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Guest Editorial
Special Section on Electric Vehicle Technology

I AM VERY GLAD to be able to present a Special Section on Electric Vehicle Technology in this issue of our TRANSACTIONS. On the eve of going to press with this special section, I was confronted by the following data. As recently as 1950, there were only 53 million motor vehicles registered in the world, and their exhaust emissions could still be tolerated because of their relatively modest effects. In 1992, our planet had well over half a billion cars and trucks! By the year 2000, their number will exceed one billion! If they were all to be powered by gasoline and diesel oil, our world could not stand it. Therefore, one of the most pressing demands of our time is an alternative clean, efficient, intelligent, and environmentally friendly urban transportation system. Electric vehicles offer a solution for improving air quality, reducing reliance on fossil fuels, and they are energy efficient. Furthermore, electric vehicles will be more intelligent to improve traffic safety and road utilization. In this special section, there are ten papers authored by researchers in academia and industry. These papers address the state of the art as well as some of the key issues and key technology of electric vehicles. The first paper, by Chan and Chau, provides an overview of current electric vehicle technology and the challenges ahead. The second paper, by Shimizu, Harada, Bland, Kawakami, and Chan, describes a unique ECO Vehicle Project in Japan with an in-wheel motor drive system, a hollow load floor which accommodates the batteries, and a new battery management system. The third paper, by Ehsani, Rahman, and Toliyat, addresses the system design philosophies of electric and hybrid vehicle propulsion systems. The dynamics are studied in an attempt to find an optimal torque-speed profile for the electric propulsion. The fourth paper, by Terashima, Ashikaga, Mizuno, Natori, Fujiwara, and Yada, describes unique in-wheel motors for a high-performance experimental electric vehicle. The fifth paper, by Profumo, Zhang, and Tenconi, describes alternative axial flux induction or synchronous in-wheel motors for electric vehicles. The sixth paper, by Rahman and Qin, presents the design, analysis, and PWM vector control of a hybrid permanent magnet hysteresis synchronous motor for electric vehicle application. The seventh paper, by Mutoh, Kaneko, Miyazaki, Masaki, and Obara, describes a torque controller which suits electric vehicle operating conditions. The eighth paper, by Kawamura, Hoshi, Kim, Yokoyama, and Kume, proposes an anti-directional-twin-rotary motor drive as a new power train for electric vehicles. The ninth paper, by Lai, presents resonant snubber-based soft-switching inverters for electric propulsion drives, which have superior performance in efficiency improvement, EMI reduction, and dv/dt reduction. The last paper, by Kim and Ha, deals with the design of interface circuits with electrical battery models. On the whole, the above ten papers address important technology for the next century. They deal with the key components in electric vehicle development, namely system design philosophy, various options of electric motor drives and energy management. These are challenges for our profession. The 21st century will be the environmental century, and electric vehicles will be the major means of urban transportation.

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Basic Considerations and Topologies of Switched-Mode Assisted Linear Power Amplifiers

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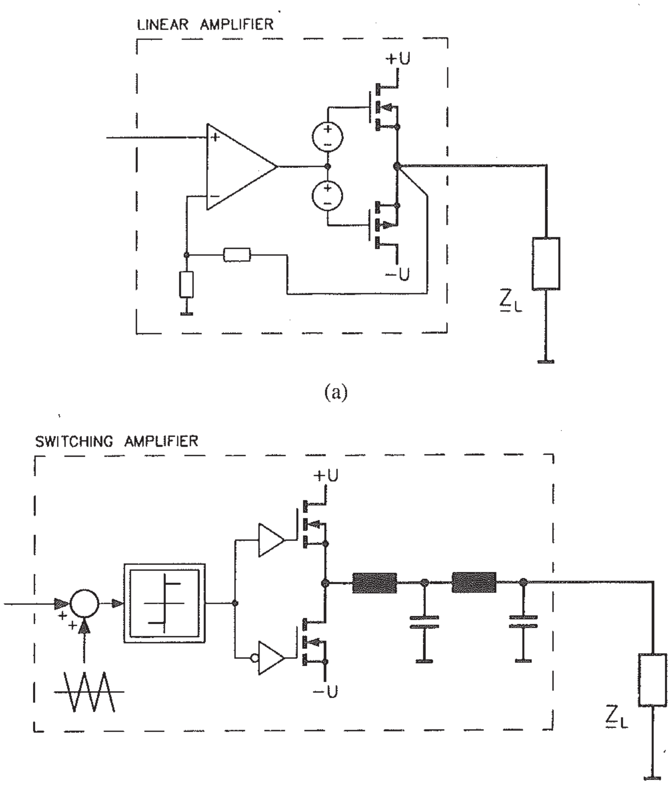
***Abstract*— This paper presents a combined power amplifier system consisting of a linear amplifier unit with a switched-mode (class D) current dumping stage arranged in parallel. With this topology, the fundamental drawback of conventional linear power amplifiers—the high loss—is avoided. Compared to a pure class D (switching) amplifier, the presented system needs no output filter to reduce the switching frequency harmonics. This filter (usually of multistage type) generally deteriorates the transient response of the system and impairs the feedback loop design. Furthermore, the low-frequency distortions of switching amplifiers caused by the interlock delay of their power transistors are avoided with the presented switched-mode assisted linear amplifier system. This can be considered as a master–slave system with a guiding linear amplifier and a supporting class D slave unit. The paper describes the operating principle of the system, analyzes the fundamental relationships for the circuit design, and presents simulation results. Finally, various further topologies of switched-mode assisted linear amplifiers are given.**

