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[1] in the 1920s feed-forward error correction has found practical application in radio frequency and microwave amplifiers[2]. But it has never been used, in Black's form, in audio power amplifiers[3]. 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 $V_p=V_iG_p+E_p$. G_p is the voltage 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 (α, γ and *SC*) to determine, isolate and extract power amplifier error; an auxiliary amplifier *AA* (low power requirement, low distortion

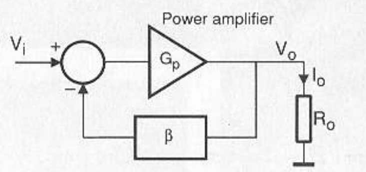


Fig. 1. Basic elements of a feedback amplifier. Analysis shows that distortion can never be completely nulled with this configuration.

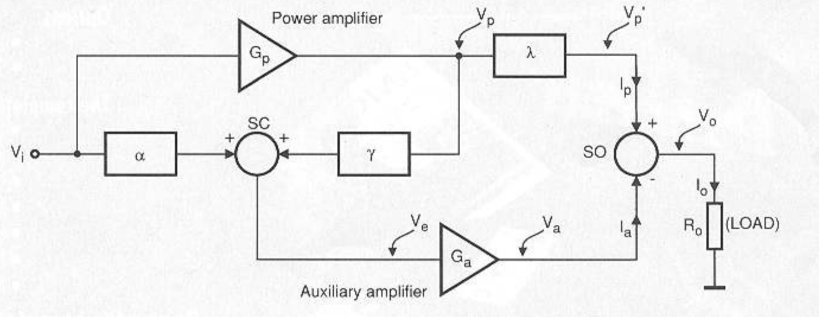


Fig. 2. Principles behind feed-forward. In this configuration, power amplifier error is extracted, determined and isolated via a criterion network comprising α, γ and SC.

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_0. Phase-amplitude equaliser network λ 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:

$V_e = V_i\alpha + V_p\gamma$

$$V_0=V_p'-V_a$$
$$=V_iG_p\lambda-G_aV_i(\alpha+\gamma G_p)+E_p(\lambda-\gamma G_a)+E_a$$

where E_a is the error component (distortion plus noise) produced by the auxiliary amplifier. Proper operation of the feed-forward technique requires that $E_a<<E_p$ 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 α 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.

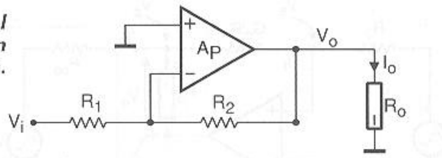
As a technique, it is less effective at the highest frequency of the audio range and in the crossover region of class AB amplifiers, where the feedback factor can be low and deviation from linearity is high[4].

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.

Fig. 3. Simplicity of a practical feedback amplifier configuration with inverting gain.



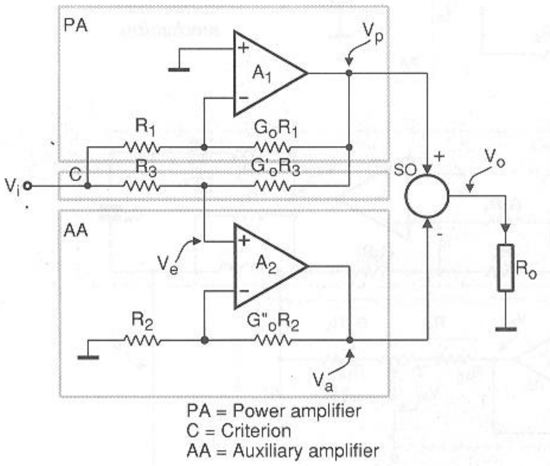
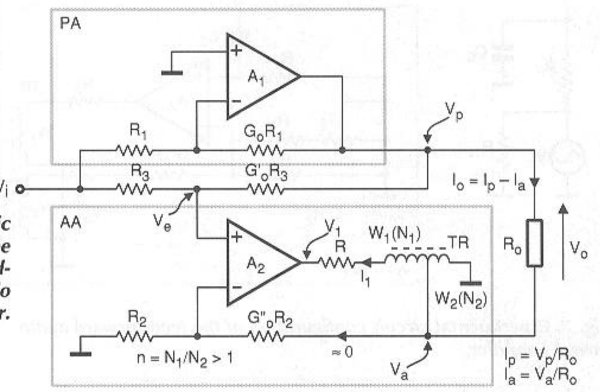


Fig. 4. A practical feed-forward amplifier incorporating feedback amplifiers.

