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Error Correction in Class AB Power Amplifiers

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ABSTRACT

An audio power amplifier design is presented which can linearize the transfer characteristics of conventional class AB output stages in the crossover region. The objective is to offer practical circuits with error correction, that overcome nonlinearities, inherent to the crossover region of class AB output stages.

0 INTRODUCTION

Error correction is a powerful tool for improving amplifier distortion. An amplifier system with error correction uses a second signal path to amplify the error signal. The error signal is defined as the difference between the input and the output of the main channel. The output of the auxiliary path is combined with the output signal of the main path so that the error, originating in the main channel, is canceled at the load terminals. In low frequency amplifiers an output-summing network comprises a resistive divider, otherwise a floating load is required.

The basic idea is that the conventional class AB output stage is already consist of two parallel

amplifying paths, that work simultaneously in the crossover region, where the output voltage passes through zero. There are many distortion mechanisms associated with that region.

The classic crossover distortion results from the non-conjugate nature of the transfer functions of output devices [1]. There is the narrow region of bias voltage (offset between transfer functions), where the crossover distortion is kept at low level. Over-bias produces the same sort of higher-order harmonics distortion spectrum as under-bias [2]. Instantaneous changes in output devices' junction temperature with output signal lead to continuous deviation from the optimal bias.

The output transistors are generally operated to cut-off. The delay in cutting the bipolar transistors off results in extra instantaneous current flow. That extra current flows through the other output device, increasing its transconductance. Thus even an optimally biased output stage will be over-biased and will exhibit crossover distortion with increase in frequency.

The crossover distortion appears audible as a noise modulation [3], subjectively reveals as loss of micro dynamics and listening fatigue.

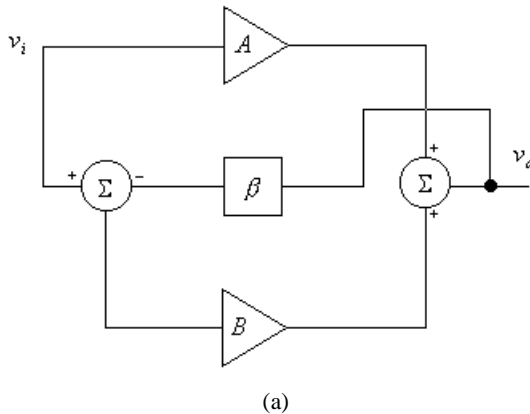
A recent paper [4] describes how the difference between the input and the output of one half of class AB output stage could be amplified in proper phase by the other half, and the resulting signal is free from nonlinearities. It was demonstrated, that the crossover distortion was significantly reduced and the distortion spikes remained below the noise level. An attempt to improve the original design [4] by keeping both halves of the output stage in class-A was presented [5]. Instead of measuring difference between the input and the output of one half, another approach deal with the difference between the input and the output of the whole stage [5].

This paper discusses practical aspects of error correction in class AB output stages, estimates the effect of error correction, presents new topologies and gives a better understanding of the method.

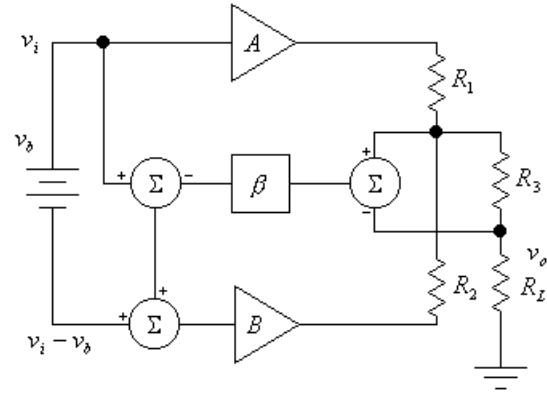
1 PRINCIPLES OF OPERATION

1.1 Error feedback and error feedforward

Two generalized error-correction topologies were discussed [6]. When analyzing these topologies, let us consider only error correction using a resistive output-summing network.



(a)



(b)

Fig.1. Output injected active error feedback (a) and its application to push-pull transconductance output stage (b).

R_3 is the output current sensor.

For the output-injected active error feedback (Fig. 1(a)) the transfer is

$$G_1 = \frac{v_o}{v_i} = \frac{A+B}{1+\beta B} \approx \frac{1}{\beta}. \quad (1.1)$$

The sensitivities of overall gain G_1 to changes in the main and auxiliary paths are

$$S[A] = \frac{A}{A+B} \approx \frac{1}{1+\beta B} = \frac{1}{1+G_1}, \quad (1.2)$$

$$S[B] = \frac{B(1-\beta A)}{(A+B)(1+\beta A)} \approx \frac{1-\beta A}{1+\beta B} = \frac{1-\beta A}{1+G_1}, \quad (1.3)$$

where $G_1 = -\beta_{eff} B = \beta B$ is effective loop gain.

The approximate forms of Eqs. (1.1)-(1.3) require both that $\beta B > 10$ (which implies a large loop gain) and also that $A \approx 1/\beta$. That implies that the deviation of the main-channel gain A from the required gain $1/\beta$ is small, thus $S[B]$ is low. The latter is a usual requirement for error-correcting systems - the distortion originated from auxiliary path should be negligible.

The analysis on the basis of total available gain [6] shows that the sensitivity to main path variations and loop gain are reciprocal. The sensitivity to variations in the main amplifier, hence distortion reduction and the stability margin, are solely determined by the total loop gain value. It ought to be mentioned that other circuit elements like summing and feedback networks are assumed to be ideal. While the decrease

of sensitivity to main path variations is exchanged for the correspondent gain lost, like in negative feedback, the error correction can be offered as an alternative design solution.

