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A New 1kW Class-D Supported Linear Power Amplifier Employing a Self-Adjusting Ripple Cancellation Scheme

HANS ERTL, JOHANN W. KOLAR, FRANZ C. ZACH

Technical University Vienna, Power Electronics Section

Gußhausstraße 27 / 359.5, A-1040 Vienna / Austria
phone: +43-1-58801-3728 fax: +43-1-505 88 70
e-mail: ertl@ps1.iaee.tuwien.ac.at

Abstract — A power amplifier system consisting of a linear amplifier stage with a class-D (switching) current injection stage arranged in parallel is presented. With this topology the fundamental drawback of conventional linear power amplifiers – the high losses – is avoided. Compared to pure class-D switching amplifiers the described switched-mode assisted power amplifier does not require an output filter to suppress the switching frequency components of the output voltage. Such a filter would generally deteriorate the dynamic response and the output impedance of the switching amplifier.

Concerning the output voltage the presented topology can be considered as a master-slave system with a guiding linear amplifier and a supporting class-D slave unit. (However, it has to be pointed out that related to the throughput power the class-D stage acts as the master.) The remaining losses of the linear stage are caused by the current ripple of the class-D stage. Therefore, a further reduction can be achieved by realizing the current injection stage using a parallel connection of two partial systems operated in a phase-shifted manner. As the paper shows, this gives a significant improvement of the efficiency especially for small output voltages.

1. Introduction

Class-A or class-B power amplifiers are replaced by switching (class-D) amplifier topologies in an increased quantity to avoid the high power losses of conventional linear power stages. Nevertheless, if high-quality output voltages are required (i.e., low amplitudes of distortion components and a low output impedance of the system, e.g., for test and measuring equipment) switching amplifiers show significant limitations. The output voltage of a class-D amplifier contains switching frequency noise of substantial amplitude which has to be reduced by a carefully designed low-pass filter. However, this filter – which has to be a higher order type – in general reduces the dynamic response and increases the

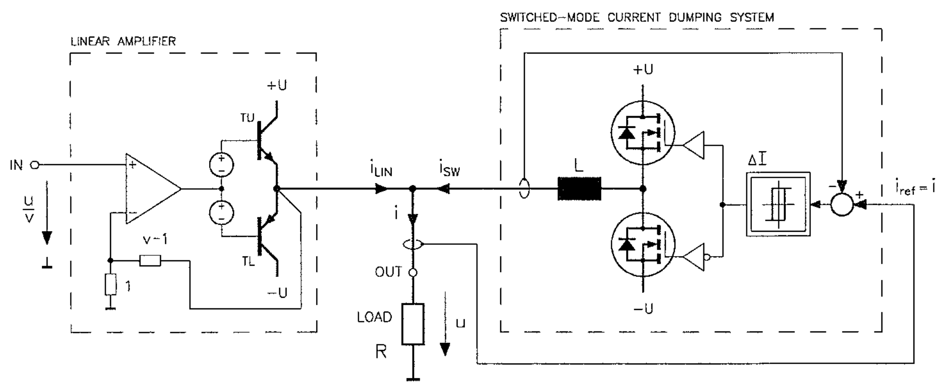


Fig.1 Basic circuit topology of a switched-mode assisted linear power amplifier system.

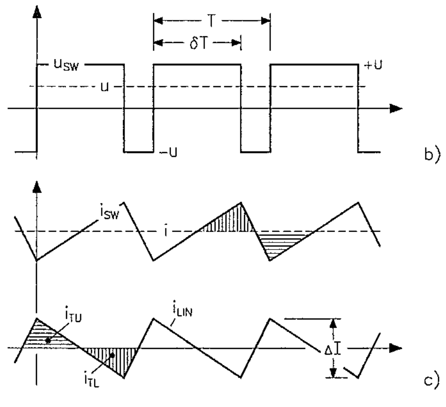
output impedance of the amplifier. Furthermore, the distortions caused by the interlock delay of the switching transistors and/or by a ripple of the DC supply voltage are not lowered by the output filter (which is designed to reduce switching-frequency components) and have to be reduced using a proper control of the system [1, 2].

To avoid the disadvantages described above, in [3] a concept was proposed (based on an idea presented originally in [4]) formed by a parallel arrangement of a conventional linear amplifier stage and a class-D switching current injection system (Fig.1). With this, the output filter of the switching stage is reduced to a single coupling inductance. The basic idea of this switched-mode assisted linear amplifier system is that considering the output voltage the linear amplifier acts as the guiding master whereas the task of the class-D slave stage is to take over the current of the linear stage. In the ideal (stationary) case the linear power amplifier then only has to deliver the current ripple of the switching stage. By this, the losses of the linear stage are significantly reduced as compared to a conventional class-A or class-B power amplifier. The linear amplifier therefore can be considered as active filter which compensates the switching frequency ripple and the interlock-delay disturbances of the switching stage. Contrary to a passive output filter of a conventional switching amplifier the linear amplifier of the proposed concept also reduces low-frequency distortions and subharmonic frequency components.