Fig. 5. Basic scheme of the proposed feed-forward audio power amplifier.



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: $G_p=V_p/V_i\cong-G_0$ and $G_a=V_a/V_e\cong(1+G_0'')$ provided $A_1/G_0>>1$ and $A_2/G_0''>>1$. 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 $G_0'=G_0$. $\alpha \approx 1/\gamma$

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 $V_p-V_a=-G_0V_i+E_p-E_p+E_a=-G_0V+E_a$, 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 $E_a<<E_p$, 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[3].

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.

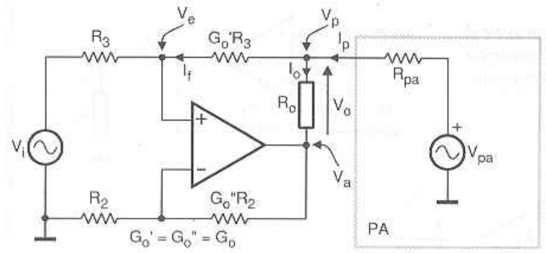


Fig. 6. Non-zero output resistance of PA does not influence the properties of the feed-forward correction mechanism.

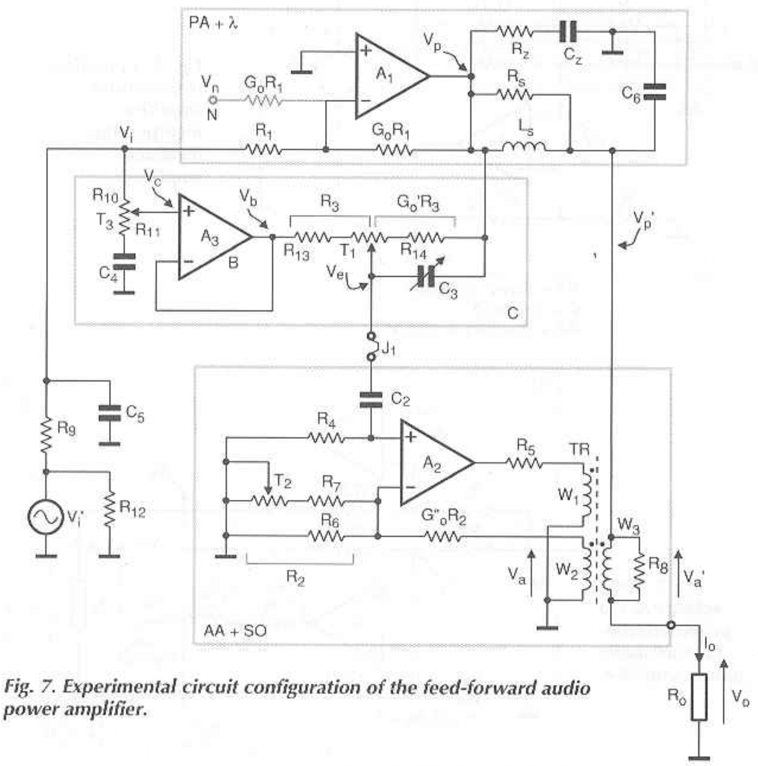


Fig. 7. Experimental circuit configuration of the feed-forward audio power amplifier.

Feed-forward error correction (always intra-loop) is sometimes used in audio power amplifiers[5, 6], 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**[7]) 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 $2\pi f B_m S_e N_2$ where S_e 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_e, 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_e at 100mm^2 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>>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 ±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_i, and power amplifier output impedance could be high and non-linear without affecting the output voltage value ($V_0=-G_0V_i$ in Fig. 5).

