

Constant Switching Frequency Self-Oscillating Controlled Class-D Amplifiers

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Abstract—The self-oscillating control approach has been used extensively in class-D amplifiers. It has several advantages such as high bandwidth and high audio performance. However, one of the primary disadvantages in a self-oscillating controlled system is that the switching frequency of the amplifier varies with the ratio of the output voltage to the input rail voltage. In other words, the switching frequency varies with the duty cycle of the output. The drop in the frequency results in lower control bandwidth and higher output voltage ripple, which are undesirable. This paper proposes a new self-oscillating control scheme that maintains a constant switching frequency over the full range of output voltage. The frequency difference is processed by a compensator whose output adjusts the total loop gain of the control system. It has been proven by simulation that a constant switching frequency self-oscillating converter is achieved and the proposed control circuit performs satisfactorily.

Index Terms—Power amplifiers, frequency control, voltage control, power electronics.

I. INTRODUCTION

Switch mode class-D audio amplifiers have been replacing the audio amplifiers made from class-A, B or AB amplifiers thanks to their superior efficiency and power density [1]–[5]. For the control of class-D audio amplifiers, the modulation strategy is an integral factor that determines the performance of the audio amplifiers. Modulation strategies can be classified into analogue modulation and digital modulation.

The analogue modulation usually has better performance due to the absence of digital delay or quantization error, which are inherent in digital modulation [6]. There have been two basic analogue modulation methods in the literature, namely the triangular carrier-based modulation method and the self-oscillating control method. The self-oscillating control method uses a comparator with either hysteresis [7], [8] or zero-hysteresis [9].

Recently, the self-oscillating control method has received great attention due to its advantages over the triangular carrier based method [6]–[13]. The advantage of the self-oscillating control system is that it allows the open-loop control system magnitude to cross the 0 dB point precisely

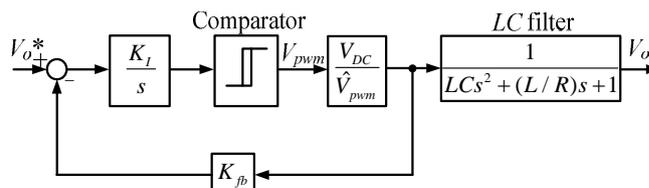


Fig. 1. A basic voltage-mode hysteresis controller.

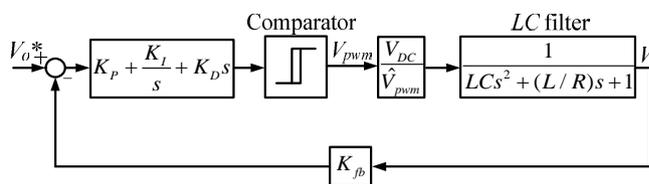


Fig. 2. A global loop integrating modulator controller.

at the oscillation frequency, providing high control bandwidth. On the other hand, the cross over frequency of the carrier-based modulation control system must be at least 2 times lower than the switching frequency, resulting in lower bandwidth and higher output ripple.

Figure 1 introduces a basic voltage-mode, hysteresis-based, self-oscillating control system. The output filter is a second-order LC filter. The switching node of a synchronous buck converter is fed back and compared with the voltage reference, which, in the case of audio amplifiers, is a reference for the audio output signal. K_{fb} is the feedback constant of the switching signal. The rail voltage is V_{DC} with respect to ground for a single supply converter, or can be from $+V_{DC}/2$ to $-V_{DC}/2$ for dual supply audio amplifiers. For consistency, a single supply voltage is assumed. In the control scheme, the integrator amplifies the error and provides infinite gain at dc. The phase shift created by the integrator is -90° . The negative feedback of the signal provides a phase shift of -180° . The remaining phase shift is provided by the nature of the hysteresis comparator, the time delay of gate driver, and the propagation of the control units. Assuming there is no other time delay produced by the other physical elements in the control loop, the hysteresis comparator provides -90° phase shift at the nominal switching frequency.