2. Switched-Mode Assisted Linear Power Amplifier Basic Principle of Operation – System Control

As shown in Fig.1, the control of the class-D part is realized by a current controller with a reference value being identical to the load current. Therefore, the linear stage only has to deliver the control error and the ripple current of the switching stage which would significantly reduce its losses. (Instead of the explicit subtraction of the reference value (load current i) and the actual value (class-D stage output current i_{SW}) as shown in Fig.1, in the case at hand this could also be done in an implicit manner by direct measurement of the linear stage output current i_{LIN}.) In the simplest case the current control mentioned before can be performed using a hysteresis type controller which would result in a not constant switching frequency within the fundamental period of the amplified signal. Alternatively many other current control schemes known from switched-mode power supplies (e.g., average current control [5]) could be applied. There, the usage of a pulse width modulator allows a switching frequency being constant which is, however, of not essential importance for this application due to the described active filtering approach of the proposed system. An advantage of the hysteresis controller is its inherent overmodulation ability which yields a more efficient utilization of the DC supply voltage $\pm U$. On the other side, PWM current controllers with their well defined switching instants allow an easier extension of the class-D current injection stage to a parallel arrangement being operated in an optimum phase shifted manner in order to reduce the total ripple current and to improve the efficiency of the system. However, as described in section 5 for two parallel converter branches a similar behavior can be achieved also applying hysteresis current controllers and a self-adjusting ripple cancellation scheme.

In the following, at first the losses of a hysteresis-controlled switched-mode assisted amplifier system with a single switching stage shall be calculated for constant width ΔI of the tolerance band.



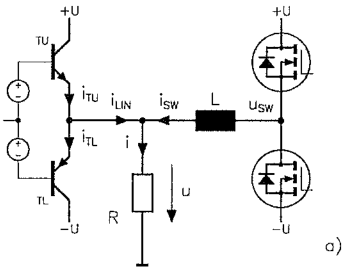


Fig.2 Voltage and current wave shapes of a switched-mode assisted linear power amplifier; (a) circuit diagram; (b) output voltage; (c) current wave shapes (for class-B-mode).

For this it is assumed that the switching frequency is sufficiently high as compared to the output signal frequency so that load current i and output voltage u can be treated as being constant within the switching interval T. Furthermore, all power semiconductor devices are assumed to be ideal (neglection of delay times, on-state voltage drops etc.). Also, all supply voltage variations are neglected. Normalizing the output voltage u to the DC supply voltage we define a time-varying modulation coefficient $m = u/U$ ($m = -1...+1$). Applying this definition we get for the switching frequency of the system

$$f_S = f_{S,max} \cdot \left(1 - m^2\right) \qquad \text{with} \qquad f_{S,max} = \frac{U}{2L \cdot \Delta I}. \tag{1}$$

As simplification for the calculation of the power losses of the linear stage we assume a pure class-B operating mode (no quiescent current for reduction of the crossover distortions). With this, as shown in Fig.2, each single power transistor of the linear stage has to deliver half of the current i_{LIN}. The local average value (averaging within the pulse period T) therefore becomes $i_{TU,avg} = i_{TL,avg} = \frac{1}{8}\Delta I$ due to the triangular shape of the currents. For the power losses of the transistors we finally receive

$$p_{TU} = U \cdot \Delta I \tfrac{1}{8}(1-m) \qquad p_{TL} = U \cdot \Delta I \tfrac{1}{8}(1+m) \qquad p_T = p_{TU} + p_{TL} = \tfrac{1}{4} U \cdot \Delta I = P_T. \tag{2}$$

The total power losses of the linear stage are not dependent on the output voltage and remain constant for any output voltage wave shape and amplitude.

According to Eq.(2), the demand for low power losses and high system efficiency implies a small ripple current amplitude ΔI. However, for a given maximum switching frequency $f_{S,max}$ defined by the semiconductor devices of the switching stage this would suggest the usage of a higher inductance value L of the coupling coil. On the other side, a higher value of L reduces the large signal power bandwidth of the current injection stage. Although this does not limit the power bandwidth of the total system in principle, it should be pointed out that the desired reduction of the losses of the linear stage only can be achieved for frequencies below the bandwidth of the class-D stage. However, the pulse response of the total system is basically not determined by the slew rate of the switching stage if the linear stage is designed to provide transient current pulses of rated amplitude. In this case the current control error leaves the tolerance band in a transient manner (see simulation results presented in section 4).

