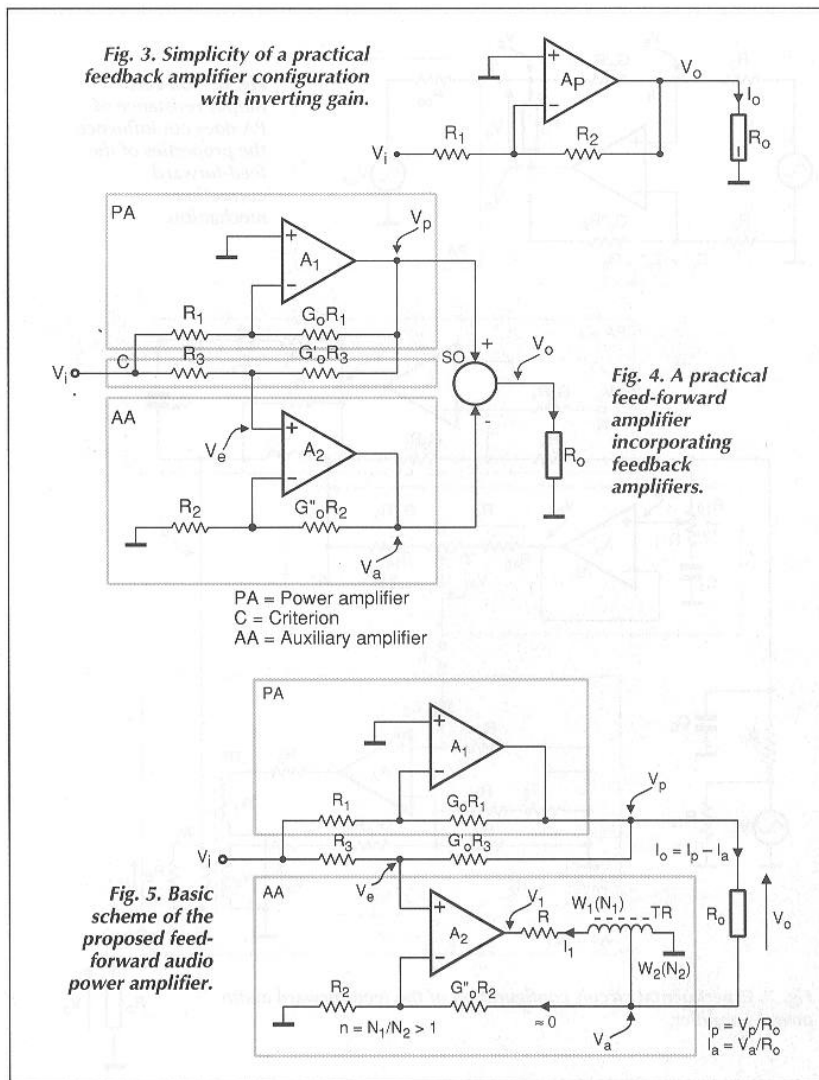
Distortion  $E_{fb}$  of a feedback amplifier can never be nulled, but it can be substantially reduced in the range of frequency and input voltage, where the feedback factor is much greater than 1.

As a technique, it is less effective at the highest frequency of the audio range and in the crossover region of class AB amplifiers, where the feedback factor can be low and deviation from linearity is high<sup>4</sup>.

On the other hand, negative feedback amplifier configurations are very simple and require no matching of components (Fig. 3).

Feed-forward error correction is much more complex. But better distortion results are possible. In theory, the error of the whole power amplifier can be reduced to that of the auxiliary amplifier alone, even at high frequencies and in the crossover region. The advantage is that the auxiliary amplifier needs to handle only moderate currents and voltages. So it can be designed to provide much lower distortion (for instance it can be operated in class A) than the power amplifier, and very low distortion can be achieved.

Neither feedback nor feed-forward error correction can completely null the output error of a power amplifier. But feed-forward is more promising, virtually nulling distortion of the



power amplifier, leaving only the low residual error of the auxiliary amplifier over the load.

**Combining feedback and feed-forward**

Tight matching of parameters in the feed-forward scheme (Fig. 2) can be achieved, simply and steadily, by using negative feedback. The strategy helps precise definition of gain in both the power amplifier and the auxiliary amplifier – provided the open loop gain of both amplifiers is high in the audio frequency range. So feedback and feed-forward techniques can be profitably combined in a true low-distortion audio power amplifiers. In a practical application (Fig. 4), the power amplifier and auxiliary amplifier have the gains defined by their respective feedback networks:  $G_p = V_p/V_i \approx -G_0$  and  $G_a = V_a/V_i \approx (1 + G_0')$  provided  $A_1/G_0 \gg 1$  and  $A_2/G_0' \gg 1$ . But there is also  $\gamma = 1/(1 + G_0')$ ,  $\lambda = 1$  and  $\alpha = G_0'/(1 + G_0')$ . As a result,  $G_0' = G_0$  and  $G_0' = G_0$ .

The scheme is a practical way of assuring that the fundamental conditions for proper operation of feed-forward technique are always satisfied. But the problem remains in

implementing the output summing network – probably the most difficult obstacle in the basic feed-forward error correction scheme.

The simplest and most intuitive way of realising this summing network is where corrective voltage  $V_a$  is directly transferred into the load's loop. Voltage  $V_0$  across the load is equal to  $V_p - V_a = -G_0 V_i + E_p - E_p + E_a = -G_0 V_i + E_a$ , which is consistent with feed-forward theory.

But if, in this scheme, the auxiliary amplifier has to sustain the full load current, the assumption that the auxiliary amplifier is a low-power, low-distortion (prospectively class A) amplifier is no longer valid.

As a result, we can not assume that  $E_a \ll E_p$ , and consequently the inherent advantage of the feed-forward technique disappears. This is why the simple feed-forward configuration has never been used in power amplifiers<sup>3</sup>.