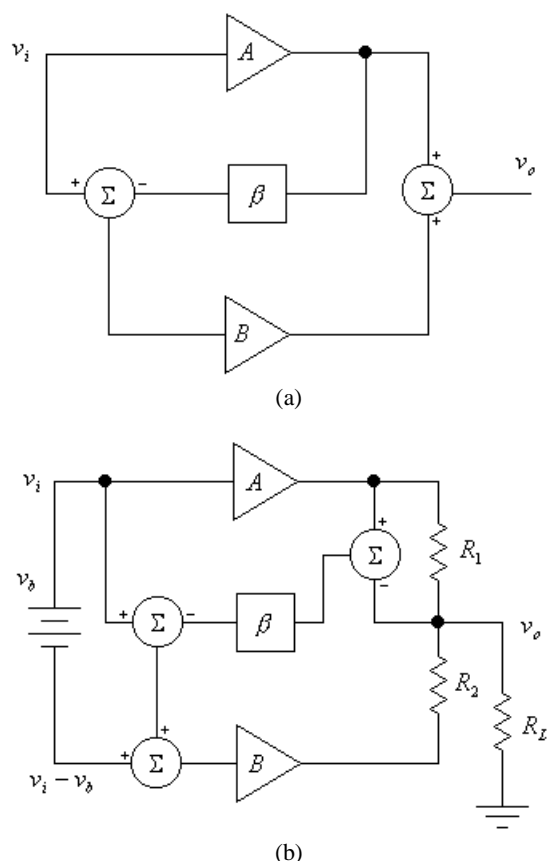


Fig.2. Feedforward error correction (a) and its application to push-pull output stage (b). R_2 is the main path output current sensor and also is a part of the summing network.

For the feedforward error correction (Fig. 2(a)) the transfer function is

$$G_2 = \frac{v_o}{v_i} = A + B(1 - \beta A) \approx A. \quad (1.5)$$

The sensitivities of overall gain G_2 to changes in the main and auxiliary paths are

$$S[A] = \frac{A(1 - \beta B)}{A + B(1 - \beta A)}, \quad (1.6)$$

$$S[B] = \frac{B(1 - \beta A)}{A + B(1 - \beta A)} . \quad (1.7)$$

$S[A]$ is nearly zero and $G_2 \approx A$, when the balance condition $\beta B \approx 1$ is satisfied.

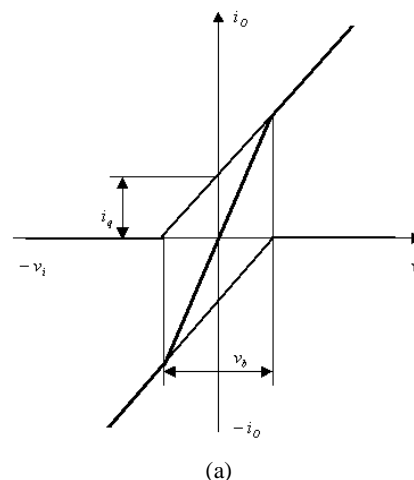
The apparent distinction from error feedforward is

that the correction signal in active error feedback is injected in the main path in front of the error-sensing β network.

The power devices used in output stage can be modeled as a voltage-controlled current sources. The principle of error correction can be used with these models, as shown in Fig.1(b) and Fig.2(b). The equations (1.1)-(1.7) can be rewritten for the output current instead of output voltage. Application of the error correction to push-pull output stage requires Y current summing network. The output devices emitter/source resistors can be utilized as components of output network. Those low value resistors can be a source of distortion themselves (from the authors' measurements up to 0.05% THD). The careful comparison of Fig.1(b) and Fig.2(b) reveals that in active error feedback these resistors are located inside the error loop, so their imbalance is not as influential as in the case of error feedforward.

1.2 Construction of the required transfer function

To lower crossover distortion, the quiescent current should be kept at the optimum point. The biasing voltage is always set at some higher level to ensure, that the quiescent current will never fall to zero due to thermal runaway. It was shown [1,2], that the increase in the quiescent current does not make the distortion figures of a push-pull stage better. Both devices conduct simultaneously in the center of the output voltage range, thus the transconductance in this region is doubled (g_m doubling). That change in transconductance lead to a wide distortion spectrum. Fig.3(a) shows the transfer characteristic of the output stage in the overbiased state.



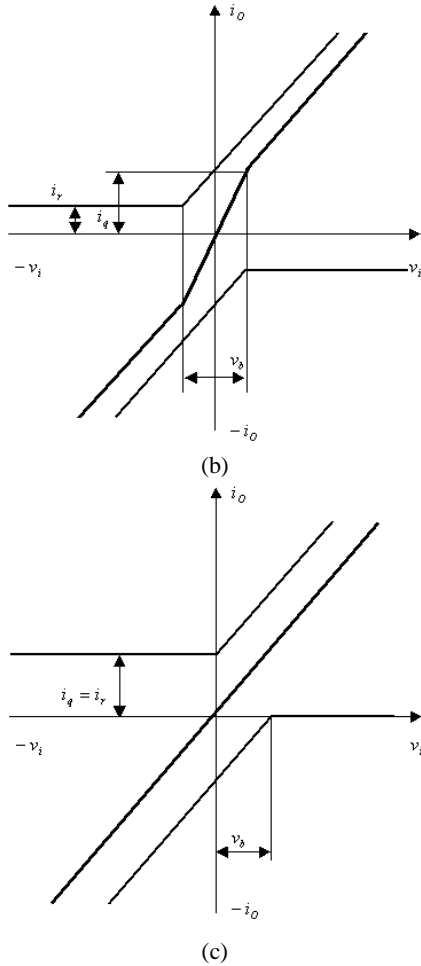


Fig.3. The crossover point characteristics show (a) over-biased, (b) nonswitching current setting and (c) error correction (i_o - the output current, v_i - input voltage, i_q - quiescent current, i_r - residual current, which prevent output device to switch off completely).

Another form of distortion is switching distortion, which associated with failure of the output devices to turn-off completely at high frequencies. A lot of circuits were proposed to eliminate switching distortion [7-9]. The main idea of these circuits is illustrated in Fig.3(b). The conventional bias circuit is modified such a way, that the current flow via any of the output device will never decreased to zero, but remains at some constant level i_r . In this region the output device operates as quasi-constant current circuit. While reducing switching distortion, these circuits are unable to prevent from g_m doubling. But then it is easier to get the right signals from the splitter, than to construct an output stage with stabilized optimum quiescent current i_q . If the splitter is not perfect, one will get under- or

overbiased state. Some of these circuits use a kind of positive feedback and they potentially carry the risk of uncontrolled common-mode conduction [1].