3. Design of the Linear Amplifier Stage

A very low high-frequency output impedance of the linear stage is of fundamental importance for a low noise level of the output voltage. The switching frequency noise of the output voltage of the whole system depends directly on the output impedance of the linear amplifiere. Therefore, the design of the linear amplifier system not only has to be done with regard to the aimed useful (low-frequency) bandwidth of the whole system, but also under consideration of the switching-frequency (high-frequency) output impedance. In general, the inner impedance of the linear amplifier output stage (a source- or emitter-follower circuit) has to be lowered to the mΩ-range using a feedback loop with a proper loop gain. A detailed analysis of this problem area is omitted here for the sake of brevity and can be found in [3] and [6].

4. Dimensioning Example – Simulation Results

A prototype system of a 1kVA amplifier system with the nominal values $U = \pm 80V$ (DC link voltage), $R = 2.5\Omega$ (load resistance); RMS-value of the output voltage: 50V, $f_B = 10kHz$ (useable power bandwidth), $f_{S,max} = 200kHz$ (maximum switching frequency) shall be calculated briefly. For the desired bandwidth we calculate the inductance of the coupling coil: $L = R/\omega_B = R/(2\pi f_B) = 2.5\Omega/(2\pi \cdot 10kHz) = 40\mu H$. Using Eq.(1) the tolerance band for the hysteresis current controller $\Delta I = U/(2L \cdot f_{S,max}) = 80V/(2 \cdot 80\mu H \cdot 200kHz) = 5A$ can be determined. From Eq.(2) we receive for the power losses of the linear amplifier stage $P_T = \frac{1}{4} U \cdot \Delta I = \frac{1}{4} 80V \cdot 5A = 100W$.

These losses are far beneath the losses of conventional linear power amplifiers, especially for the case of non-resistive or non-linear loads. However, it has to be mentioned that the losses calculated before do not include the losses of the switching stage. However, the efficiency of switched-mode bridge topologies usually lies far above 95% so that the total losses of switched-mode assisted amplifiers would not be increased significantly as compared to conventional linear amplifiers.

The current wave shapes of the simulated 1kW amplifier system are given in Fig.3 for sine-wave ($m = M \cdot \sin \omega t$, $M = \hat{U}/U$, $M = 0 \ldots 1$) and pulse response. There, the pulse response clearly shows the

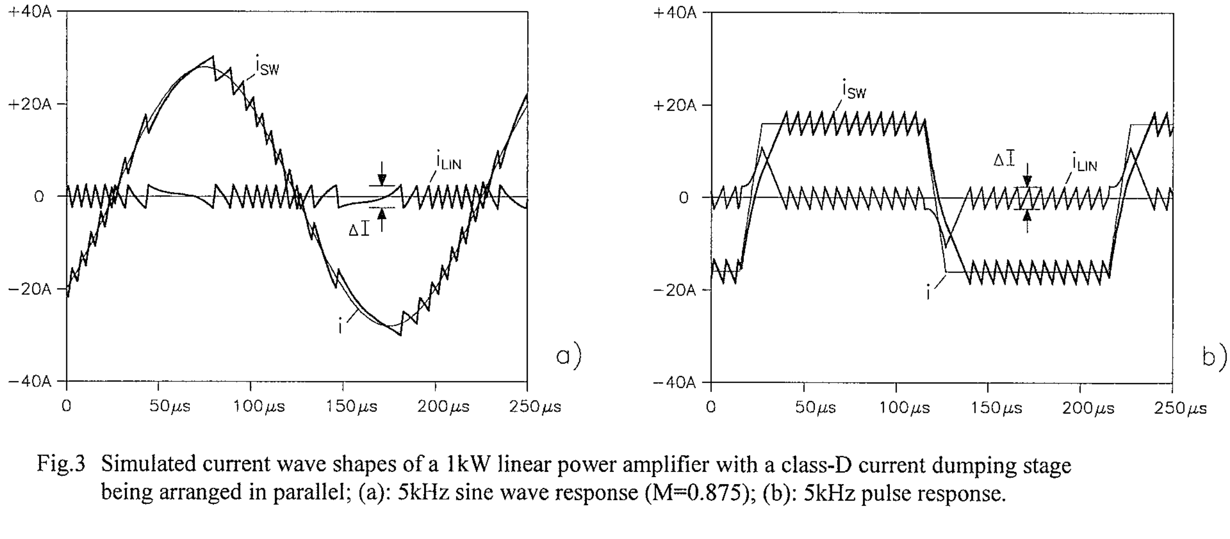


Fig.3 Simulated current wave shapes of a 1kW linear power amplifier with a class-D current dumping stage being arranged in parallel; (a): 5kHz sine wave response (M=0.875); (b): 5kHz pulse response.

limited slew-rate of the switched-mode current injection system. For higher du/dt-rates the output current of the linear amplifier i_{LIN} not only has to compensate the ripple of the switching state but also has to take over the dynamic current peaks (i_{LIN} therefore cannot be guided completely within the tolerance band ΔI). This effect results in increased power losses of the linear stage.