It also explains why, though the advantages of the feed-forward technique, in conjunction with feedback, are generally recognised, Black's feed-forward error correction technique has found only limited application by audio designers.

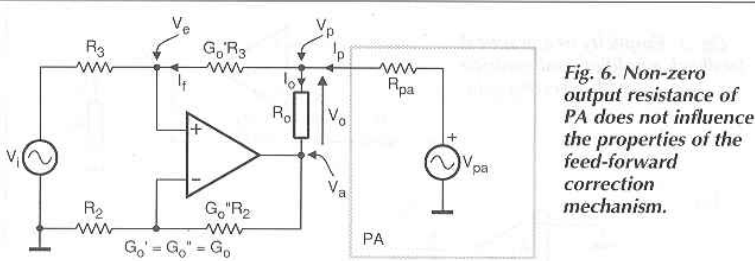


Fig. 6. Non-zero output resistance of PA does not influence the properties of the feed-forward correction mechanism.

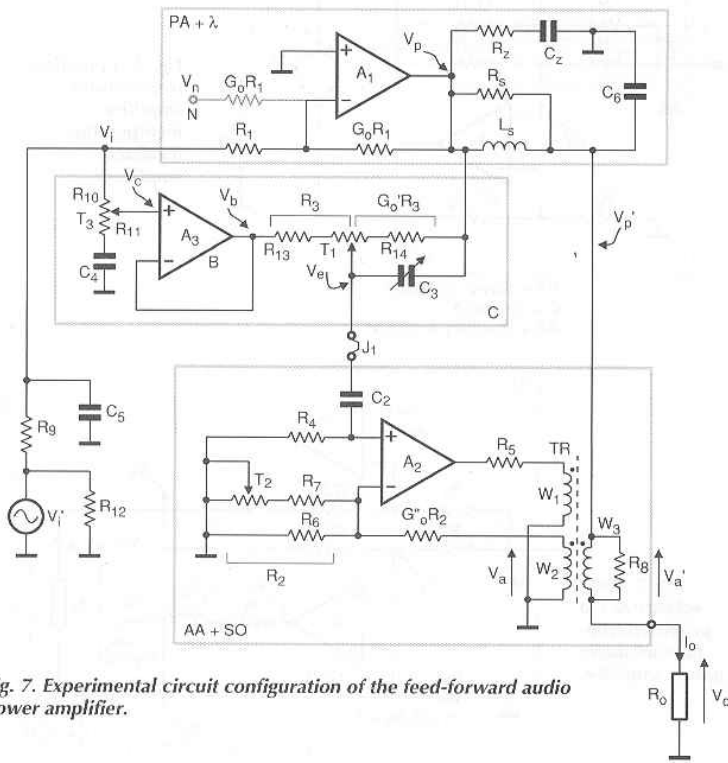


Fig. 7. Experimental circuit configuration of the feed-forward audio power amplifier.

Feed-forward error correction (always intra-loop) is sometimes used in audio power amplifiers<sup>5, 6</sup>, but Black's basic scheme has yet to be incorporated into audio power amplifier design.

**A feed-forward power amplifier that works**

We have seen that, in the feed-forward scheme (Figs. 2, 4), the most critical part to be implemented in audio applications is the output summing network (SO). Here the power signal  $V_p$  coming from the power amplifier and low level corrective signal  $V_a$  produced by the auxiliary amplifier have to combine without undesired interaction (ie cross-modulation, frequency instability, gain impairment) to provide a distortion-free output.

What is more, this combination must be performed efficiently without requiring much power from the auxiliary amplifier, and must not be affected by impedance-variations of loads – even loads as difficult as loudspeakers.

In amplifiers for hf use, such problems are less critical. Appropriate networks can be used to implement the output summing function,

due mainly to the favourable frequency range and fixed system impedance (50Ω).

But audio applications span an unfavourable frequency range and imply complex and unpredictable load impedances. As a result, circuit techniques commonly used in radio frequency and microwave feed-forward power amplifiers are not practical, and different solutions have to be found.

An effective approach (Fig. 5<sup>7</sup>) has PA as the power amplifier to be corrected (usually class AB), and AA as the auxiliary amplifier. AA should be operated in class A for the best performance and incorporates transformer TR in its feedback loop. (Resistor R also includes the resistance of winding  $W_1$  and the output resistance of  $A_2$ .)

The unique role of TR is to provide both the wide-band impedance matching of the auxiliary amplifier to  $R_0$  and the power-efficient means for injecting the corrective signal  $V_a$  into the load's loop.

Transformers are usually avoided in solid state audio power amplifiers, as they are expensive, bulky, band-limited and not suited for very low distortion applications. But when

used in unconventional ways, as in this case, their unique properties can prove useful.

Putting transformer TR in the feedback loop of the auxiliary amplifier has two very important effects. The flux produced in the magnetic core of TR by the power component of the load current is automatically annulled by the feedback that forces voltage  $V_a$  to be insensitive to power component variation. So no restrictions are imposed on transformer size and core material by the amount of power that the power amplifier transfers into the load. In most cases a small transformer can be used.

Open loop output impedance of the auxiliary amplifier can also be extremely low (a few mΩ) in the full audio frequency range and above. The consequence is that undesired interactions and cross-modulations between power amplifier and auxiliary amplifier, as well as the sensitivity of the auxiliary amplifier to load impedance variations, are strongly reduced. A further benefit is that the primary winding of TR is driven, virtually, by a voltage source, since R tends to zero. This widens the frequency bandwidth of TR, whose practical low frequency corner  $f_0$  turns out to be as low as a few Hz, even if a small ferrite core is used to improve its bandwidth and linearity.

**Transformer operational requirements**

The function of the transformer – to permit injection of the corrective current into the load without interaction with the main current component – is performed by cancelling the core flux generated by the main current component.

This flux neutralisation is carried out by the coercive action of the auxiliary amplifier's feedback loop and is effective as long as the current and voltage available at the output of  $A_2$  are adequate and the loop gain remains high. The only effective flux in the transformer core is therefore produced by the corrective voltage  $V_a$ .