The original idea [4] is to combine error correction principle with the non-switching. It is illustrated in Fig.3(c). One half in that design remains usual highly nonlinear half, and all the significant crossover non-linearity will originate in it. The other half is low distortion non-switching, have only low level of mainly second harmonic distortion, which can be reduced further by the overall negative feedback. The bottom half supplies all the negative load current and small current i_q to prevent the upper half from switching off, while the upper one supplies positive current, and it needs only a small quiescent current i_q to keep it in constant current state at negative load current. The error correction keeps that current i_q (or the biasing voltage v_b) at such a level, so that the overall transfer function becomes a straight line in the crossover region and near it. So if the quiescent current can be fixed, the error correction will compensate the thermal drift in the other output device. It should be noted, that the output impedance of the stage with error correction will remains constant. While one output device is conducting, the other will be turned off or will be in the high impedance current source mode.

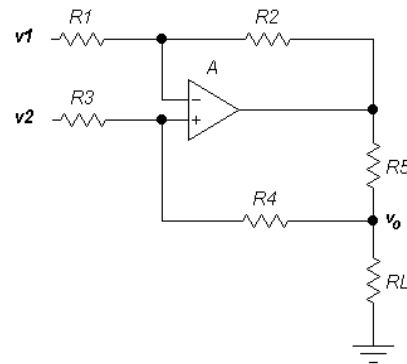


Fig.4. Conventional op-amp based current source.

For combination of current circuit and error feedback one can start with the conventional op-amp based current source [4] as a non-switching half. It is shown in Fig.4. The output current i_o of the circuit

$$i_o = \frac{1}{R_5} \left(\frac{R_4}{R_3} v_2 - \frac{R_2}{R_1} v_1 \right)$$

will be independent from the R_L value when

$$\frac{R_1}{R_1 + R_2} = \frac{R_3}{R_3 + R_4}.$$

Instead of that traditional design other elegant and useful topologies [10] can be suitable.

It should be again underlined, that in contrast to previous design [11] two halves are essentially different: one is a low distortion non-switching amplifier, while other is a conventional output stage.

2 PRACTICAL REALIZATIONS

There have been two alternatives for error correction: active error feedback and error feedforward and two variables, which the error signal can control: quiescent current i_q and biasing voltage v_b . A large number of circuit topologies can be generated. Some of them are presented below.

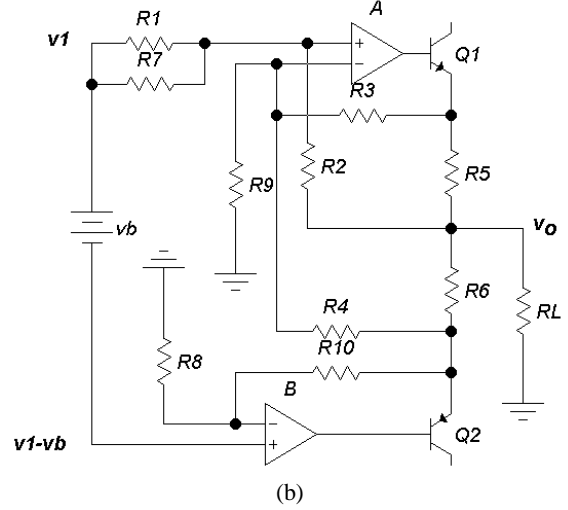
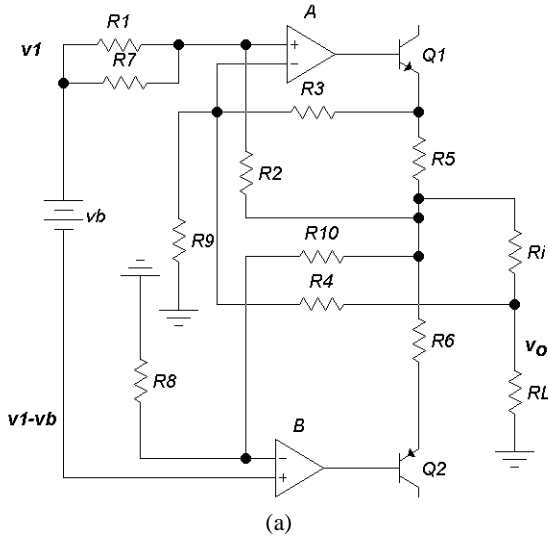


Fig.5. The amplifier with the op-amp based current source as the push-pull half in active error feedback (a) (R_i is the output current sensor) and feedforward (b) (R_5 is the main path output current sensor and also is a part of the summing network) configuration.

For the system gain G the circuit parameters are

$$R_1 = R_2 = R_3 = R_4, \quad R_7 = R_8 = R_9 = \frac{R_1}{G-1},$$

$$R_5 = R_6 = R_i \text{ for the circuit, presented in Fig.5(a)}$$

$$\text{and } R_1 = R_2 = R_3 = R_4, \quad R_7 = R_8 = R_9 = \frac{R_1}{G-1},$$

$$R_5 = R_6 = R_i \text{ for the circuit Fig.5(b).}$$

The balance condition for the circuit Fig.5(b) is $R_4 R_5 = R_3 R_6$.

A problem with the circuits Fig.5 is that if a discrete component output stage is used the input impedance may be relatively low and non-linear, and feeding this impedance from two resistors R_1, R_2 may add gain error and distortion at this point. R_1, R_2 could be reduced in value to decrease these effects, but it also reduces the impedance seen by the driver stage and makes it dependent on the load, so this is not entirely satisfactory.

Fortunately there is a simple way to eliminate these resistors, as shown in Fig.6. The input is taken directly to the top differential stage, giving a higher gain to the non-inverting input. The gain to the inverting input must be increased to compensate for this to maintain the correct output voltage, and this can be achieved by adding one resistor to the bottom

class-B half.

The in-depth analytical study is presented in Appendix 1.