5. Phase-Shifted Current Injection Stage

As stated by Eq.(2) the linear amplifier power losses P_T caused by the ripple of a single class-D current injection stage described so far remains constant for the entire output voltage area. This would result in a relatively poor efficiency of the amplifier system for output signals of small amplitudes as will be shown in the following section (cf. Fig.7c, *SMAL-1*). This fundamental drawback can be avoided using a current injection stage where the ripple becomes smaller for lower values of the output voltage u. (Unfortunately a two-level converter branch supplied with a DC voltage of $\pm U$ shows the highest ripple current for $u = 0$ due to the fact that zero output voltage has to be generated as average value of positive and negative pulses with an amplitude determined by the supply voltage U.) Using a three-level converter topology it would (theoretically) be possible to supply $u = 0$ with zero current ripple but the control unit for a three-level branch causes a relatively high realization effort (four isolated gate drive circuits).

As an alternative to three-level structures a current injection stage consisting of 2 two-level converter branches connected in parallel can be applied. As mentioned, a two-level branch shows a high current ripple amplitude for $u \approx 0$ but if two branches are operated in opposite phase, a cancellation of the ripples of the partial systems results and the total ripple (which determines the losses of the linear stage) becomes rather small. A system where the current ripples are phase-shifted in an optimum manner can, e.g., be realized by using a current controller with two underlying voltage-mode PWM stages operated in phase-opposition. But in this case special precautions have to be taken to guarantee the current sharing between the two branches. Alternatively, in [7] a concept has been proposed which allows a synchronized operation of two hysteresis-controlled converters arranged in parallel in a sub-optimum manner by cross-coupling of the current measurement signals (cf. Fig.4) in

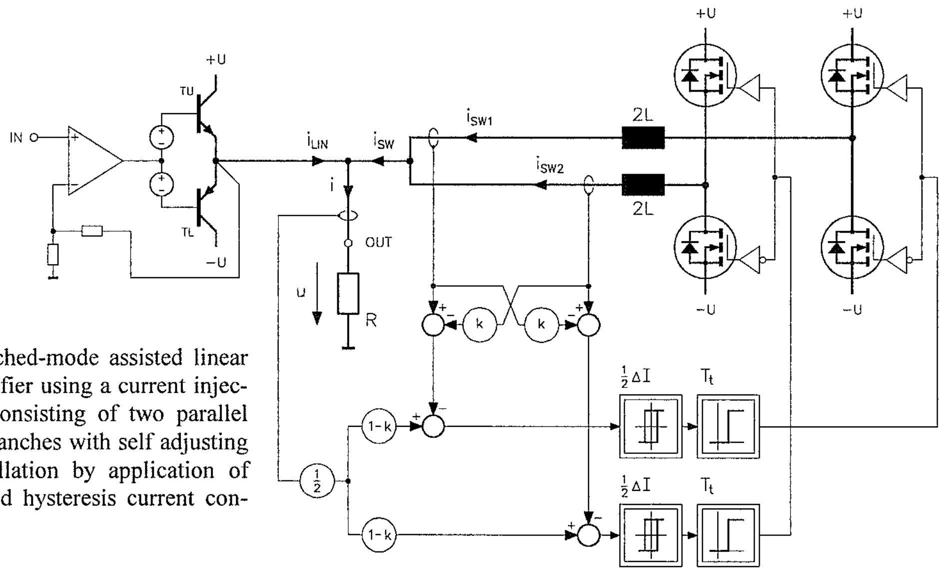


Fig.4 Switched-mode assisted linear power amplifier using a current injection stage consisting of two parallel converter branches with self adjusting ripple-cancellation by application of cross-coupled hysteresis current controllers.