For frequency  $f > f_0$ , the peak flux density  $B_p$  and the peak voltage  $V_{ap}$  are linked by  $2\pi f B_p S_c N_2$  where  $S_c$  is the effective cross-sectional area of the transformer core. So the amount of corrective voltage  $V_{ap}$  that can be provided to the load is limited by the core geometry, through  $S_c$ , and the core material, through  $B_s$  (ie the saturation flux density), since it must always be  $B_p \leq B_s$ . The amount of available corrective voltage can also be seen to increase in direct proportion to frequency.

As an example, take a toroidal ferrite core with  $S_c$  at 100mm<sup>2</sup> and  $B_s$  at 200mT. If  $N_2$  is 20,  $V_{ap}$  is 50mV at 20Hz and  $V_{ap}$  is 5V at 2kHz. Compared to an output of 100W/8Ω, they represent peak correctable errors of 0.12% and 8.7% respectively.

Performance matches well with that of class AB audio power amplifiers, exhibiting non-linear distortion that rises with increasing frequency, and extends normally, say, from 0.01% to 1% in the audio frequency range.

**Amplifier requirements**

Class A operation is mandatory for amplifier  $A_2$  to achieve the lowest distortion with low level error signals. High gain and low noise

are also recommended, as is low output offset voltage to stop any noticeable direct current flowing into the transformer winding.

Other requirements are wide bandwidth and high slew rate so that the correction capability of the auxiliary amplifier is extended to the highest harmonics of audio signals.

Output voltage and current handling capabilities of the auxiliary amplifier are dictated by the peak values of load current and corrective voltage:  $V_1 \cong nV_{ap}$  and  $I_1 \cong I_{op}/n$  where  $n=N_1/N_2$  is the turns ratio and  $is \gg 1$ . The turns ratio is used to trade-off voltage for current to reach the best level of performance of  $A_2$  with a reasonable power consumption. For example,  $I_{op}=8A$ ,  $V_{ap}=0.5V$  and  $n=40$  gives  $V_1=20V$  and  $I_1=0.2A$ . Therefore,  $A_2$  can be powered from  $\pm 25V$  and its output stage biased at 0.2A for class A operation.

Only 10W of power is consumed by the auxiliary amplifier – a reasonable and worthy amount if compared with the 256W of undistorted audio power furnished to an 8Ω load.

**Auxiliary considerations**

An important, yet often overlooked, characteristic of feed-forward schemes based upon Black's principle, is that the voltage across the load is not defined by the power amplifier, but by the auxiliary amplifier only. In other words  $V_0$  is not theoretically dependent on power amplifier parameters (output impedance, gain, linearity etc). Power amplifier output could even be completely uncorrelated with  $V_1$ , and power amplifier output impedance could be high and non-linear without affecting the output voltage value ( $V_0 = -G_0V_1$  in Fig. 5).

We can also deduce, mathematically, that  $V_a = V_1G_0 + V_p$  and  $V_0 = V_p - V_1G_0 - V_p = -V_1G_0$  to show that output voltage is always equal to the desired value, regardless of the power amplifier output voltage  $V_p$ . So any deviation of  $V_p$  from its ideal value affects the output of the auxiliary amplifier but not output voltage  $V_0$ .

We reach the same conclusion if we take into account the non-zero output impedance of the power amplifier Fig. 6 – particularly important in the crossover and clipping regions, where comparatively high values of output impedance can be experienced.

Assume  $G_0R_3 \gg R_{pa}$ . In this case we have  $I_f \ll I_p$  so  $I_p \cong I_0$  and the node voltages are:

$$V_p = (V_{pa} - V_a) / (R_0 + R_{pa}) + V_a,$$

and,

$$V_a = V_{pa} + G_0V_1.$$

Solving simultaneously, substituting and assuming in normal operation that  $V_{pa} = -G_0V_1 + E_p$ , we have:

$$V_p = -G_0V_1(1 - R_{pa}/R_0) + E_p$$

and

$$V_a = G_0V_1R_{pa}/R_0 + E_p$$

You can see that the auxiliary amplifier has to contribute a certain amount of signal voltage to the load in addition to the copy of the error voltage  $E_p$ . This amount is proportional to the ratio  $R_{pa}/R_0$ .

Power amplifiers with poorly biased class AB output stages can actually have closed loop-output impedances comparable with load impedance, at least in the crossover region. In such cases, the auxiliary amplifier contribution to the output signal voltage in the crossover region can prove significant.

Clearly, the role of forcing the desired voltage across the load is undertaken by the auxiliary amplifier.

In well-designed feed-forward amplifiers, the power amplifier provides the power to the load, while the auxiliary amplifier is limited to providing accuracy and precision only. Nevertheless, should the power amplifier fail to do its job (for instance due to crossover mechanism), the auxiliary amplifier would be forced to provide power as well as precision. Obviously, the auxiliary amplifier is designed to provide only a limited amount of precise corrective voltage.

It is worth noting that the auxiliary amplifier is always stable because even with the worst-case positive feedback factor ( $R_{pa} = \infty$ ),

$$F_p = R_3 / [R_3(1 + G_0') + R_0]$$

of the auxiliary amplifier.

Extremely low open loop output impedance means the auxiliary amplifier's closed loop gain and distortion performance are insensitive to load impedance variations.

Wide bandwidth achievable for the low power auxiliary amplifier also allows a large reduction of the highest harmonics of audio signals.

Error correction technique (Fig. 5) can be applied to all power amplifier circuit configurations, with inverting as well as non-inverting gain.

**Magnetic flux cancellation**

This diagram helps analyse the mechanism of magnetic flux cancellation  $\Phi(V_1)$  in the transformer core, due to main signal current  $I_p \cong V_p/R_0 \cong -C_0V_1/R_0$ . It is operated by the auxiliary amplifier.