Figure 2 shows another hysteresis-based, self-oscillating control system for class-D amplifiers. It feeds back the output voltage instead of the switching signal. This scheme is usually referred to as the Global Loop Integrating Modulator (GLIM) scheme [12]. Since the transfer function

of the out-put filter is taken into the control loop, the integrator in Fig. 1 is replaced by a proportional-integral-derivative (PID) controller. The PID controller cancels out the two poles created by the output filter and provides high dc gain with its integral term. The PID controller, together with the output filter, can be approximately treated as an integral, making the open loop transfer function similar to that in Fig. 1. The advantage of this scheme compared to the basic scheme in Fig. 1 is that the errors from both the output voltage and output filter are attenuated [12].

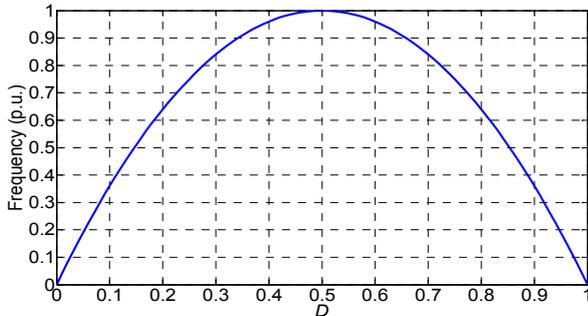


Fig. 3. Frequency (nominal to the maximum frequency) versus changes of duty cycle.

However, a major problem with the self-oscillating control system applied in class-D amplifiers is that the switching frequency varies with the duty cycle of the output stage. The variation is based on a parabolic curve with peak value at a duty cycle of 0.5. This is due to the decrease in the gain when the duty cycle is different from 0.5. From previously published papers, it can be shown that the switching frequency has the following expression [6]–[8]

$$f_{sw} = \frac{K_{sw}D(1-D)}{v}, \quad (1)$$

where v is the height of the hysteresis threshold (hysteresis window), K_{sw} is a constant dependant on the circuit gain, and D is the duty cycle. In the case of a single supply system, the modulation index is defined by the proximity of the output signal V_o , which is biased by half the rail supply voltage, to the closest supply rail. Thus, the modulation index is defined as

$$M = \frac{|V_o - V_{DC}/2|}{V_{DC}/2} = |2D - 1|. \quad (2)$$

The *centre duty cycle* is the duty cycle that produces the maximum frequency in a self-oscillating controlled system. In this case, the centre duty cycle is $D = 0.5$. A plot of the switching frequency, nominalized to the maximum frequency, versus the change in duty cycle is shown in Fig. 3. It can be seen that the frequency drops rapidly with the change of duty cycle away from the centre duty cycle. For example, when the duty cycle is 0.1 or 0.9, the switching frequency drops 64 %, or nearly two thirds from the nominal frequency.

This drop in the switching frequency is undesirable because it results in several undesired consequences. First, the drop of the switching frequency increases the output

voltage ripple. Second, the drop causes the open-loop bandwidth and loop gain to drop as well, resulting in higher distortion and slower dynamics. As a result, it is desired that the switching frequency in self-oscillating controlled systems be fixed at the nominal frequency in order to counteract the aforementioned drawbacks.

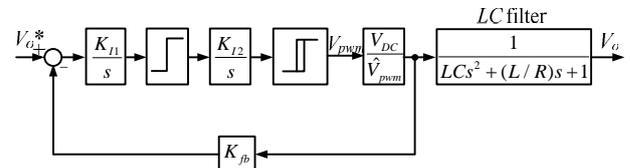


Fig. 4. The fixed frequency self-oscillating control system proposed in [6].