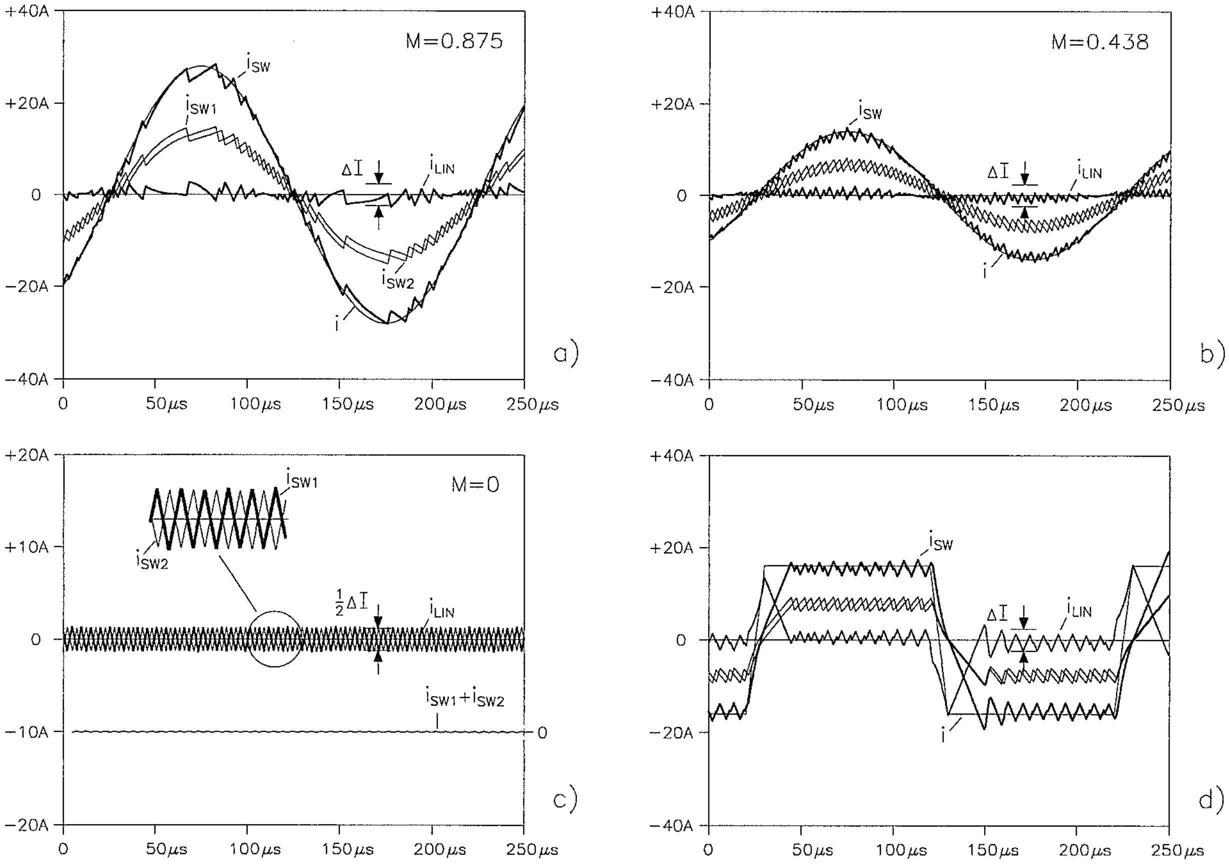


Fig.5 Simulated current wave shapes of a 1kW linear power amplifier with cross-coupled hysteresis controlled dual-stage class D current injection; (a): 5kHz sine wave response (M=0.875); (b): as (a) but M=0.438; (c): zero output voltage (total ripple cancellation); (d): 5kHz pulse response.

connection with a small control signal delay T_t. The reference value of each branch is half the load current i. The effect of the cross-coupling on the reference value has to be corrected by a coefficient $1-k$ in the reference value signal path. Remark: This dual-stage system is dimensioned in the way that power bandwidth and maximum switching frequency equals to the respective specifications of the single-stage concept (i.e., $L \rightarrow 2L$, $\Delta I \rightarrow \frac{1}{2}\Delta I$, cf. Fig.4).

The simulation results for this system are presented in Fig.5 for different values of the modulation indices M (coupling coefficient $k = 0.3$, delay time $T_t = 0.5\mu s$) according to Fig.3 of the single-stage system. As described in [7], the proposed control method provides a phase shift of the ripple of the two partial currents i_{SW1} and i_{SW2} which is dependent on the duty-cycle δ of the converter. For $\delta \approx 0.5$ (which for the analyzed structure is given for $u \approx 0$) we receive an optimum phase-shift of the two ripple currents as can be seen from Fig.5c. For output voltages of higher amplitude the ripple-cancellation becomes suboptimal, but in any case the total ripple of the dual-stage current injection system will be lower as compared to the single-stage system.