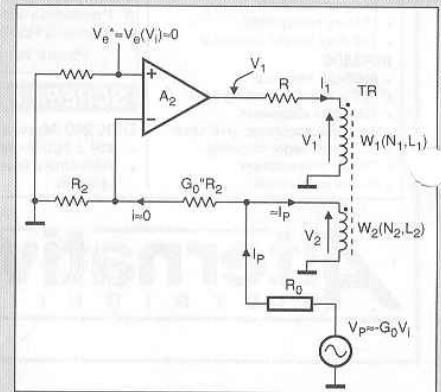
$$\Phi(V_1) \cong V_2(V_1) / (2\pi f N_2)$$

$$V_2(V_1) \cong -G_0V_1 \frac{R}{n^2R_0} \cdot \frac{1}{[1 + A_2 / \{n(1 + G_0'')\}]}$$

From this,

$$\lim V_2(V_1) = \Phi(V_1) = 0$$

$$(A_2 \rightarrow \infty \text{ and/or } R \rightarrow 0)$$



is always lower than the negative feedback factor  $F_n = R_2 / [R_2(1 + G_0'')]$  when  $G_0' = G_0''$ .

**Practical circuit**

In a practical circuit implementation (Fig. 7), the first-order error due to the finite gain-bandwidth product of the amplifiers can be taken into account and compensated for.

Transformer TR is modified to provide error correction to a grounded load and its secondary windings  $W_2$  and  $W_3$  are close-coupled to assure that  $V_a' = V_a$ . Follower B buffers the input voltage source and the phase-amplitude-equaliser ( $R_{10}$ ,  $R_{11}$  and  $C_4$ ) from the criterion network.  $R_6$  and  $C_5$  form an input low-pass filter cutting off input frequencies above 100kHz. Decoupling  $C_2$  avoids undesired dc operation with the auxiliary amplifier. Trimmers  $T_1$ ,  $T_2$  and  $T_3$  facilitate calibration of the complete amplifier and help achieve the best distortion performance.

In analysing the circuit, the effects of  $C_2$  on the output voltage can be neglected, with  $C_2$  assumed to be  $\infty$ .

In the Zobel network ( $R_2$ ,  $C_2$ ,  $R_8$  and  $L_8$ ) commonly used at the output of class AB audio power amplifiers,  $R_8$  and  $L_8$  in conjunction with  $C_6$  implement the  $\lambda$  amplitude-phase equalisation network (Fig. 2).

As well as its normal role of separating the power amplifier from the load at frequencies far above the audio range, this network also limits positive feedback around the auxiliary amplifier at the highest frequencies, improving

**Turning Black's feed-forward principle into practice**

Load impedance 'seen' by amplifier  $A_2$  is high enough to allow true low distortion operation of the auxiliary amplifier.

Auxiliary amplifier has to process only small error components, and being class A operated, its percent error contribution to load current is extremely low.

There is no appreciable common-mode-induced distortion because component  $V_e^* = V_e(V_1)$  is virtually zero.

Transformer distortion, if any, is reduced in proportion to the loop gain

of the auxiliary amplifier.

Extremely low open loop output impedance means the auxiliary amplifier's closed loop gain and distortion performance are insensitive to load impedance variations.

Wide bandwidth achievable for the low power auxiliary amplifier also allows a large reduction of the highest harmonics of audio signals.

Error correction technique (Fig. 5) can be applied to all power amplifier circuit configurations, with inverting as well as non-inverting gain.

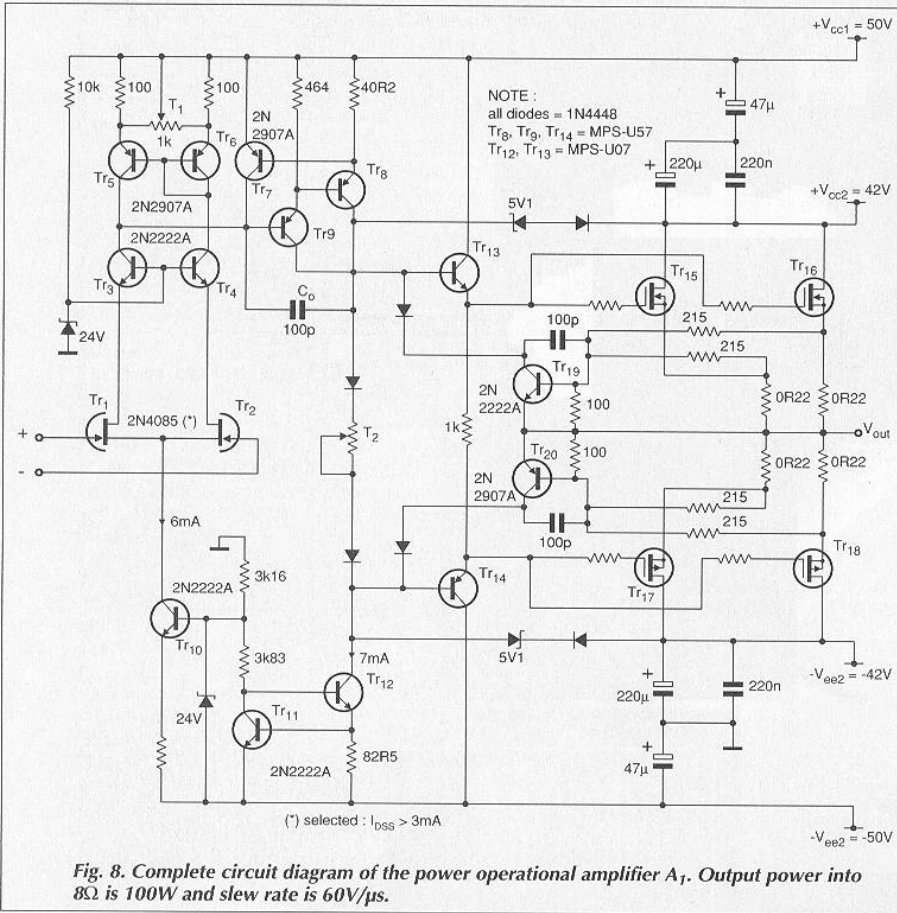


Fig. 8. Complete circuit diagram of the power operational amplifier A1. Output power into 8Ω is 100W and slew rate is 60V/μs.

frequency stability of the whole amplifier.