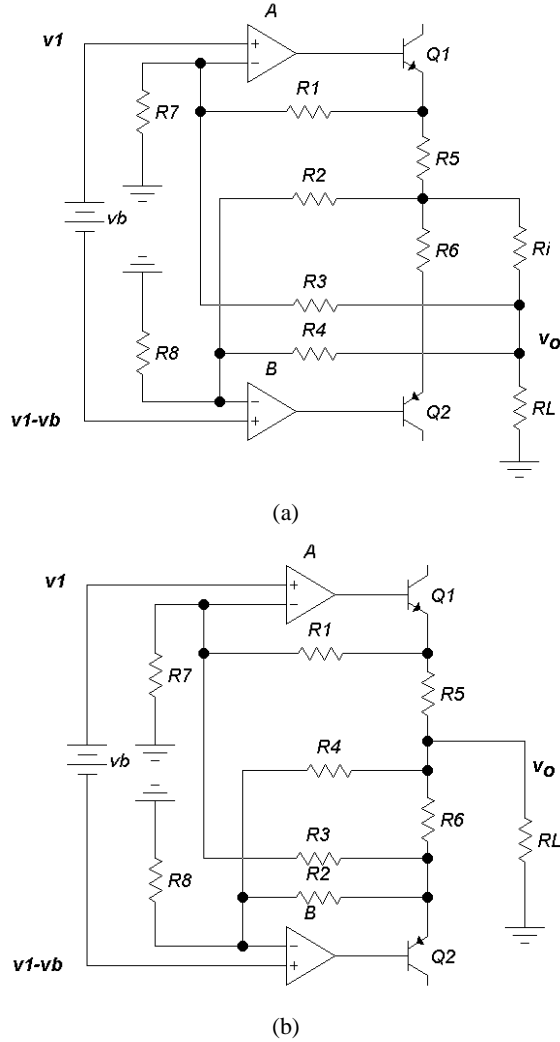


Fig.6. The amplifier topology in active error feedback (a) (R_i is the output current sensor) and feedforward (b) (R_5 is the main path output current sensor and also is a part of the summing network) configuration, which eliminates undesirable loading of the driver stage.

For the system gain G the circuit parameters are

$$R_1 = R_2 = R_3 = R_4, R_7 = R_8 = \frac{R_1}{2(G-1)},$$

$$R_5 = R_6 = R_i \text{ for the circuit, presented in Fig.6(a)}$$

$$\text{and } R_1 = R_2 = R_3 = R_4, R_7 = R_8 = \frac{R_1}{2(G-1)},$$

$$R_5 = R_6 \text{ for the circuit Fig.6(b).}$$

The balance condition for the circuit Fig.6(b) is $R_3 R_5 = R_1 R_6$.

The Quad triples [2] can be used as the amplifying blocks A, B [4]. While the design is very simple, it can have certain stability problems. At high power levels there is an additional limitation. Emitter current flow through the feedback resistors of the triple can lead to a substantial voltage drop across them. This requires an additional input voltage to the output stage, which in the negative direction will certainly be enough to switch off the class-A current mode error amplifier. The error correction mechanism then can become inoperative at high output currents. The illustration is presented in Appendix 2. Choosing more complex high gain stages able to include frequency compensation [12,13] could help to avoid these problems.

It seems appealing to apply the described method to the conventional common-emitter/source output power stage. In this application, the floating ground amplifier may be helpful. The floating ground is a circuit whose electrical common point is not at earth ground potential or the same ground potential as the circuitry it is associated with. Let us address to the floating ground follower. It is shown in Fig.7.

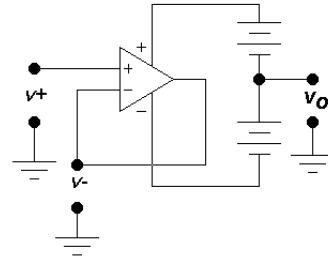


Fig.7. Floating ground voltage follower.

The output voltage v_o of the circuit is

$$v_o = v^+ - v^-.$$

While the input of the follower is single-ended the floating follower performs just as well as an amplifier whose input is differential. That circuit can work very well as a subtractor for the conventional common-emitter output power stage, as shown in Fig.8(a). The feedforward topologies will be described forth below, the similar circuits can be obtained for the error feedback.

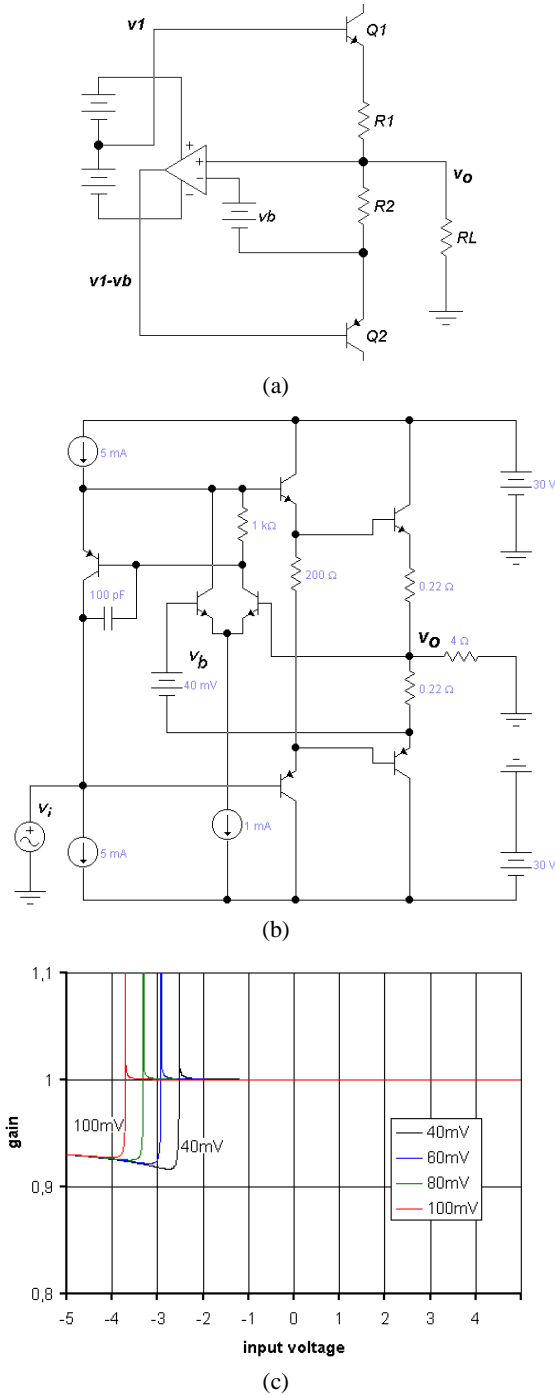


Fig.8. Emitter follower output stage with a floating ground voltage follower as a subtractor (a). The output stage with a 'novel current controller' [14] as a subtractor (b). The

output stage small-signal gain $\frac{dv_o}{dv_i}$ versus input voltage

v_i , with bias v_b as a parameter (c).

In Fig.8(b) the output stage incorporated a 'novel current controller' [14] as a subtractor. The small-

signal gain deviations for this stage is shown in Fig.8(c), that can be compared with emitter follower small-signal gain without correction, as shown in Fig.15, Appendix 3. The crossover region is vanished, and the gain changed to perfect straight line at the unity level for the positive input voltages, when the feedforward is in the action. Using feedback (e.g. Sziklai feedback pair) in the half without error correction can make an improvement to symmetry.