6. Comparison of Losses

In this section the losses of the dual-stage switched-mode assisted amplifier shall be calculated in order to get information concerning the improvement of the system efficiency as related to a system

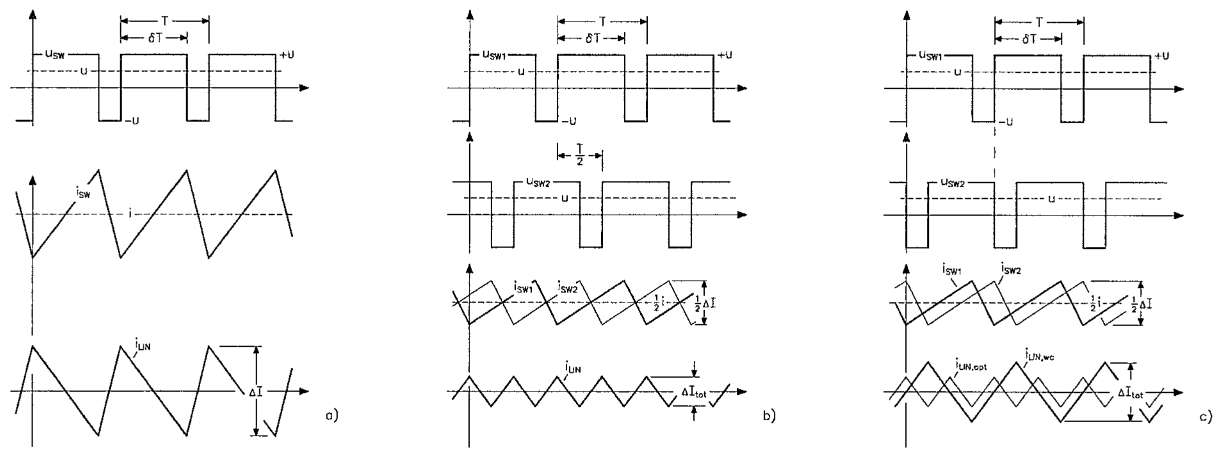


Fig.6 Voltage and current wave shapes of a single-stage class-D current injecting system (a) and for dual current injecting systems with optimum (b) and worst-case (c) ripple current cancellation.

with a single-stage current injection. Due to the relatively involved operating behavior of the applied self-adjusting ripple cancellation (cf. [7]), the analysis is limited to (i) *optimum current cancellation* and (ii) *sub-optimal current cancellation* and to pure sinusoidal modulation $m = M \cdot \sin \omega t$ for the sake of brevity.

Optimum Current Cancellation. This will be achieved if the system operates in the mode where both output voltages of the converter branches are shifted by half of a switching period (cf. Fig.6b). Omitting the details of the calculation we get for the amplitude ΔI_{tot} of the total ripple of the two superimposed inductor currents as compared to the ripple amplitude ΔI of a single-stage approach

$$\frac{\Delta I_{tot}}{\Delta I} = \frac{m}{m+1} \quad \text{for} \quad m = 0 \ldots +1 \qquad \text{and} \qquad \frac{\Delta I_{tot}}{\Delta I} = \frac{m}{m-1} \quad \text{for} \quad m = -1 \ldots 0 . \tag{3}$$

Calculating the losses of the transistors of the linear amplifier in a similar way as performed in section 2 we finally receive

$$P_T = \bar{p}_T = \tfrac{1}{4} U \cdot \Delta I \cdot \frac{1}{\pi} \int_0^{\pi} \frac{M \cdot \sin(\omega t)}{M \cdot \sin(\omega t) + 1} d\,\omega t = \tfrac{1}{4} U \cdot \Delta I \cdot \left[1 - \frac{1}{\sqrt{1-M^2}} \left(1 - \frac{2}{\pi} \arctan \frac{M}{\sqrt{1-M^2}} \right) \right] . \tag{4}$$

Due to this equation, the losses of the idealized dual-stage system (shown in Fig.7a or Fig.7b, denoted by *SMAL-2*$_{opt}$) contrary to a single-stage system (*SMAL-1*) are dependent on the modulation index M and tend to zero for $M \rightarrow 0$. Even in the case of $M = 1$ the losses amount to only about 36% of the losses of a single-stage system. Due to this operating behavior the efficiency in the area of low output voltage amplitudes is significantly increased (cf. Fig.7c).

Sub-Optimal Current Cancellation. According to [7] the proposed system shows a minimum phase-shift (giving the worst-case condition concerning the current ripple cancellation) which can be defined in a first step by linking the intervals $u_{SW} = -U$ for $m = 0 \ldots 1$ (or $u_{SW} = +U$ for $m = -1 \ldots 0$) of both converter stages as shown in Fig.6c). As a detailed analysis shows, the amplitude of the total current ripple in this case is exactly twice the value of the optimal case determined by Eq.(4) (cf. Fig.7 *SMAL-2*$_{wc}$).