***Index Terms*— Class D converters, dc–ac power conversion, power amplifiers, switching amplifiers.**

I. INTRODUCTION

CONVENTIONAL linear power amplifiers (Fig. 1(a)) are replaced by switching (class D) amplifiers (Fig. 1(b)) in an increased quantity to overcome the essential drawback of linear amplifier systems, i.e., the high losses (especially in the case of nonresistive or nonlinear loads or if signals with high peak-to-rms ratio are amplified [1]). Nevertheless, if the output voltages have to be of very high quality (e.g., for high-end audio applications or for test and measuring equipment), switching amplifiers show significant limitations. The output voltage of a class D amplifier implies substantial switching-frequency components (high frequency distortions) which have to be reduced by a proper low-pass filter. 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, and/or a ripple of the dc supply voltage $\pm U$ and/or the on-state voltages of the power semiconductor devices (transistors and freewheeling diodes), 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 subharmonic frequency components which may result for a small signal-to-switching-frequency ratio, or if a pulse width modulation strategy with not constant switching frequency (e.g., hysteresis control or sigma–delta modulation) is applied. This subharmonic noise basically cannot be lowered by the output low-pass filter because the relevant frequency components lie within the power bandwidth of the amplifier.



(a)

(b)

Fig. 1. Simplified circuit diagram of (a) a linear power amplifier and (b) a class D switching amplifier (the internal body diodes of the switching power MOSFET's are not shown).

To avoid the disadvantages described above, a concept originally proposed in [5] consisting of a parallel arrangement of a class D switching system and a conventional linear amplifier stage (Fig. 2) is analyzed. There, the output filter of the switching amplifier is reduced to a single coupling inductor determining the switching frequency ripple. Although the linear amplifier, therefore, 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

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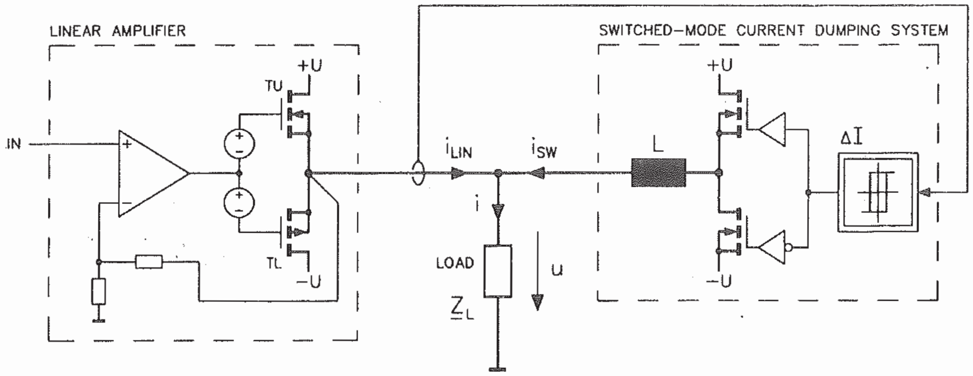

Fig. 2. Circuit diagram of a switched-mode assisted linear power amplifier.

(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 subharmonic components. It has to be pointed out, however, that a very low output impedance of the linear system part is of paramount importance in order to get a high noise rejection. This circumstance has to be considered by an appropriate design of the linear amplifier circuitry and feedback system. Furthermore, the switched-mode assisted linear amplifier only allows a significant reduction but not a complete loss elimination as an idealized class D amplifier. Therefore, considering the losses, the proposed system can be seen as an intermediate solution between pure linear and pure class D power amplifiers. As an advantage of the proposed system, it has to be mentioned that the dynamic response of the whole system is determined by the linear stage and, therefore, not influenced by an output filter.

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 have to be delivered by the linear stage. Instead of an explicit subtraction of reference value (load current i) and actual value (class D stage output current i_{SW}), the calculation of the controlling quantity can be done in an implicit manner by direct measurement of the linear stage output current i_{LIN}. In the simplest case, the current controller can be a hysteresis controller (Fig. 2), which results in a nonconstant switching frequency within the fundamental period of the amplified signal. As an alternative, a pulse width modulator (PWM) with a superimposed linear current controller, or other types of current controllers being well-known from switched-mode power supplies (e.g., conductance control), can be applied. The usage of a PWM 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 its inherent overmodulation ability which yields a more efficient utilization of the dc supply 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 (Section V).

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 ΔI is applied. It is assumed that the load current i and the output voltage u can be treated as constant within the switching interval T, or that there exists a sufficient signal-to-switching frequency ratio, respectively (Fig. 3). Furthermore, the power transistors are assumed to be ideal (neglection of delay times, on-state voltages, etc.). Also, dc supply voltage variations are neglected.

Switching Frequency: With the assumptions given above, the output voltage u (averaged within a pulse interval T) is determined by the duty cycle δ. If we apply the definition $m = u/U$ for normalizing the output voltage ($m = -1 \cdots +1$), we get

$$\delta = \frac{1 + u/U}{2} = \frac{1+m}{2}. \tag{1}$$

According to $u_L = L\ di_{\mathrm{SW}}/dt$, the switching frequency $f_s = 1/T$ can be calculated

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

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 following table gives the local losses (i.e., the losses averaged within a switching period T) of the upper transistor TU and the lower transistor TL of the linear stage, where it is assumed that for class A mode the quiescent current is as small as possible ($I_{\mathrm{Q}} = I_{\mathrm{Q,min}} = \Delta I/4$) (see Fig. 3(e)).

For $I_{\mathrm{Q}} = \Delta I/4$, the class A mode losses are twice the losses of the class B mode. The total transistor losses p_T are not dependent on the modulation index m and, therefore, the local transistor losses p_T also represent the global losses

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