Resistor  $R_8$  controls the high frequency output impedance of AA. Trimmers  $T_1$  and  $T_2$  permit conditions to be satisfied.  $C_3$  compensates the frequency response of the auxiliary amplifier, and the input capacitance of  $A_2$ .

In the analysis, a single pole frequency response is taken for all amplifiers. So, in addition to condition  $G_0' = G_0'' = G_0$ , we shall assume  $V_p = [-V_i G_0 / (1 + s/p_1)] + E_p$ . Also,

$$V_b = V_c / (1 + s/p_3) \approx V_c = V_i (1 + s/z_0) / (1 + s/p_0),$$

where  $p_0$  is  $1/[C_4(R + R_{11})]$ ,  $z_0$  is  $1/C_4 R_{11}$  and  $V_a$  is  $V_c [(1 + G_0) / (1 + s/p_2)] + E_a$ .

In general we have  $p_3 \gg p_1$  and  $p_3 \gg p_2$  so that the assumption  $V_b \approx V_c$  has no appreciable consequences.

As for  $\lambda$  - assuming a first order response of the auxiliary amplifier - we expect a first order low-pass frequency response so that  $\lambda \approx 1/(1 + s/p_5)$  where  $p_5 = 1/(C_6 R_6)$ , corresponding to the assumption that  $R_5 \ll sL_s$ .

Regarding the value of the distortion rejection factor, the validity of the above expressions for  $\lambda$  and  $p_5$  is substantially independent on the load impedance value, since  $V_a'$  is applied in series with  $R_0$ . Then, according to Black's scheme in Fig. 2, applied to our circuit (Fig. 7), the error voltage  $V_e$  can be written as:

$$V_e = V_i (\alpha + \gamma G_p) + \gamma E_p$$

where

$$\alpha = [(1 + s/z_0) / (1 + s/p_0)] \cdot [G_0 / (1 + G_0) (1 + s/p_4)],$$

$$\gamma = (1 + s/z_3) / [(1 + G_0) (1 + s/p_4)],$$

$$\gamma G_p = -G_0 (1 + s/z_3) / [(1 + G_0) (1 + s/p_1) (1 + s/p_4)],$$

$$z_3 = 1 / (G_0 R_3 C_3)$$

and

$$p_4 = (1 + G_0) z_3.$$

The condition that  $\gamma G_p = -a$  can be met if  $p_0 = p_1$  and  $z_0 = z_3$  so that the error  $V_e$  reduces to  $\gamma E_p$ .

We also want to satisfy the condition  $\gamma G_a = \lambda$ . Substituting for each term reveals that this condition is true if  $z_3 = p_2$  and  $p_5 = p_4$  so that  $\rho = 1/(\lambda - \gamma G_a) = \infty$ . Therefore, the power amplifier error  $E_p$  turns out to have been completely removed from the output voltage and the above can be written as:

$$p_5 = p_4 = (1 + G_0) z_3 = (1 + G_0) p_2 \approx 2\pi f_{T2},$$

where  $f_{T2}$  is the nominal gain-bandwidth product of the auxiliary amplifier.

The interpretation of making  $z_3$  equal to  $p_2$  and  $p_5 = p_4$  is that zero  $z_3$  is introduced to compensate for the first-order phase-amplitude errors caused by the pole  $p_2$  of the auxiliary

amplifier, while the low residual errors due to the collateral pole  $p_4$ , associated with block of the criterion network, are counteracted by means of the high frequency pole  $p_5$  of the phase-amplitude equalising network  $\lambda$ .

In other words we can state that all potential limitations of Black's feed-forward error-correction mechanism, due to the dominant pole of both the power and the auxiliary amplifier, have been actually counterbalanced in this practical implementation.

The more accurate expression of  $\lambda$ , written as  $\lambda = (1 + s/z_s) / (1 + s/z_s + s^2/\omega_0^2)$  where  $z_s$  is  $R_s/L_s$  and  $\omega_0^2$  is  $1/L_s C_6$  shows the additional potential of the  $\lambda$  network to compensate for a more realistic second-order frequency response of the auxiliary amplifier, by suitable choice of  $z_s$  and  $\omega_0$ .

This accounts for the high distortion rejection factor (30 to 60dB for frequencies up to 1MHz) that has been measured after calibration on the prototypes (see 'Measurement results') with different load conditions. The end result is that very low distortion figures can be expected - and attained.

An additional property of the feed-forward implementation depicted in Fig. 7, with its floating winding ( $W_3$ ) able to inject the corrective voltage  $V_a$  into the load's loop, is that it easily lends itself to iterative application, reducing output error to extremely low levels.

### Power op-amp A1 in practice

The viability of the error-correction technique discussed so far has been demonstrated by prototypes of an 100W/8Ω audio power amplifier (Figs. 8, 9 and 10), assembled and calibrated according to Fig. 7 using the theory analysed above.

Power op amp output stage (Fig. 8) includes

#### Main characteristics of A1

Output power into 8Ω	100W
Output power into 4Ω	160W
Slew rate	±60V/μs
Power bandwidth	≈200kHz
Gain-bandwidth product (measured at 1MHz)	≈11MHz

two pairs of complementary n- and p-channel power mosfets whose quiescent current can be adjusted with trimmer  $T_2$ . Different supply rails are used to improve amplifier efficiency.

Output offset voltage can be adjusted with trimmer  $T_1$ .