We can also deduce, mathematically, that $V_a=V_iG_0+V_p$ and $V_0=V_p-V_iG_0-V_p=-V_iG_0$ 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 $G_0R_3>>R_{pa}$. In this case we have I_f $<<I_p$ so $I_p \cong I_0$ and the node voltages are:

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

and,

$$V_a=V_{pa}+G_0V_i.$$

Solving simultaneously, substituting and assuming in normal operation that $V_{pa}=-G_0V_i+E_p$, we have:

$$V_p=-G_0V_i(1-R_{pa}/R_0)+E_p$$

and

$$V_a=G_0V_iR_{pa}/R_0+E_p$$

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 ($R_{pa}=\infty$),

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

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_g 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 ∞.

In the Zobel network (R_z, C_z, R_s and L_s) commonly used at the output of class AB audio power amplifiers, R_s and L_s in conjunction with C_6 implement the λ 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

Magnetic flux cancellation

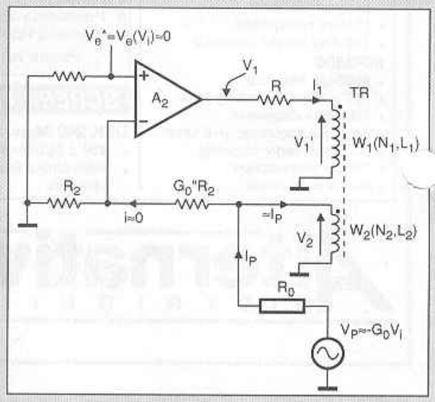
This diagram helps analyse the mechanism of magnetic flux cancellation $\Phi(V_i)$ in the transfomer core, due to main signal current $I_p \approx V_p/R_0 \approx -G_0V_i/R_0$. It is operated by the auxiliary amplifier.

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

$$V_2(V_i) \cong -G_0V_i\frac{R}{n^2R_0}\cdot\frac{1}{\{1+A_2/[n(1+G_0'')]\}}$$

From this,

$$\lim V_2(V_i)=\Phi(V_i)=0$$
$$(A_2 \rightarrow \infty \text{and/or} R \rightarrow 0)$$



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 $V_e^* = V_e(V_i)$ 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.