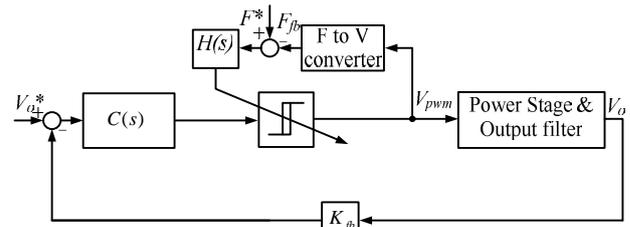


Fig. 5. The fixed frequency self-oscillating control system proposed in [8] and [13].

Based on the independence of the switching frequency to the threshold and the gain, as shown in equation (1), the switching frequency can be fixed by either altering the gain [6] or online-adjusting the hysteresis threshold [2], [7], [8], [13]. Between these approaches, the later has dominated.

In [6], the frequency is fixed by modifying the control loop. Specifically, a second integrator was inserted into the circuit together with a second comparator. The control scheme is redrawn in Fig. 4. The drawback is that the switching frequency is nearly but not exactly constant.

The methods proposed in [8] are very similar to those in [13], wherein the frequency is converted to voltage, which is then processed by a high gain compensator. The output of the high gain compensator is used directly to adjust the hysteresis window through, for example, a flip-flop circuit. The only difference between [8] and [13] is that the method in [8] is applied to a different type of converter, namely the buck converter with a fourth-order output filter. However, ideas proposed in [13] also includes phase compensation by use of a phase-lock-loop circuit, unlike [8]. Their control block diagram is redrawn in Fig. 5, where $C(s)$ is a generalized compensator for the output voltage, and $H(s)$ is the aforementioned high gain compensator.

Reference [2] introduces the concept of fixing the frequency by adjusting the hysteresis threshold from the output of a compensator that processes feed-forward input reference signals or feedback PWM signals. However, the discussion of this concept in [2] is general, and there is neither proper analysis nor associate evidence to support its claims.

This paper seeks to solve the problem in a different way. Similar to [6], the hysteresis threshold v will be fixed. However, the gain of the open loop control system is adjusted by a feedback control of the frequency signal, so that the whole open loop gain crosses the 0 dB point at the desired switching frequency. The compensating gain will be

injected to the open loop system via a multiplication unit (multiplier). The detailed description of the proposed method will be presented next.

II. PROPOSED CONSTANT SWITCHING FREQUENCY CONTROLLER

In the studied system, a rail voltage of 60 V dc is used to supply the power stage. The converter is a synchronous buck converter. The nominal load is 26 . The nominal switching frequency is 1 MHz. The specifications of the converter to be examined are listed in Table I.

The proposed fixed-frequency, self-oscillating control system is shown in Fig. 6. The PWM signal generated by the hysteresis comparator is converted to a voltage proportional to its frequency. This is done by passing the PWM to a mono-stable multi-vibrator (MMV) or one-shot circuit with a fixed pulse width output. Each time the input to the MMV has a rising edge, the output of the MMV generates a pulse with a fixed width, which is 100 ns in this case. The resulting signal of the MMV is low-pass filtered by a first order RC filter circuit with a cut-off frequency of 1.6 kHz. Only the dc value of the MMV remains, and it is proportional to the switching frequency. The operation of the MMV and the F-to-V converter are illustrated in Fig. 7 and Fig. 8.

The measured switching frequency is compared to the reference switching frequency and processed by a compensator $F(s)$. $F(s)$ can be implemented with a PI or PID controller. The output of $F(s)$ is the compensating gain for the self-oscillating control loop, and it is inserted to the control loop through a multiplier. It adjusts the switching frequency of the converter to track the reference frequency, which is represented by F^* .

III. SIMULATION RESULTS

The simulation studies examine different responses of the converter using the self-oscillating control approach without frequency compensation as well as with the proposed frequency compensation. The responses are based on both dc reference signals and sinusoidal audio reference signals. Simulation model was built in MATLAB®/Simulink environment. Most parts of the controllers and feedback were modelled and simulated by realistic commercially available discrete components: non-ideal operational amplifiers with limited gain-bandwidth product and limited output voltage ability, etc.