Instead of differential pair [14] in a floating ground voltage follower one can use conventional [7] or enhanced [9, 15] current mirrors. One of the possible configurations is presented in Fig.9.

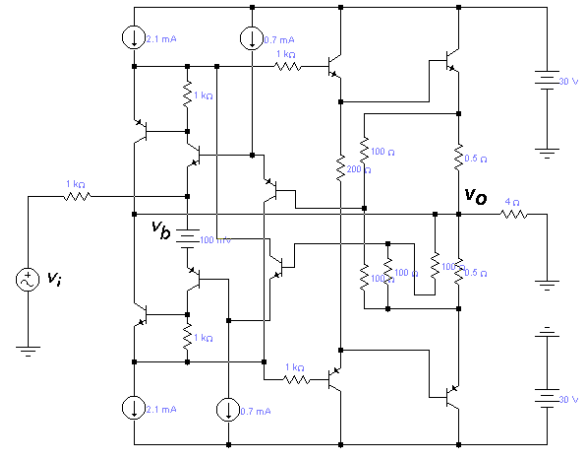


Fig.9. The double emitter-follower output stage with a complementary current mirror as a subtractor and error amplifier.

For the accurate error correction the driver stage should transfer the correction signal from the subtractor to the bases/gates of the output devices as accurate as possible. Three of the possible double emitter-followers configurations are discussed in Appendix 3.

3 CONCLUSIONS

A series of circuits were presented which are suitable for the output stage of audio amplifier. It is hoped that some of that circuits will prove useful for the designers.

The circuits presented here solve the most serious problems of class AB output stage design. The simple enhancement of a standard output stage dramatically reduces crossover distortion. It can also remove the need for accurate setting and thermal compensation of quiescent current - making the design suitable for volume production. The efficiency is a little lower than that of a conventional class B

amplifier due to small extra power dissipation in the non-switching half.

4 REFERENCES

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APPENDIX 1

In this section the virtue of error correction will be estimated. The feedforward error correction will be taken as an example. Exactly the same results can be obtained for the error feedback, with the exception that the balance condition does not exist for the error feedback [6]. The diagram of the first circuit [4] is shown in Fig.10.

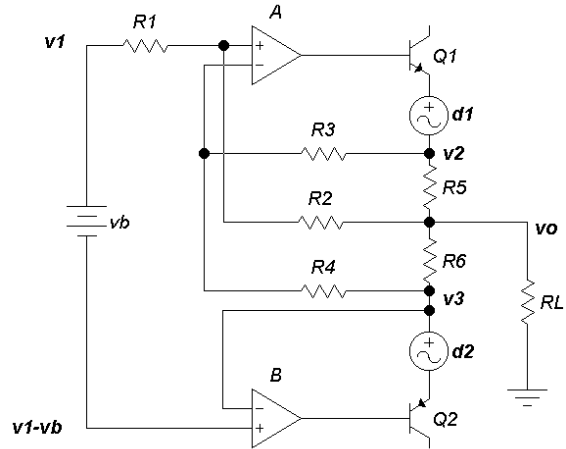


Fig.10. The amplifier with the op-amp based current source as the push-pull half in feedforward configuration. d_1 and d_2 - distortion, generated by the upper/lower half, e.g. in the power stages.

The voltage on the noninverting input of A amplifier is

$$v_1 \frac{R_2}{R_1 + R_2} + v_o \frac{R_1}{R_1 + R_2}$$

The voltage on the inverting input of A amplifier is

$$v_3 \frac{R_3}{R_3 + R_4} + v_2 \frac{R_4}{R_3 + R_4}$$

The voltage on the noninverting input of B amplifier is

$$v_1 - v_b ; v_b - \text{bias voltage}$$

The voltage on the inverting input of B amplifier is

$$v_3$$

Now for the output voltage for B amplifier one can get

$$v_3 = B(v_1 - v_b - v_3) + d_2$$

$$v_3 = \frac{B}{1+B}(v_1 - v_b) + d_2 \frac{1}{1+B}$$

and for A amplifier

$$v_2 = A \left(v_1 \frac{R_2}{R_1 + R_2} + v_o \frac{R_1}{R_1 + R_2} - v_3 \frac{R_3}{R_3 + R_4} \right) - A v_2 \frac{R_4}{R_3 + R_4} + d_1,$$

where d_1 and d_2 is the distortion, generated by the upper/bottom half, respectively.

Under assumption $R_1 = R_2 = R_3 = R_4$

$$v_2 = \frac{A}{2}(v_1 + v_o - v_2 - v_3) + d_1$$

$$v_2 = \frac{A}{2+A}v_1 + \frac{A}{2+A}v_o - \frac{A}{2+A}v_3$$

For the output summing network one can derive

$$\frac{v_2}{R_5} + \frac{v_3}{R_6} = v_o \left(\frac{1}{R_5} + \frac{1}{R_6} + \frac{1}{R_L} \right)$$

Under assumption $R_5 = R_6$

$$v_o = (v_2 + v_3)K,$$

where

$$K = \frac{\frac{2}{R_5}}{\frac{2}{R_5} + \frac{1}{R_L}} = \frac{1}{2} \frac{R_L}{R_L + \frac{R_5}{2}}.$$

(if $R_L \gg R_5$, then $K \approx \frac{1}{2}$)

$$v_o = K \left(\frac{A}{2+A}v_1 + \frac{A}{2+A}v_o - \frac{A}{2+A}v_3 + \frac{2}{2+A}d_1 + v_3 \right) =$$

$$= K \left(\frac{A}{2+A}v_1 + \frac{A}{2+A}v_o + \frac{2}{2+A}v_3 + \frac{2}{2+A}d_1 \right) =$$

$$= K \left(\frac{A}{2+A}v_1 + \frac{A}{2+A}v_o + \frac{2}{2+A} \left(\frac{B}{1+B}(v_1 - v_b) + d_2 \frac{1}{1+B} \right) + \frac{2}{2+A}d_1 \right) =$$

$$= K \left(\frac{2}{2+A} \left(\frac{A}{2} + \frac{B}{1+B} \right) v_1 + \frac{A}{2+A}v_o - \frac{2}{2+A} \frac{B}{1+B}v_b + \frac{2}{2+A} \frac{B}{1+B}d_2 + \frac{2}{2+A}d_1 \right)$$