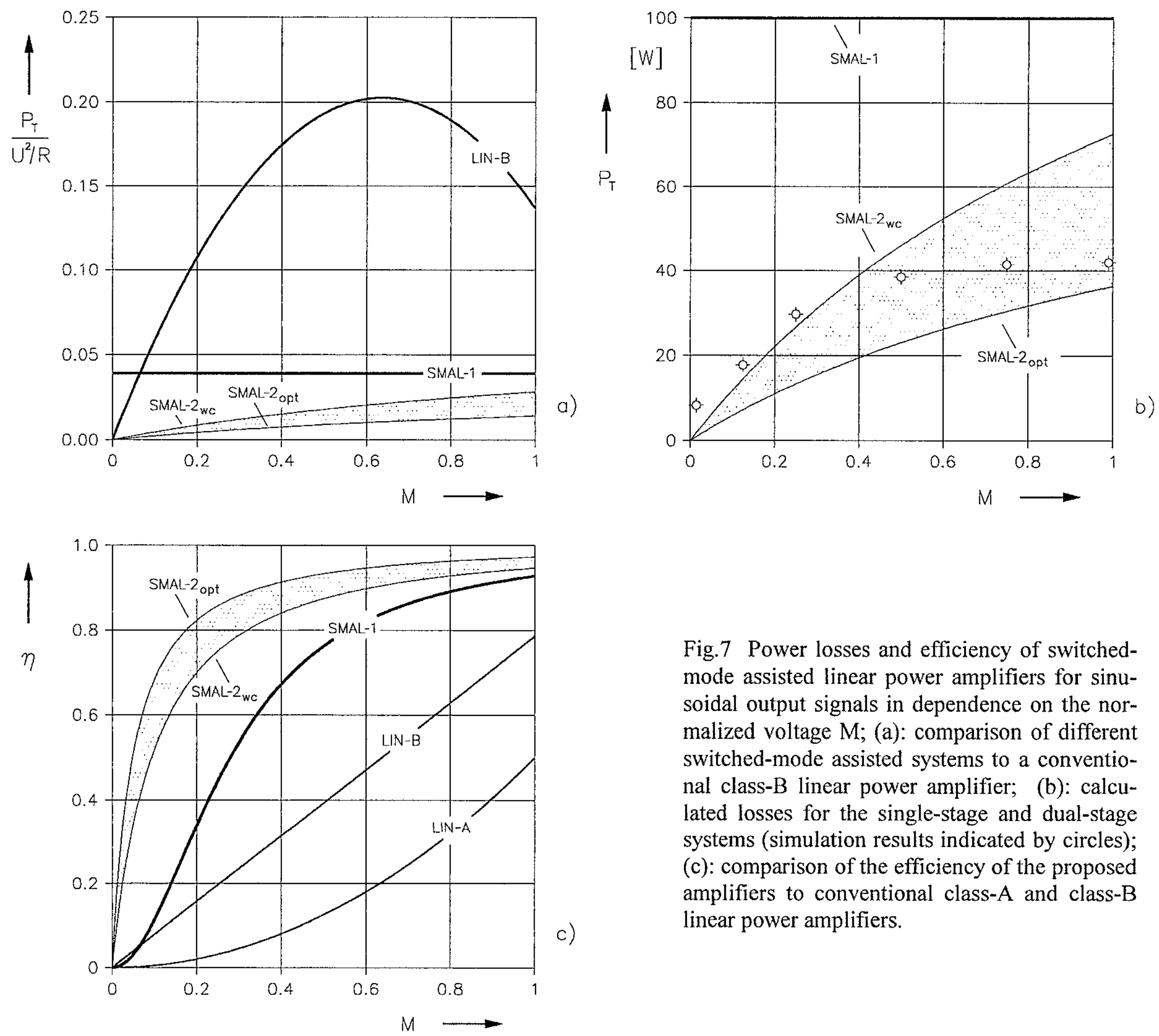


Fig.7 Power losses and efficiency of switched-mode assisted linear power amplifiers for sinusoidal output signals in dependence on the normalized voltage M; (a): comparison of different switched-mode assisted systems to a conventional class-B linear power amplifier; (b): calculated losses for the single-stage and dual-stage systems (simulation results indicated by circles); (c): comparison of the efficiency of the proposed amplifiers to conventional class-A and class-B linear power amplifiers.