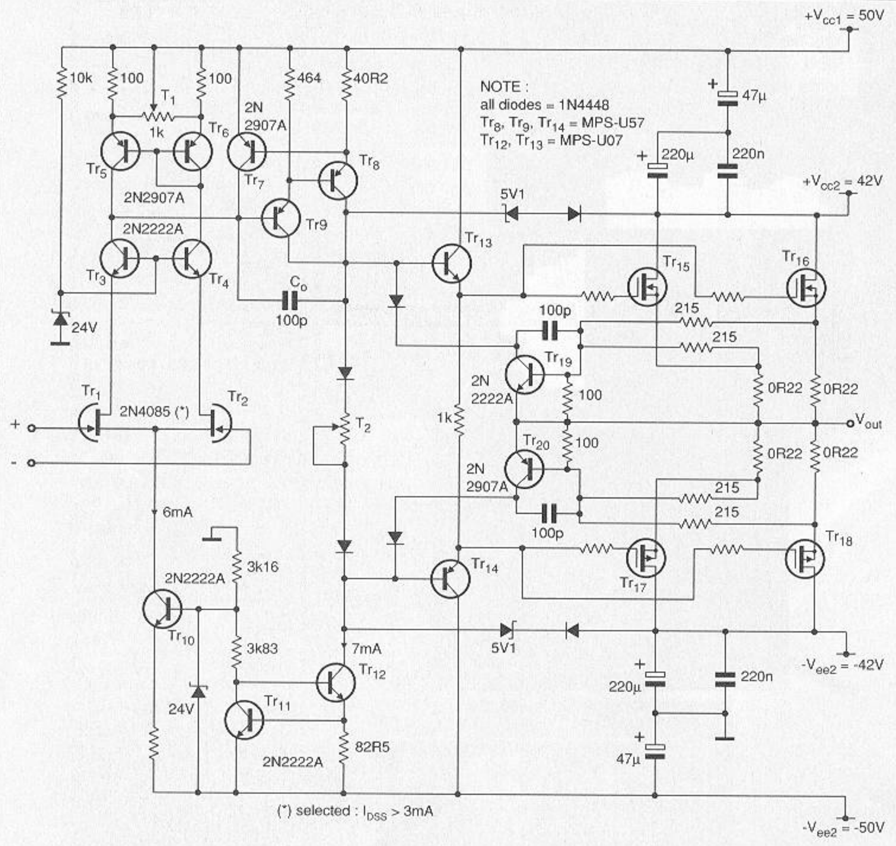


Fig. 8. Complete circuit diagram of the power operational amplifier A_1. Output power into 8Ω is 100W and slew rate is 60V/µs.

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 $V_p=[-V_iG_0/(1+s/p_1)]+E_p$. Also,

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

where p_0 is $1/[C_4(R+R_{11})]$, z_0 is $1/C_4R_{11}$ and V_a is $V_e[(1+G_0)/(1+s/p_2)]+E_a$.

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

As for λ – assuming a first order response of the auxiliary amplifier – we expect a first order low-pass frequency response so that $\lambda\cong 1/(1+s/p_5)$ where $p_5=1/(C_6R_6)$, corresponding to the assumption that $R_5<<sL_s$.

Regarding the value of the distortion rejection factor, the validity of the above expressions for λ 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

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

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

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

$$z_3=1/(G_0R_3C_3)$$

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 γE_p.

We also want to satisfy the condition $\gamma G_a=\lambda$. Substituting for each term reveals that this condition is true if $z3=p_2$ 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:

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

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 λ.

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 λ, 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 $1/L_sC_6$ shows the additional potential of the λ network to compensate for a more realistic second-order frequency response of the auxiliary amplifier, by suitable choice of z_s and ω_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 A_1 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 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.

