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Linear Power Amplifiers: Basic Considerations of Switched-Mode Assisted Amplifiers

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Abstract - The paper presents a combined high efficient amplifier system consisting of a linear amplifier unit with a switched-mode(class D) current and voltage stage arranged in parallel. With this topology the fundamental drawback of conventional linear power amplifiers - the high loss is avoided. Compared to a pure class D (switching) amplifier the presented system needs no output filter to reduce the switching frequency harmonics. This filter (usually of cascaded type) generally try to avoide the transient response of the system and impairs the feedback loop design. Furthermore, the low-frequency distortions of switching amplifiers caused by the inter system response delay of the power transistors is avoided with the presented switched-mode assisted linear amplifier system. This consideration as a master-slave system with a guiding linear amplifier and a supporting class D slave unit. The work describes the operating principle of the system, analyzes the fundamental relationships for the circuit design and presents simulation results with the help of virtual lab.

Keywords - class D, harmonics, transient error, etc

I. INTRODUCTION

Conventional linear power amplifiers are replaced by switching (class D) amplifiers in an increased quantity to overcome the essential drawback of linear amplifier systems[l]. The output voltages of a class D amplifier imply a switching frequency component (harmonics) which may be reduced by a proper filtering technique and circuitry. However, this filter - which has to be in general of higher order type - reduces the dynamic response and increases the output impedance of the whole amplifier system. Also, the interlock delay time of the usually applied bridge topologies, a ripple of the DC supply voltage +V and the on-state voltages of the power semiconductor devices, transistor and freewheeling diode may result in low-frequency distortion [2] which hardly can be reduced by the described switching frequency output filter but has to be lowered by using a special control loop design [3, 4]. A further problem of switching amplifiers is the possible occurrence of harmonic frequency components which may result for a small signal-to switching-frequency ratio or if a pulse width modulation strategy with variable switching frequency is applied. This harmonic noise basically cannot be lowered by the output low pass filter because the frequency components lie within the power bandwidth of the amplifier. A concept is proposed consisting of a parallel arrangement of a class D switching system and a conventional linear amplifier stage.

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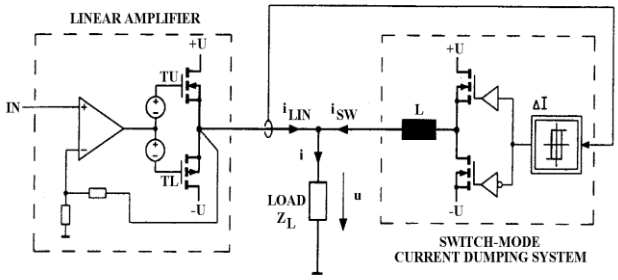

Fig. 1 Simplified circuit diagram of a linear power amplifier

The output filter of the switching amplifier is reduced to a single coupling inductor determining the switching frequency ripple. Although the linear amplifier can be considered as active filter which compensates the switching frequency ripple and the modulation noise, the basic idea of the proposed switched-mode assisted linear amplifier is that the linear amplifier acts as the guiding master system whereas the task of the class D (slave) stage is to take over the current of the linear stage (current dumping). In the ideal (stationary) case the linear power amplifier only has to deliver the ripple of the class D stage which significantly reduces its power losses. Contrary to a passive output filter of a conventional switching amplifier the linear amplifier of the proposed concept also reduces low-frequency distortions and harmonic components. However, a very low output impedance of the linear system part is importance to get a high noise rejection. This performance is to be considered by an appropriate design of the linear amplifier circuitry and feedback system. Furthermore, the switched mode assisted linear amplifier allows a significant reduction **as** an idealized class D amplifier. Therefore, concerning the losses the proposed system can be seen as intermediate solution between pure linear and pure class D power amplifiers.

II. SYSTEM CONTROL - CALCULATION OF POWER LOSSES

The guidance of the class D part is realized by a current controller whose reference value is identical to the current through the load**.** Thus, only the control error and the ripple is delivered by the linear stage. Instead of an explicit subtraction of reference value (load current i) and actual value (class D stage output current is**,**) the calculation of the controlling quantity can be done in an implicit manner by direct measurement of the linear stage output current . As an alternative, a pulse width modulator (PWM) with a superimposed linear current controller or other types of modulation is to maintain the harmonics develop in the master and slave and reduced by the low pass filter current controllers being well known from switched-mode power supplies (e.g., conductance control) can be applied.