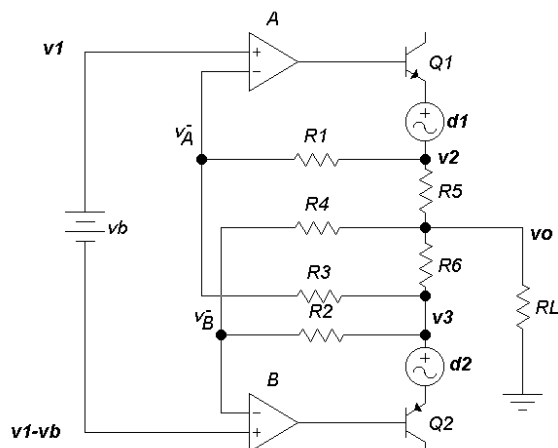
$$v_o = \frac{K(2+A)}{2+A(1-K)} \left(\frac{2}{2+A} \left(\frac{A}{2} + \frac{B}{1+B} \right) v_1 - \frac{2}{2+A} \frac{B}{1+B}v_b + \frac{2}{2+A} \frac{B}{1+B}d_2 + \frac{2}{2+A}d_1 \right)$$

From that equation it is evident that if $A \gg 1$, $B \gg 1$ then $v_o = v_1$, v_o is independent from v_b and d_1 is reduced accordingly with the loop gain $\frac{A}{2}$, while d_2 is reduced accordingly with the

product of loop gains $\frac{A}{2}$ and $\frac{B}{2}$ of the both halves.

In this circuit the feedforward error correction reduces distortion in the bottom highly nonlinear half.

The diagram of the improved circuit [4] is presented in Fig.11.



The voltage on the noninverting input of A amplifier is

The voltage on the inverting input of A amplifier is

$$v_3 \frac{R_1}{R_1 + R_3} + v_2 \frac{R_3}{R_1 + R_3}$$

 $v_1 - v_b$; v_b - bias voltage

The voltage on the inverting input of B amplifier is

$$v_3 \frac{R_4}{R_2 + R_4} + v_o \frac{R_2}{R_2 + R_4}$$

For the noninverting input of the bottom half the following equation can be written

$$\frac{v_B^- - v_3}{R_2} = \frac{v_o - v_B^-}{R_4},$$

$$v_\theta - v_3 = i_{R6} R_6,$$

where i_{R_6} is the current flowing through R_6 , v_B^- is the input voltage on the inverting input of B amplifier, $v_3 = B(v_1 - v_b - v_A^-)$.

Assuming $R_2 = R_4$, $B \gg 10$, $v_0 \approx v_1$ (v_0 is maintained nearly equal to v_1 due to the error correction in the upper half), the current

$$i_{R6} = -\frac{2v_b}{R_6}$$

will be nearly constant, thus provide the constant

Now one can put the equations for the output voltage for B amplifier

$$v_3 = B \left(v_1 - v_b - v_3 \frac{R_4}{R_2 + R_4} - v_o \frac{R_2}{R_2 + R_4} \right) + d_2$$

$$v_3 = \frac{B}{1 + B \frac{R_4}{R_2 + R_4}} \left(v_1 - v_b - v_o \frac{R_2}{R_2 + R_4} \right) + d_2 \frac{1}{1 + B \frac{R_2}{R_2 + R_4}}$$

and for A amplifier

$$v_2 = A \left(v_1 - v_3 \frac{R_1}{R_1 + R_3} - v_2 \frac{R_3}{R_1 + R_3} \right) + d_1$$

$$v_2 = \frac{A}{1 + A \frac{R_3}{R_1 + R_3}} \left(v_1 - v_3 \frac{R_1}{R_1 + R_3} \right) + d_1 \frac{1}{1 + A \frac{R_3}{R_1 + R_3}},$$

there d_1 and d_2 is the distortion, generated by the upper/bottom half, respectively.

Under assumption $R_1 = R_2 = R_3 = R_4$

$$v_3 = \frac{B}{2+B}(2v_1 - 2v_b - v_o) + d_2 \frac{2}{2+B}$$

$$v_2 = \frac{A}{2+A}(2v_1 - v_3) + d_1 \frac{2}{2+A}$$

Under assumption $R_5 = R_6$

$$v_o = (v_2 + v_3)K,$$

where

$$K = \frac{1}{2} \frac{R_L}{R_L + \frac{R_5}{2}}.$$

$$\begin{aligned}
v_o &= \left(\frac{A}{2+A} (2v_1 - v_3) + d_1 \frac{2}{2+A} + v_3 \right) K = \left(v_1 \frac{2A}{2+A} + v_3 \frac{2}{2+A} + d_1 \frac{2}{2+A} \right) K = \\
&= \left(v_1 \frac{2A}{2+A} + \frac{2}{2+A} \left(\frac{B}{2+B} (2v_1 - 2v_b - v_o) + d_2 \frac{2}{2+B} \right) + d_1 \frac{2}{2+A} \right) K = \\
&= \left(\frac{2}{2+A} \left(A + \frac{2B}{2+B} \right) v_1 - \frac{2}{2+A} \frac{2B}{2+B} v_b - \frac{2}{2+A} \frac{B}{2+B} v_o + \frac{2}{2+A} \frac{2}{2+B} d_2 + \frac{2}{2+A} d_1 \right) K \\
v_o &= \frac{K}{1 + K \frac{2}{2+A} \frac{B}{2+B}} \left(\frac{2}{2+A} \left(A + \frac{2B}{2+B} \right) v_1 - \frac{2}{2+A} \frac{2B}{2+B} v_b + \frac{2}{2+A} \frac{2}{2+B} d_2 + \frac{2}{2+A} d_1 \right)
\end{aligned}$$

From that equation one can clearly see that if $A \gg 1$, $B \gg 1$

then $v_o = v_1$, v_o is independent from v_b and

d_1 is reduced accordingly with the loop gain $\frac{A}{2}$,

while d_2 is again reduced accordingly with the product of loop gains $\frac{A}{2}$ and $\frac{B}{2}$ of the both halves.