### Auxiliary op-amp A2

Amplifier  $A_2$ , Fig. 9, has a mosfet output

#### Main characteristics - amplifier A2

Voltage gain	≥10 <sup>8</sup>
Output voltage range	≈±20V
Output current range	
Class A	≈±200mA
Class AB	≈±800mA
Slew rate	≈±500V/μs
Gain-bandwidth product (measured at 1MHz)	≈300MHz
Output offset voltage	≈±600μV

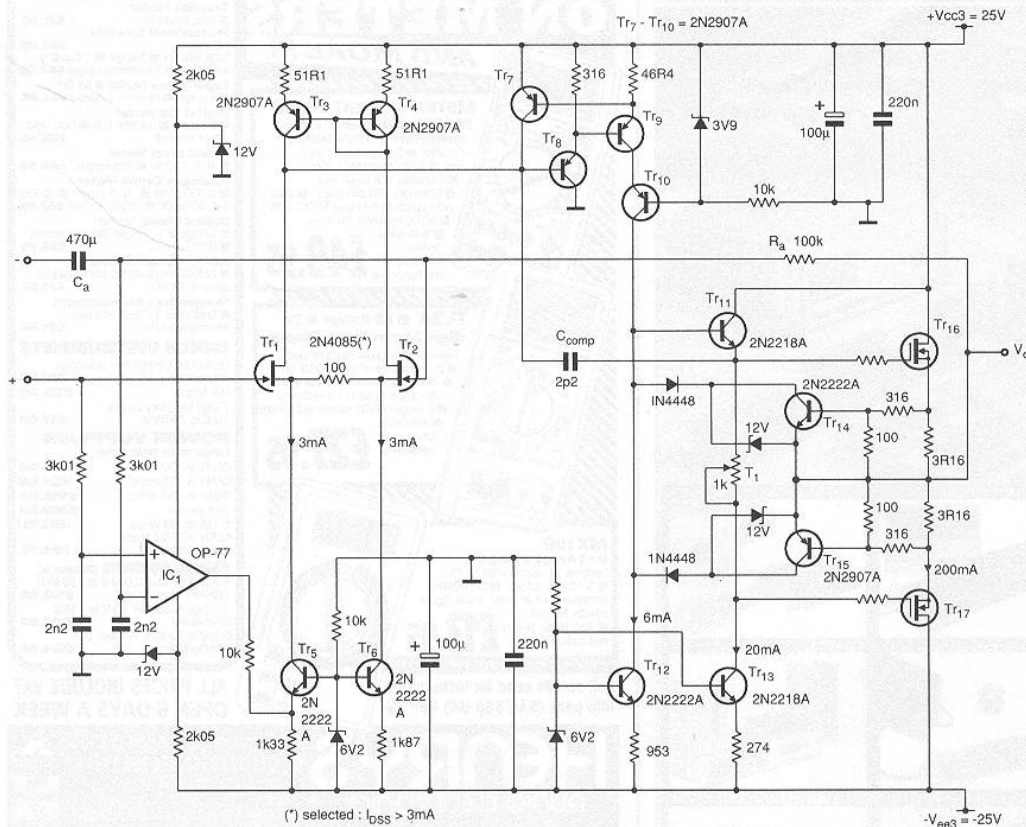


Fig. 9. Power output requirement for the auxiliary operational amplifier,  $A_2$ , is lower. In class A, output is  $\pm 200\text{mA}$  while in class AB it rises to  $\pm 800\text{mA}$ .

stage, biased by adjusting trimmer  $T_1$  at  $200\text{mA}$  for class A operation in normal working conditions. But it can operate in class AB operation, when  $A_2$  is forced to sink or source higher currents, due to variations in load impedance or clipping for example.

The circuit is a combination of a high-speed, high dynamic-range amplifiers ( $Tr_{1-17}$ ) and a precision integrated op-amp  $IC_1$ . The main task of  $IC_1$  is, with the help of coupling capacitor  $C_a$  and feedback resistor  $R_a$ , to keep the offset voltage below a few hundred millivolts and to increase the low frequency open loop gain of the overall amplifier.

$IC_1$  also helps reduce the low-frequency voltage noise ( $1/f$  noise) associated with jfet pair  $Tr_{1,2}$ .

**Main characteristics – buffer**

- TDH with  $1V/600\Omega$
- from 20Hz to 20kHz  $\leq 0.0005$
- 3dB small-signal bandwidth  $\approx 15\text{MHz}$
- Voltage noise density  $\approx 1\text{nV}/\sqrt{\text{Hz}}$
- Slew rate  $\approx \pm 20V/\mu\text{s}$

**Buffering**

Buffer  $B$  of Fig. 10 consists of an op-amp voltage follower and makes use of a high-performance integrated operational amplifier, featuring low noise and very low distortion.

**Transformer**

Transformer  $TR$ 's core is a small toroid – 23mm external diameter, 14mm internal diam-

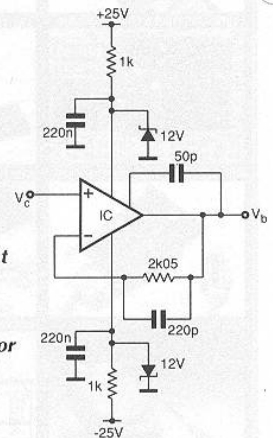
**Complete auxiliary amplifier and transformer characteristics**

Voltage gain	$\approx 18.1$
Gain-bandwidth product	$\approx 10\text{MHz}$
Slew rate	$\approx \pm 17V/\mu\text{s}$
Thd+noise	
@5kHz	$\approx 0.01\%$
Thd, $V_a=0.5V/1\Omega$	
@50kHz	$\approx 0.05\%$

eter and 7mm high. Its cross-sectional area  $S_c$  is approximately  $31\text{mm}^2$ . The core material is Ferroxcube-grade 3E2, having a saturation flux density  $B_s$  of about  $350\text{mT}$ , and a useful linear range of  $\pm 200\text{mT}$ . Turns ratio  $n$  is 30 ( $N_1=300, N_2=N_3=10$ ).