A. Converter without a Frequency Compensator

Figure 9 shows the voltage transient (step) response of the converter without a frequency compensator. The dc reference values are swept so that the duty cycle of the output voltage is changed. At time $t = 0$, the duty cycle is 0.5. At times $t = 70 \mu\text{s}$, $t = 140 \mu\text{s}$, $t = 210 \mu\text{s}$, the duty cycle is 0.2, 0.1, and 0.7, respectively. The bottom of Fig. 9 shows the switching signal generated by the comparator. It can be seen that the switching frequency varies with the change of the duty cycle. Its peak value is at the centre frequency where $D = 0.5$. This phenomenon can be observed more clearly from a partial zoom of Fig. 9 which is shown in

Fig. 10.

TABLE I. THE SPECIFICATION OF THE CONVERTER.

Parameter	Value
L	36 μH
C	100 nH
V_{DC}	60 V
\hat{V}_{pwm}	5 V
V	0.66 V
Nominal load	26
Nominal switching frequency	1 MHz
tMMV	100 ns

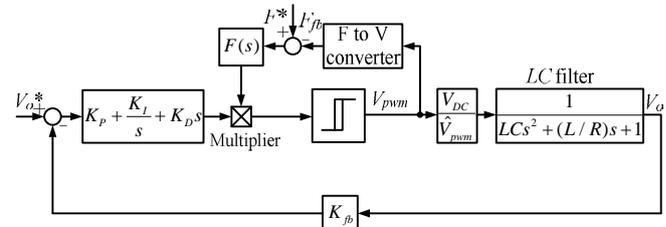


Fig. 6. The proposed fixed frequency self-oscillating control system.

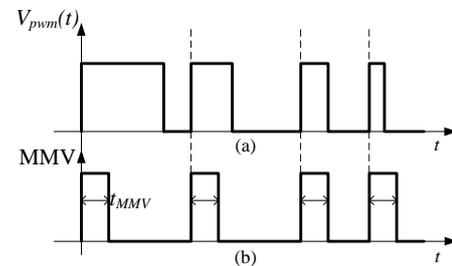


Fig. 7. Operation of the MMV a) the PWM output signal of the comparator, b) the output signal of the mono-stable multi-vibrator.

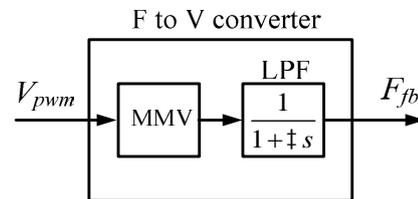


Fig. 8. Frequency to voltage (F to V) converter.

Figure 11 shows response to an audio signal of the conventional converter without a frequency compensator. The frequency of the signal is 20 kHz, which represents an audio signal for typical human hearing ability. It can be confirmed again from the simulation result, that the switching frequency varies with the magnitude of the audio signal, or in other words, it varies with the switching duty cycle. As Fig. 3 suggests, the drop of the frequency at higher modulation index creates larger output ripple and distortion. This phenomenon can be observed from the top and bottom of the output voltage signal.

B. Converter with the Proposed Frequency Compensator

Figure 12 shows the step response of the self-oscillating control amplifier with the proposed frequency compensator. As can be seen, the switching frequency of the converter in steady state held constant regardless of the variation in the output voltage or duty cycle. The steady-state response of the output voltage has been improved, while a desirable transient response is preserved. The ripple of the output voltage is reduced at all switching duty cycles that are

different from the centre duty cycle. This feature can be seen in the partial zoom of Fig. 12 shown in Fig. 13.

Converter without a frequency compensator is shown in Fig. 9–Fig. 11.