It is shown, that there is additional reduction of distortion generated in bottom half d_2 for both circuits. The analysis was conducted including d_2 inside amplifier B feedback loop. Another option is to include distortion voltage v_d at the output of amplifier B . It will lead to the same result - the distortion is reduced in proportion to $\frac{A}{2}$ assuming exact resistor values. It is misleading to draw a conclusion that distortion is solely reduced by the product of the loop gains. One should keep in mind that getting precise values for the feedback resistors could have a far greater effect than even an order of magnitude increase in loop gain. As in real life the errors introduced by 1% tolerance components, it seems unpractical to obtain loop gain much more than 100.

Rather than repeat the entire calculation for the whole signals let us considered only the distortion voltage

v_d ($v_3 = v_d$) at the output of the class-B section, and put in resistor values with x error, i.e.

$R_3 = (1+x)R_1$. For the upper half the following equation can be written

$$\frac{v_2 - v_A^-}{R_1} = \frac{v_A^- - v_3}{R_3},$$

where v_A^- is the input voltage on the inverting input of A amplifier

$$v_2 = A(v_1 - v_A^-), \quad v_1 = 0, \quad v_3 = v_d$$

$$\frac{v_2 - \frac{v_2}{A}}{R_1} = \frac{\frac{v_2}{A} - v_3}{R_3}, \quad R_3 = (1+x)R_1.$$

The output voltage becomes

$$v_o \approx v_d + v_2 \approx \left(x + \frac{2}{A} + \frac{x}{A} \right) v_d.$$

It is clearly evident, that the error in resistor value x plays the same role as inadequate loop gain $\frac{A}{2}$. Here

A can be assumed frequency dependent and the appropriate frequency dependent x can be obtained for perfect null condition. Adjustment of resistor values can compensate for finite gain. For example if the upper half gain A is equal to 100 for the inverting input, then x should be -0.02 and the accurate

cancellation of v_d would be achieved. That can be done by reduction R_3 from $1 \text{ k}\Omega$ to 980Ω or by

R_1 increase from $1\text{ k}\Omega$ to $1020\text{ }\Omega$. Also, if A decreases with frequency, that can be compensated by a corresponding decrease in R_3 (capacitive shunting).

The same sort of equation can be obtained for the emitter/output summing resistors R_5, R_6 . That is because the summing resistors are located outside correction loop in feedforward error correction. This is not the case for feedback error correction, where the summing resistors are located inside the error correction loop.

APPENDIX 2

The Quad triples output stage with feedforward error correction is shown in Fig.12. The increase in output current lead to the increase in voltage drop across feedback resistors of the triple. This cause the decrease of the quiescent current i_q in class-A current mode amplifier, as shown in Fig13(b). At substantially high output current the output device switches off and the error correction mechanism then becomes inoperative (the trace 1 in Fig13(b)).

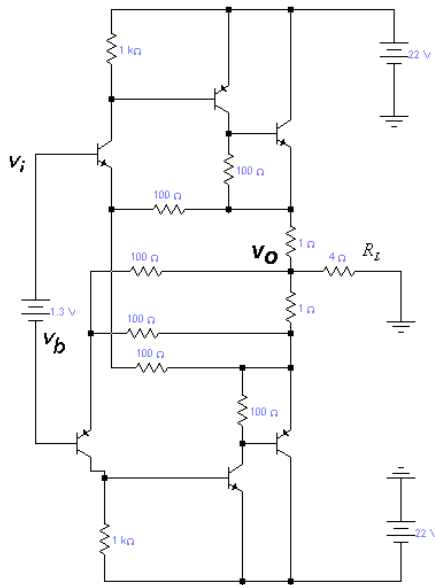
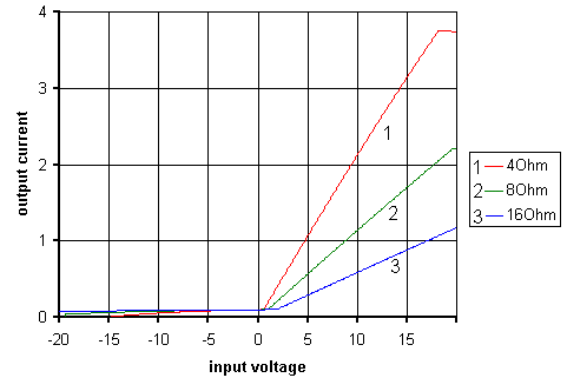
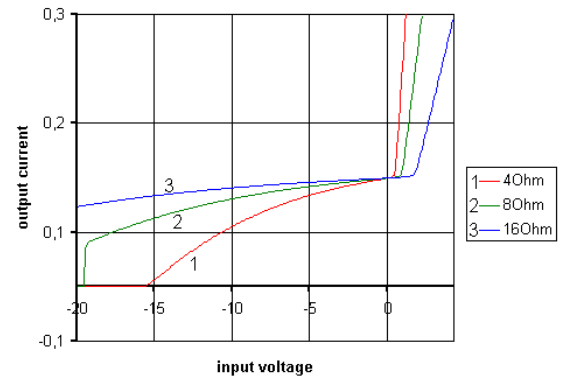


Fig.12. Triple configuration output stage with feedforward error correction.



(a)



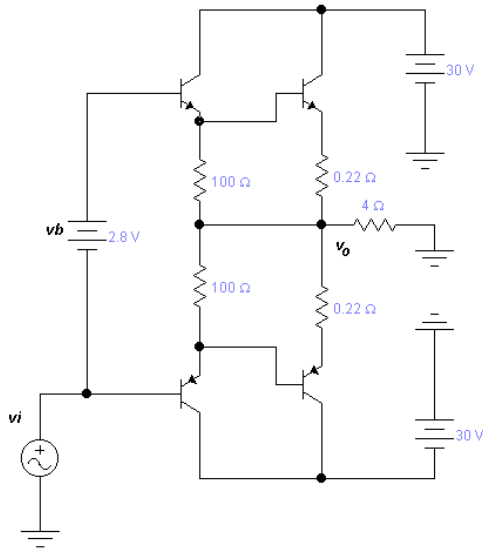
(b)

Fig.13. Collector current of the *n*pn output transistor versus input voltage v_i , with load resistance R_L as a parameter.