Secondary windings  $W_2$  and  $W_3$  are close-coupled with parallel and cross-coupled thick

Fig. 10. Buffer  $B$ . At frequencies from 20Hz to 20kHz, distortion of this design is less than or equal to 0.0005%.



wires and all windings are uniformly wound along the core length. When driving the primary winding with a source resistance of  $5\Omega$ , the -3dB small signal bandwidth of the transformer extends from 5Hz to about 13MHz.

**Amplifier calibration**

Amplifier calibration has been performed with a load of  $8\Omega$  in parallel with  $0.2\mu\text{F}$  using the following procedure.

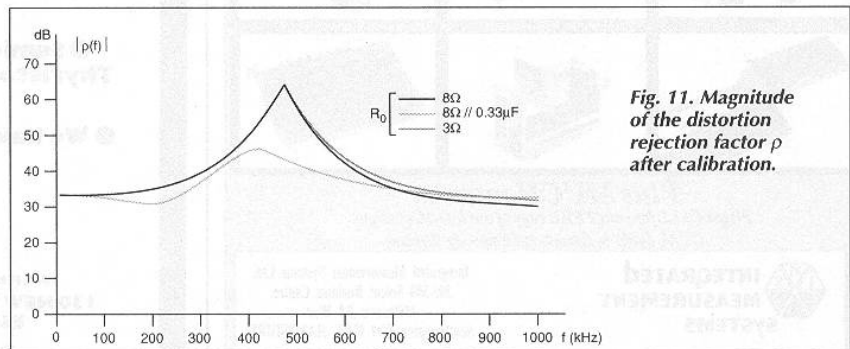


Fig. 11. Magnitude of the distortion rejection factor  $\rho$  after calibration.

Step 1: Jumper  $J_1$  is opened to isolate AA.

Step 2: Corner frequency of both PA and AA is measured and recorded ( $G_0'' \cong G_0 \cong 18.1$ ).

$$f_{c1} = p_1 / 2\pi \cong 422 \text{ kHz}$$

$$f_{c2} = p_2 / 2\pi \cong 650 \text{ kHz}$$

Step 3: Nominal value of  $C_3$  is determined with  $z_3 = p_2$ :

$$C_3 = 1 / (G_0 R_3 p_2) \cong 13.5 \text{ pF}$$

Step 4: Nominal values of  $C_4$ ,  $R_{10}$  and  $R_{11}$  are found by applying  $p_0 = 1 / C_4 (R_{10} + R_{11})$  and  $z_0 = 1 / C_4 R_{11}$  and using  $p_0 = p_1$  and  $z_0 = z_3$ . Since  $R_{10} + R_{11} = R_{T3} = 2 \text{ k}\Omega$  we obtain:

$$C_4 = 1 / [(R_{10} + R_{11}) p_1] \cong 188 \text{ pF}$$

$$R_{11} = 1 / C_4 z_3 = 1 / C_4 p_2 \cong 1.3 \text{ k}\Omega$$

Step 5: The aim is to meet the condition defined by  $\gamma G_a = \lambda$ . Signal  $V_i' = 100 \text{ mV} / 3 \text{ kHz}$  is applied to the input and trimmer  $T_1$  is adjusted so that  $V_e^* = V_e(V_i)$  reaches a minimum. Then the frequency is increased to 100 kHz and trimmer  $T_3$  is adjusted so that  $V_e^* = V_e(V_i)$  is again at a minimum.

Step 6: Connect jumper  $J_1$  and repeat step 5.

Step 7: Input of the amplifier is grounded and a forced error signal  $E_{pn}$  is produced at the output of PA by applying the input voltage  $V_n \cong 50 \text{ mV}$  (the amplitude of  $V_n$  must be kept below the limits set by  $V_{am} = 2\pi f B_{pm} S_e N_2$ , as shown in Fig. 7. Since  $E_{pn} = E_n$ , this method maximises, in a wide frequency range (up to 1 MHz), the distortion rejection factor  $\rho = 1 / (\lambda - \gamma G_p)$  of the auxiliary amplifier.

Frequency of  $V_n$  is first set at 3 kHz and trimmer  $T_2$  is adjusted so that the output voltage  $V_0(V_n)$  is at a minimum. Then, the frequency is increased to 300 kHz and  $C_3$  is adjusted again to have maximum rejection. A network analyser would simplify amplifier calibration, allowing optimization of  $\pi$  in the 1 kHz to 1 MHz frequency range.

Step 8: Repeat step 5

**Measurement results**

Figure 11 shows the magnitude of the distortion rejection factor as a function of frequency,

achieved for the experimental prototypes which have been calibrated.

We see that  $\pi$  values extending from magnitudes of 30-60 dB have been achieved in the wide frequency range 200 Hz to 1 MHz. Even better results can be expected with more care taken in layout and power distribution design. These values translate into an equivalent degree of reduction of the total harmonic distortion, thd, of the power amplifier, as demonstrated by test results (Figs. 12 and 13).

Two significant levels of the total bias current  $I_{bias}$  of the power amplifier mosfet output stage are taken into account. The first one,  $I_{bias} = 1 \text{ mA}$ , representing a very poor biasing level, helps prove the ability of the proposed technique to counter-balance the effects of the comparatively high output impedance of the power amplifier in the crossover region.

In addition, it shows the ability of feed-forward to reject high-order harmonics normally generated by poorly-biased output stages.

The second level,  $I_{bias} = 100 \text{ mA}$ , is closer to the normal level of biasing of power mosfet output stages and demonstrates the effectiveness of the proposed technique to correct small amounts of distortion.

Results (Figs. 12 and 13) show that the measured improvement ratio of about 30 dB is in good agreement with the value of distortion rejection reported in Fig. 11, and gives clear evidence of the effectiveness of the proposed feed-forward technique.

Only the worst case ( $f = 20 \text{ kHz}$ ) thd+noise versus output level (volt peak-to-peak/8 $\Omega$  load) is reported. All other measurements taken at  $f < 10 \text{ kHz}$  are, after applying the error correction technique, very close to the instrumentation limits.