Main characteristics of A_1

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

Auxiliary op-amp A_2

Amplifier A_2, Fig. 9, has a mosfet output

Main characteristics – amplifier A_2

Voltage gain	≥10^8
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

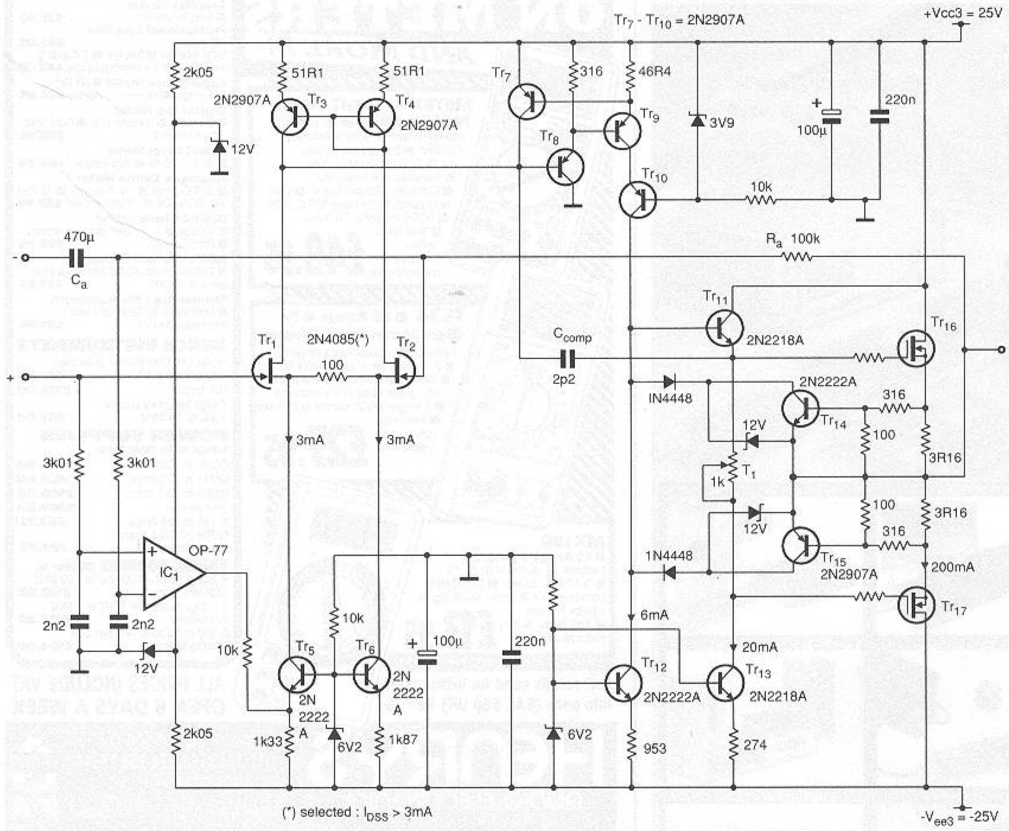


Fig. 9. Power output requirement for the auxiliary operational amplifier, A_2, is lower. In class A, output is ±200mA while in class AB it rises to ±800mA.

stage, biased by adjusting trimmer T_1 at 200mA 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 ($Tr_{1\text{-}17}$) 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/600Ω from 20Hz to 20kHz	≤0.0005
—3dB small-signal bandwidth	≅15MHz
Voltage noise density	≅1nV/√Hz
Slew rate	≅±20V/µs

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 diameter and 7mm high. Its cross-sectional area S_e is approximately 31mm^2. The core material is Ferroxcube-grade 3E2, having a saturation flux density B_s of about 350mT, and a useful linear range of ±200mT. 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 wires and all windings are uniformly wound along the core length. When driving the primary winding with a source resistance of 5Ω, the –3dB small signal bandwidth of the transformer extends from 5Hz to about 13MHz.

Complete auxiliary amplifier and transformer characteristics

Voltage gain	≅18.1
Gain-bandwidth product	≅10MHz
Slew rate	≅±17V/µs
Thd+noise @5kHz	≅0.01%
Thd, V_a=0.5V/1Ω @50kHz	≅0.05%

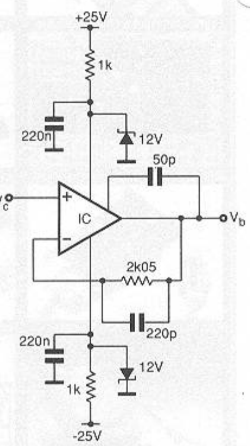


Fig. 10. Buffer B. At frequencies from 20Hz to 20kHz, distortion of this design is less than or equl to 0.0005%.

Amplifier calibration

Amplifier calibration has been performed with a load of 8Ω in parallel with 0.2µF using the following procedure.

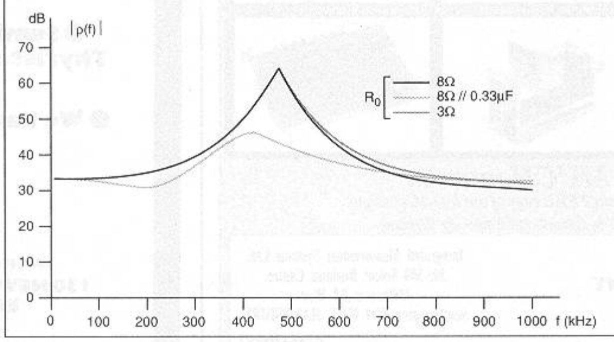


Fig. 11. Magnitude of the distortion rejection factor ρ after calibration.

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

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

$f_{c1}=p_1/2\pi \cong 422kHz$
$f_{c2}=p_2/2\pi \cong 650kHz$

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

$C_3=1/(G_0R_3p_2) \cong 13.5pF$.

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 $z_0=1/C_4R_{11}$ and using $p_0=p_1$ and $z_0=z_3$. Since $R_{10}+R_{11}=R_{T3}=2k\Omega$ we obtain:

$C_4 = 1/[(R_{10} + R_{11})p_1] \cong 188pF$
$R_{11} = 1/C_4z_3 = 1/C_4p_2 \cong 1.3k\Omega$.