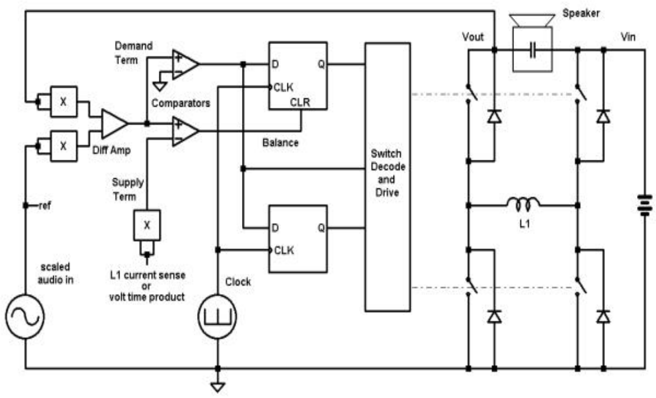

Fig.2 Circuit diagram of a switched-mode assisted linear power amplifier.

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The pulse width modulator allows a switching frequency being constant which is, however, of not essential significance for this application as stated before. An advantage of the hysteresis controller is it's inherent over modulation ability which yields a more efficient utilization of the DC supply voltage *V. On the other side, PWM current controllers with their well defined switching instants allow an easier extension of the class D stage to a parallel arrangement being operated in an optimum phase shifted manner in order to reduce the total ripple current, or increase the effective switching frequency, respectively. However, it should be mentioned that there exist solutions for two hysteresis controlled converter branches (arranged in parallel) where a suboptimal phase shift can be achieved in a very simple way. In the following, the losses of the linear amplifier stage shall be calculated for the case that a hysteresis current controller with a constant tolerance band is applied. It is assumed that the load current I and the output voltage V can be treated as constant within the switching interval T or, that there exists a sufficient signal-to-switching frequency ratio, respectively (Fig.3). '**Switching Frequency**

The output voltage U (averaged within a pulse interval T) is determined by the duty cycle. According to V, = Ldi,,/dt the switching frequency f, = 1/T can be calculated.

Power Losses

The power losses of the linear stage depend on its operating mode, where one has to distinguish between class A (linear amplifier with quiescent current eliminating crossover distortions) and class B (without quiescent current) mode. the linear stage where it is assumed that for class A mode the quiescent current is as small as possible, the class A mode losses are twice the losses of the class B mode.

Frequency effect on the Amplifier Bandwidth.

The demand for low power losses implies a small ripple amplitude . However, for a defined maximum switching frequency f, this would result in the usage of a high value of the inductance L. On the other side, a higher value of L reduces the power bandwidth of class B of the switched-mode current dumping stage. If we normalize gain bandwidth with respect to the value V/R (maximum load current, resistive load E_L = R assumed), kA = AI/(V/R), we receive for a class-B linear stage (kA ... normalized ripple amplitude). The power bandwidth of the current dumping stage can be defined as f_B = R/ (2nL) .

The switching frequency to- bandwidth ratio is linked to the losses of the linear system. For a given maximum switching frequency and a required power bandwidth of the whole amplifier the current ripple (and therefore the power losses) is fixed. However, there are some possibilities to overcome this fundamental limitation:

(1) Usage of a higher supply voltage for the switching

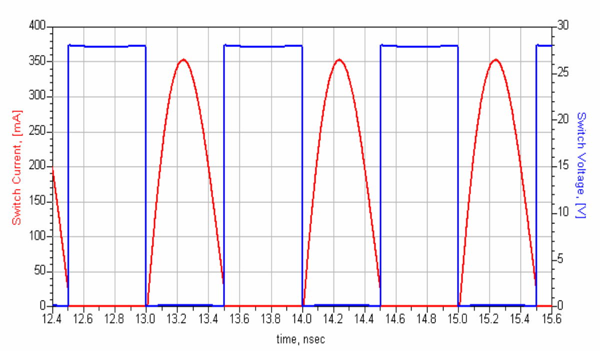

Fig.3 Voltage and current waveforms of a switched-mode assisted linear power amplifier;