Effectiveness of the distortion rejection mechanism with audio programs, has also been simulated by superposing a white-noise voltage at the output of the power amplifier.

A white-noise level of  $V_n = 0.5 \text{ Vrms}$  was injected at input node N while the amplifier was delivering 20 Vpk-pk to the load with  $f$  at 1 kHz. Unfiltered noise appearing across the load was 32 dB lower than that measured at the output of PA – a high level of rejection in agreement with theoretical expectations.

The final test report refers to the output noise levels of the amplifier, before and after correction. They are 0.79 mV and 0.38 mV, respectively. ■

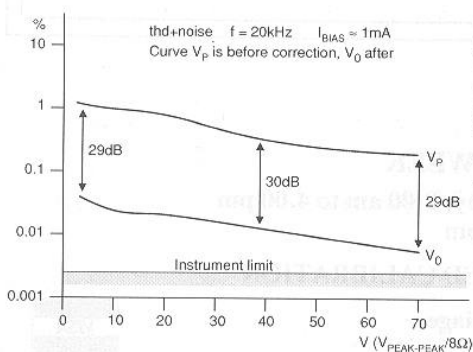
**Components**

These component values were used in the prototype. All resistors have 1% tolerance.

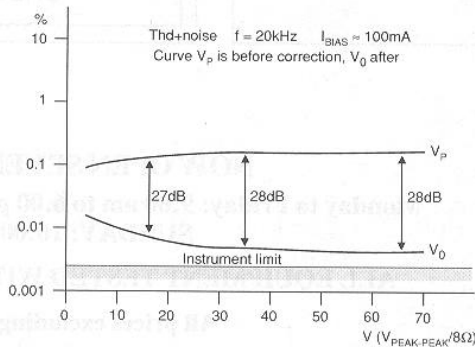
$R_1$	= 12.5 k $\Omega$
$G_0 R_1$	= 226 k $\Omega$
$G_0$	= 18.08
$R_2$	= 89.6 $\Omega$ (nom)
$G_0'' R_2$	= 1.62 k $\Omega$
$R_4$	= 10 k $\Omega$
$R_5$	= 5 $\Omega$
$R_6$	= 110 $\Omega$
$R_7$	= 332 $\Omega$
$R_8$	= 5.11 $\Omega$
$R_9$	= 1 k $\Omega$
$R_{10}$	= 0.7 k $\Omega$ (nom)
$R_{11}$	= 1.3 k $\Omega$ (nom)
$R_{12}$	= 4.7 k $\Omega$
$R_{13}$	= 909 $\Omega$
$R_{14}$	= 17.8 k $\Omega$
$C_2$	= 1 $\mu\text{F}$
$C_3$	= 13.5 pF (nom)
$C_4$	= 188 pF (nom)
$C_6$	= 13.5 nF
$T_1$	= 220 $\Omega$
$T_2$	= 1 k $\Omega$
$T_3$	= 2 k $\Omega$
$L_5$	= 1 $\mu\text{H}$
$R_s$	= 8 $\Omega$
$R_z$	= 10 $\Omega$
$C_z$	= 47 nF

**References**

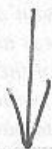
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**Fig. 12. Total harmonic distortion + noise versus output level. Test conditions:  $I_{bias} = 1 \text{ mA}$ ,  $f = 20 \text{ kHz}$ .**



**Fig. 13. Total harmonic distortion + noise versus output level. Test conditions:  $I_{bias} = 100 \text{ mA}$ ,  $f = 20 \text{ kHz}$ .**



## Feed-forward feedback

I read with interest Giovanni Stochino's article (*Audio design leaps forward*, EW+WW, October, pp. 818-824) on the application of feed-forward in Audio.

But I believe that the concept can be developed much more clearly and simply. First, we have his block diagram showing how power amplifier error is extracted, determined and isolated via a criterion network comprising  $\alpha$ ,  $\gamma$  and  $SC$ ; and the formula  $V_o = V_p - V_a$ .

Next, Mr Stochino introduces  $E_p$  and  $E_a$  (the error components in the main and auxiliary amplifiers), which is confusing as these are not shown in the figure. It is much simpler to assume that all errors, noise etc. are contained in the terms  $G_p$  and  $G_a$ , describing the two amplifiers. The expression is then developed straightforwardly by substitution:

$$\begin{aligned}V_p &= V_i \cdot G_p; \\V_p' &= \lambda \cdot V_p; \\V_p'' &= \lambda \cdot V_i \cdot G_p.\end{aligned}$$

Similarly,

$$\begin{aligned}V_a &= V_e \cdot G_a; \\V_e &= V_i \cdot \alpha + V_p \cdot \gamma; \\V_e &= V_i \cdot \alpha + V_i \cdot G_p \cdot \gamma.\end{aligned}$$

thus:

$$V_a = G_a \cdot V_i (\alpha + \gamma \cdot G_p).$$

The output  $V_o$  then becomes:

$$V_o = V_i \cdot G_p \cdot \lambda - G_a \cdot V_i \cdot (\alpha + G_p \cdot \gamma).$$

The system gain is:

$$V_o/V_i = G_p \cdot \lambda - G_a \cdot \alpha - G_p \cdot G_a \cdot \gamma.$$

Clearly, the system is independent of the main amplifier if:

$$G_p \cdot \lambda - G_p \cdot G_a \cdot \gamma = 0.$$

This leads to the condition (assuming  $\lambda = 1$ )

$$G_a \cdot \gamma = 1,$$

which is equivalent to Mr. Stochino's result, but is much easier to understand.

From this result it can also be concluded that in this case the system gain becomes:

$$V_o/V_i = -G_a \cdot \alpha.$$

Furthermore, the condition  $G_o' = G_o'' = G_o$  is not necessary.

However,  $\alpha = 1/\gamma$ , which follows from Figs. 2 and 5.

By the way, I wonder whether a copy of the patent application would be available. What is it exactly that Mr. Stochino wishes to patent?

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