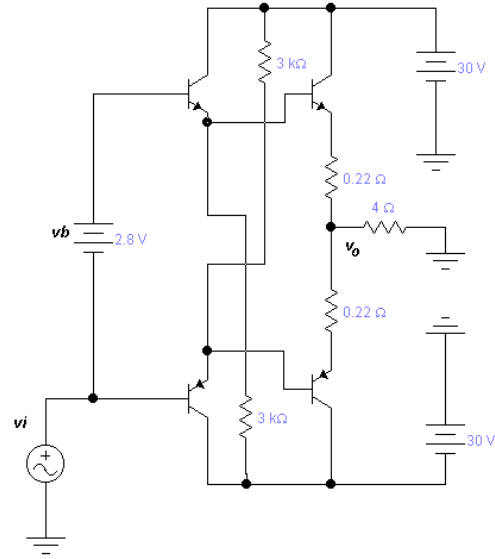
(b) – magnified region of negative input voltage. Note that the output transistor switches off with increasing load ($R_L = 4\text{ }\Omega$).

APPENDIX 3

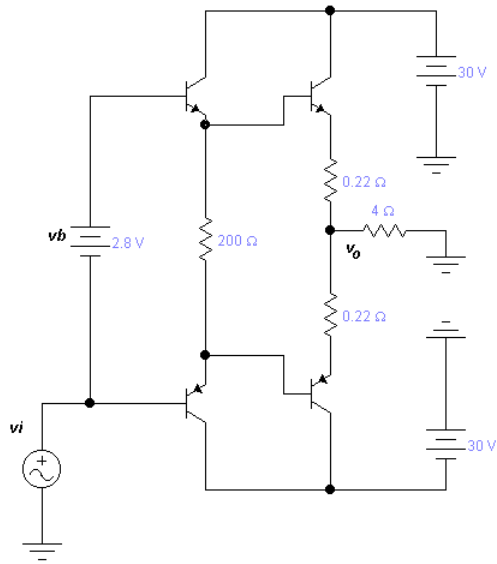
Three versions of the double-emitter follower output stage are shown in Fig.14. The small signal linearity of these three configurations is nearly identical and are presented in Fig.15. The Locanthi (Fig.14(b)) and Diamond (Fig.14(c)) circuits maintain the driver transistors in class A, as shown in Fig.17. These circuits are also capable to maintain voltage between bases/gates of the output devices at nearly constant level (see Fig.16). Without speed-up capacitor the stored charge in output transistors is removed in the Locanthi circuit two times faster than in Diamond circuit and one and a half times faster than in conventional one [6]. One should avoid conventional configuration (Fig.14(a)), as the driver transistors switch off (Fig.17(b), trace (a)) and add one more singularity to transfer function.



(a)



(c)



(b)

Fig.14. Three types of emitter follower output stage with different path for the driver transistors emitter current flow. (a) - with base-bleeder resistors, (b) - Locanthi circuit [16] and (c) - Diamond circuit [17].

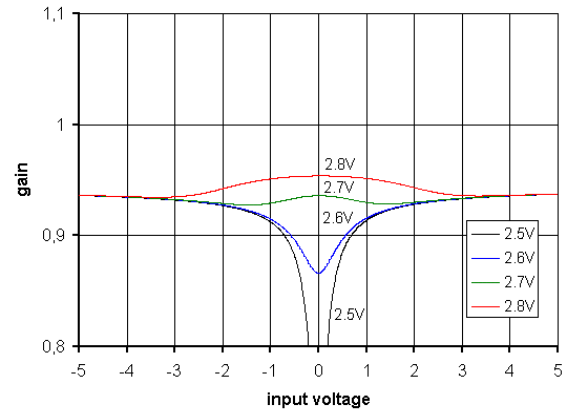


Fig.15. Emitter follower small-signal gain $\frac{dv_o}{dv_i}$ versus input voltage v_i , with bias v_b as a parameter.

Locanthi circuit (Fig.14(b)) driver transistor works in nearly constant-current nonswitching mode of operation (trace (b)).

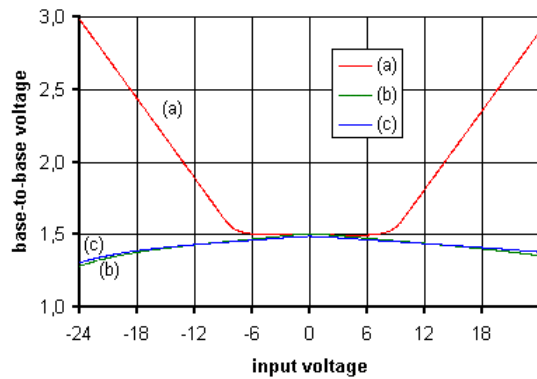


Fig.16. Base-to-base voltage of the *nnp* and *pnp* output transistors for the three types of emitter follower output stage, shown in Fig.14.

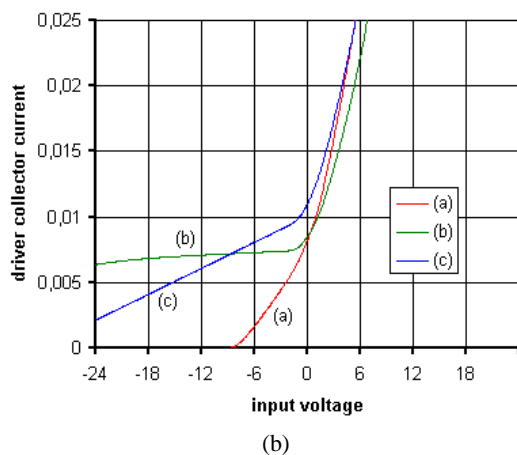
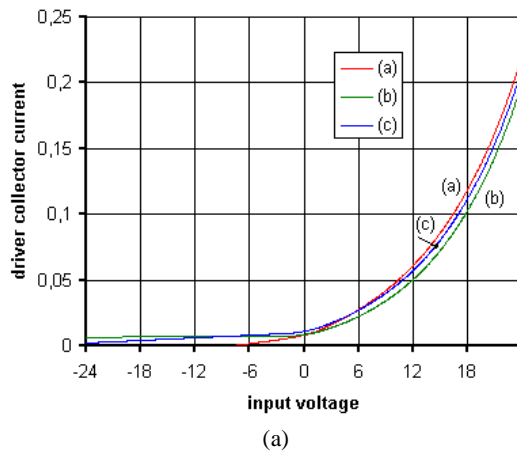


Fig.17. Collector current of the *nnp* driver transistor versus input voltage V_i for the circuits Fig.14. (b) – magnified region of negative input voltage. Note that the driver transistor of the conventional circuit (Fig.14(a)) switches off at the negative input voltages (trace (a)), while in