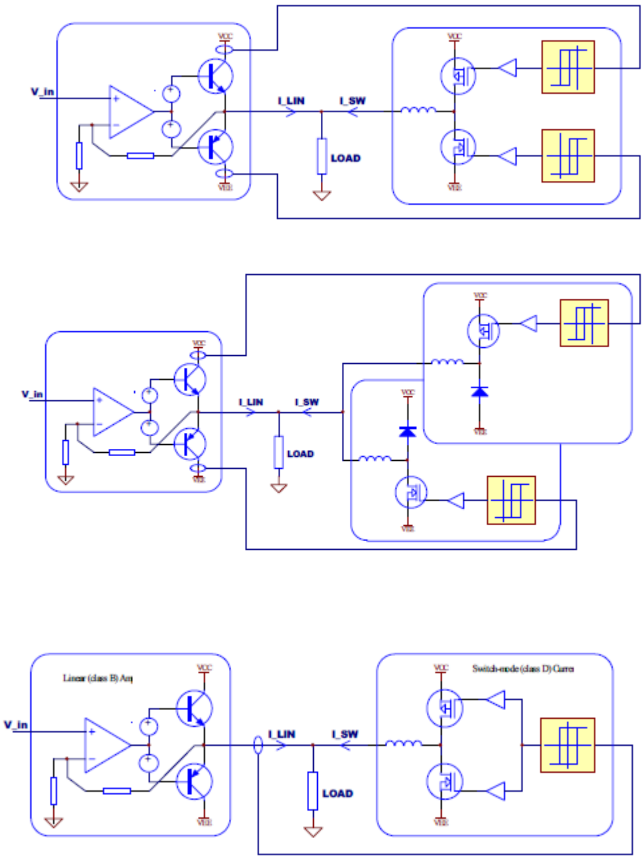

Fig. 4: switching technology module.

(a): switchmg stage output voltage; **(b):** output currents of the class D system and of the linear stage; (c) transistor currents for class B mode of the linear amplifier part;

(2) splitting-up the current dumping stage in several parallel branches operated in a phase shifted manner or application of a three-level topology

(3) higher order type coupling impedance of the switching stage . It has to be noted that the described effect only limits the power bandwidth of the current dumping stage and not of the whole amplifier system whose dynamic response (especially the slew-rate) **is** determined by the linear stage.

MOSFET safe operating area a) and the power losses b) of conventional linear power amplifiers and switched mode assisted linear **(SMALA)** amplifiers (both class B mode) for sinusoidal output voltage and different load current displacement factors.

III. SIMULATION RESULTS

A prototype system of a switched-mode assisted amplifier system with the nominal values V = 20V, R = **2.5R** (resistive **load** $\mathbf{Z_L}$ = R ; RMS value of the sinusoidal output voltage: 50V), f_B = l0 kHz,

f = 200kHz shall be calculated briefly. The power losses of the proposed system are far beneath the lsosses of conventional linear power amplifiers, especially for the case of non-resistive loads (e.g., the losses of a conventional linear amplifier would be P = 1kW for M=l and C_{osc}=0.5). However, it has to be admitted that the losses shown in Fig.4 for the switched mode assisted amplifier do not include the losses of the switching stage. On the other side, the efficiency of switched-mode bridge topologies

usually lies 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 shown in Fig.5. There, the pulse response demonstrates the limited slew-rate of the switched-mode current dumping system. In this case the output current of the linear amplifier i,,, not only has to compensate the ripple of the switching state but also has to take over the dynamic current
peaks This effect results in increased power losses of the linear stage.

IV. LINEAR STAGE DESIGN

A very low magnitude Z of the high-frequency output impedance z of the linear stage is of fundamental importance for a high input voltage signal-to-noise ratio (SNR) of the system because the ripple current AI of the switching stage generates a noise voltage. Today, the output stages of linear amplifiers usually are realized by using power MOSFET source followers [5]. The output impedance of source followers is defined by the transconductance (g) of, e.g., the upper transistor and is also influenced by the output impedance $\mathbf{R_i}$ of the driver stage in the upper frequency region.