Step 5: The aim is to meet the condition defined by $\gamma G_a=\lambda$. Signal V_i'=100mV/3kHz 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 100kHz 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 $V_n \cong 50mV$ (the amplitude of V_n must be kept below the limits set by $V_{am} = 2\pi fB_mS_eN_2$, as shown in Fig. 7. Since $E_{pn} = E_n$, this method maximises, in a wide frequency range (up to 1MHz), the distortion rejection factor $\rho = 1/(\lambda - \gamma G_p)$ of the auxiliary amplifier.

Frequency of V_n is first set at 3kHz 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 300kHz and C_3 is adjusted again to have maximum rejection. A network analyser would simplify amplifier calibration, allowing optimization of π in the 1kHz to 1MHz 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 π values extending from magnitudes of 30-60dB have been achieved in the wide frequency range 200Hz to 1MHz. 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}=1mA, 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}=100mA, 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 30dB 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=20kHz) thd+noise versus output level (volt peak-to-peak/8Ω load) is reported. All other measurements taken at f<10kHz 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=0.5Vrms was injected at input node *N* while the amplifier was delivering 20Vpk-pk to the load with f at 1kHz. Unfiltered noise appearing across the load was 32dB 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 0.79mV and 0.38mV, respectively. ■

Components

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

R_1	= 12.5kΩ
G_0R_1	= 226kΩ
G_0	= 18.08
R_2	= 89.6Ω (nom)
$G_0''R_2$	= 1.62kΩ
R_4	= 10kΩ
R_5	= 5Ω
R_6	= 110Ω
R_7	= 332Ω
R_8	= 5.11Ω
R_9	= 1kΩ
R_{10}	= 0.7kΩ (nom)
R_{11}	= 1.3kΩ (nom)
R_{12}	= 4.7kΩ
R_{13}	= 909Ω
R_{14}	= 17.8kΩ
C_2	= 1μf
C_3	= 13.5pF (nom)
C_4	≅ 188pF (nom)
C_6	= 13.5nF
T_1	= 220Ω
T_2	= 1kΩ
T_3	= 2kΩ
L_s	= 1μH
R_s	= 8Ω
R_z	= 10Ω
C_z	= 47nF

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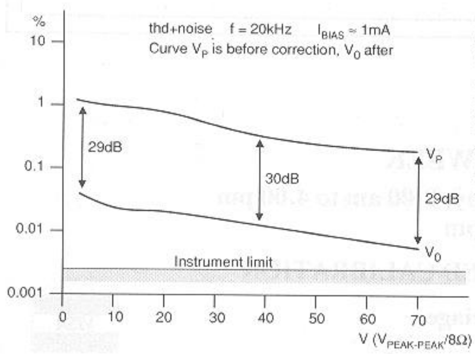

Fig. 12. Total harmonic distortion + noise versus output level. Test conditions: I_{bias} = 1mA, f = 20kHz.

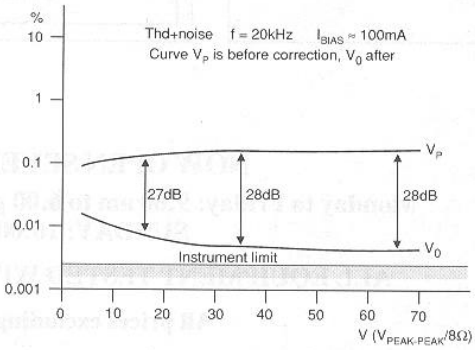

Fig. 13. Total harmonic distortion + noise versus output level. Test conditions: I_{bias}=100mA, f=20kHz.

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 α, γ 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:

$V_p = V_i.G_p;$
$V_p' = \lambda.V_p;$
$V_p' = \lambda.V_i.G_p.$

Similarly,

$V_a = V_e.G_a;$
$V_e = V_i.\alpha + V_p.\gamma,$
$V_e = V_i.\alpha + V_i.G_p.\gamma.$

thus:

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

The output V_o then becomes:

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

The system gain is:

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

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

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

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

$G_a.\gamma = 1,$

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:

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

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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