The operating behavior of the real system lies between the two border lines calculated before (in Fig.7 indicated by the dotted areas) which is verified by digital simulation for the proposed system (Fig.7b). For low output voltage amplitudes the simulation results even are slightly above the calculated sub-optimal border line which is unexpected because for low output voltages the ripple current cancellation should appear in a rather optimal manner. However, the delay-time T_t (which is essential for the self-adjusting ripple cancellation) results in a higher actual ripple amplitude than defined by the tolerance bandwidth of the hysteresis controller. Therefore, the switching frequency is lower as compared to the single-stage system where no switching delay is introduced. For a more accurate comparison of the systems we would have to lower the width the tolerance band of the dual-stage current controllers to receive switching losses being identical to the single stage system. By this, the simulation results would come closer to the expected optimum curve. Furthermore, T_t is also the reason for the small fundamental frequency component of the control error as demonstrated (for a single-stage system) by [4] and can be seen also in Fig.5a or Fig.5b. The "saturation" of the simulated losses for $M \approx 1$ (cf. Fig.7b) is caused by the not triangular ripple current wave shape resulting due to operating the system near the border of the voltage overmodulation range.

The essential advantage of the dual-stage switched-mode assisted amplifier is shown clearly by the efficiency diagram Fig.7c. Due to the ripple-cancellation as compared to the single-stage approach we get a far higher efficiency especially for low output voltage amplitudes.

7. Future Development

As described in section 2 the trade-off between power-bandwidth and losses gives a basic limitation of a switched-mode assisted linear power amplifier. A possibility to improve this behavior (besides the application of the described multi-stage class-D current injection systems) is to substitute the coupling inductance L by an impedance of higher-order (e.g., $L \to LCL$). A further characteristic of the presented system is that the linear stage has to cover the full output voltage range which implies several difficulties if output signals of higher voltage amplitudes have to be generated. Both drawbacks can be avoided applying the structure shown in Fig.8.

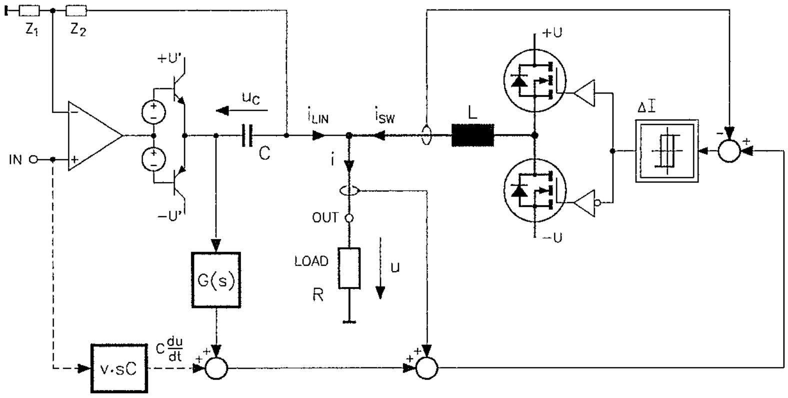


Fig.8 Schematic circuit diagram of a switched-mode assisted linear amplifier with improved efficiency formed by capacitive blocking of the output voltage u.

The topology consists of a class-D-stage similar to the basic switched-mode assisted amplifier shown in Fig.1 with an LC-output filter. Contrary to conventional filters one terminal of the filter capacitor C is not grounded but connected to the output of a linear amplifier. The signal to be amplified appears at the input of this amplifier. The feed-back loop of the linear system again (as the circuit of Fig.1) senses the total output voltage. As in the case of the conventional switched-mode assisted amplifier the linear stage has to sink the current ripple of the class-D switching stage, but now the filter capacitor takes over a large amount of the output voltage (the capacitor blocks off the output signal u from the linear amplifier). This results in a relatively small voltage at the output of the linear amplifier, so that this stage can be supplied with a relatively low (as compared to the supply voltage U of the class-D system) DC voltage $\pm U'$ which reduces the losses of the system significantly.

8. Conclusions

The presented system allows a significant improvement of the efficiency of linear power amplifiers without reduction of the high output voltage quality being characteristic for this type of amplifiers. As opposed to class-D switching amplifiers the proposed switched-mode assisted linear system does not require any output filtering which reduces the dynamic performance of conventional class-D amplifier systems. Also, an output filter would take influence on the total output impedance and would impair the desired voltage source characteristic of the amplifier system. The switched-mode assisted linear approach yields a much better and robust (load independent) guidance of the output voltage than a conventional class-D system with output filter and is, therefore, an ideal concept for the realization of a high-quality voltage source as being used for test and measuring purposes.

A further increase of the efficiency (especially for output voltages of small amplitudes as compared to the supply voltage) can be achieved by using a switched-mode assisted amplifier with a dual-stage current injection system operated in a ripple current cancellation mode. The required phase-opposition of the two stages which is necessary for an efficient ripple current cancellation can be provided in a very simple way applying two cross-coupled hysteresis current controllers.

Further research will be on an extension of the proposed system by using a capacitive coupling of the linear stage to the load. There, the output voltage swing of the linear amplifier is limited to small values which results in a further reduction of the power losses as compared to the system analyzed in this paper.

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