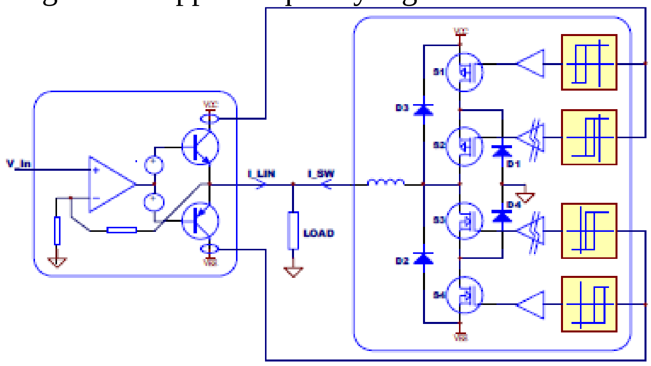

Fig5 : three level NPC switching level

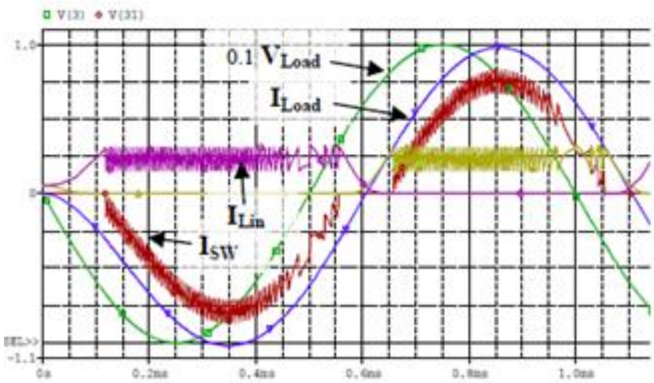

Fig. 6 : Output impedance of source followers.

In general, the transconductance of power MOSFETs is far too low to get an output impedance in the desired level. This fact is not of primary significance because the effective output impedance is reduced by the loop gain of the feedback system. For the described system we have to adjust the loop gain to 50dB at 200kHz. A higher loop gain would allow to further increase the SNR but would reduce the stability margin of the linear amplifier system. The frequency response of the amplifier mainly is determined by that of the voltage booster stage because the output current buffer usually shows a much higher bandwidth due to the application of MOSFETs and a high-frequency driver stage using bipolar video transistors. Contrary to conventional linear power amplifiers the frequency design of the voltage booster has to be performed not only regarding the power bandwidth but also has to consider the switching frequency of the current dumping stage in order to get the described reduction of the output impedance. Therefore, we use a symmetric wide-band push pull differential amplifier arrangement with a relatively low gain of 10 which, on the other side is high enough to use a conventional OP-AMP as feedback amplifier This OP-AMP is used as a PI-controller to increase the loop gain in the region of lower frequencies and to enhance the linearity of the system. A further improvement of the loop gain could be achieved by the well known principle of splitting-up the voltage booster into a low frequency part with full output voltage swing and a high-frequency small-signal path being **ar**ranged in parallel to increase the loop gain in the switching frequency region [6]. However, in any case the design of the feedback loop has to be adopted if the load impedance shows a capacitive portion due to the then given additional phase shift. In this case it would be more efficient to directly improve the output impedance of the current buffer stage using a feed forward compensation [7] or an inner feedback feed forward corrector scheme as proposed in SI.. the power MOSFETs have the advantage of a rectangular safe operating area which is of importance for the pulse response of the amplifier (Fig. ***5,*** right hand side).

V. ISOLATION OF SWITCH SIGNALS

The NPC topologies as shown would allow a SMALA to be built with equal power rating to a full bridge design using an equal number of equal The hysteresis control strategy using a pair of controllers retains the very simple sensing structure of the original. The drive of the
outer MOSFETS is also simple, requiring non-isolated
drivers. The inner NPC MOSFETs do require isolated gate drives. The low N channel device can be isolated using a level shifting boot strap driver HVIC. Most logically, this would be a half bridge driver with independent channels, such as the IR21xx family of devices [12], driving S3 and S4.
However, no equivalent device is available for the P channel devices, capable of level shifting to a more *negative* voltage for device S2. Recent digital isolators using IC scale magnetic coupling techniques from Analog Devices [13] and NVE corporation [14] have low propagation delay, low power consumption, and excellent noise immunity.

NPC SMALA ADVANTAGES

There are some key advantages of a three level NPC SMALA compared to a conventional two level SMALA which will assist with the overall implementation of the SMALA concept at higher power levels. The switching voltage waveforms applied to the inductor have half the voltage swing of a conventional two level SMALA. This allows the halving of the maximum switching frequency for a given inductor value and given inductor ripple current specification. The switching frequency and current ripple at near 0V (an important case for audio) is also much reduced due to the diode freewheeling current from 0V rather than the opposite voltage rail to the MOSFET delivering current. As mentioned, 100V rather than 200V MOSFETs will be able to be used at power levels of hundreds of Watts. This is particularly important given the limited availability of high voltage P channel MOSFETs.
Current experimental work is underway to build a two
and then three level SMALA using 80V supply rails and
capable of approximately 400W into 8Ω.

VI. CONCLUSIONS

It is observed that a simple extension of the hysteresis current control demonstrated in a previous two level SMALA implementation is suitable for controlling a three level Neutral Point Clamped (NPC) converter. The NPC topology allows the use of lower voltage switches and lower switching frequencies to implement high power audio amplifiers. It has to be noted that concerning the output impedance the realization of the output stage using bipolar power transistors would probably be a better solution because of their higher trans-conductance as compared to MOSFETs

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