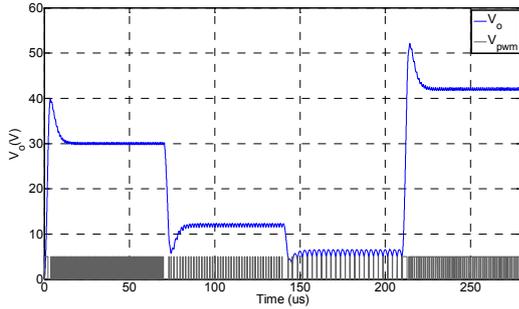


Fig. 9. Step response of output voltage without a frequency compensator. Top: V_o (V), bottom: V_{pwm} (V), the duty cycle is swept from 0.5 to 0.2, 0.1, and 0.7.

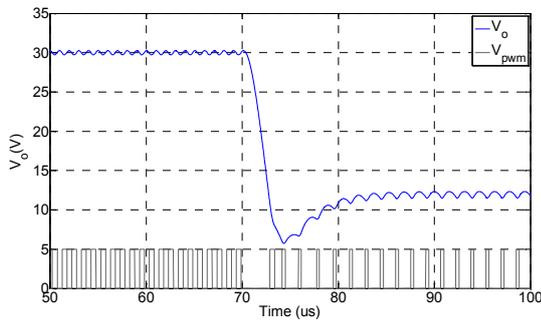


Fig. 10. Partial zoom of the step response of output voltage without a frequency compensator. Top: V_o (V), bottom: V_{pwm} (V), the duty cycle is shifting from 0.5 to 0.2 at time 70 μ s.

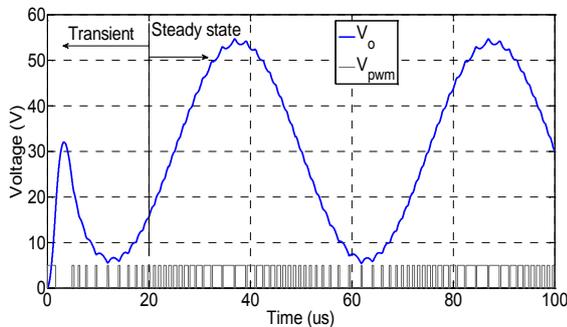


Fig. 11. Output response to a 20-kHz audio reference without a frequency compensator. Top: V_o (V), bottom: V_{pwm} (V), the duty cycle is continuously varying from 0.1 to 0.9 and vice versa.

The response to an audio input signal of 20 kHz is presented in Fig. 14. It can be seen that the switching frequency of the converter has been held constant even with a large variation in the duty cycle over one period. The ripple at the top peak and bottom peak of the output signal, therefore, has been significantly improved compared to Fig. 11. The amplifier with the proposed constant-frequency compensator outperforms the traditional one without a frequency compensator. Different simulations are performed with both converters at various duty cycles. The switching frequencies are plotted in Fig. 15 for four cases: calculated by equation (1), simulated model without a frequency compensator, simulated model with the proposed frequency compensator, and simulated model with method proposed in [6]. As can be seen, the simulated switching frequency

without a frequency compensator matches well with the calculation. Moreover, a constant switching frequency is guaranteed with the proposed method.

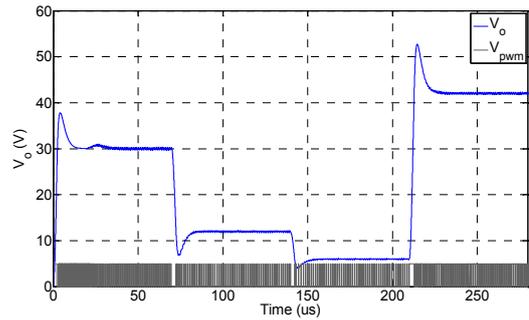


Fig. 12. Step response of output voltage with the proposed frequency compensator. Top: V_o (V), bottom: V_{pwm} (V), the duty cycle is swept from 0.5 to 0.2, 0.1, and 0.7.

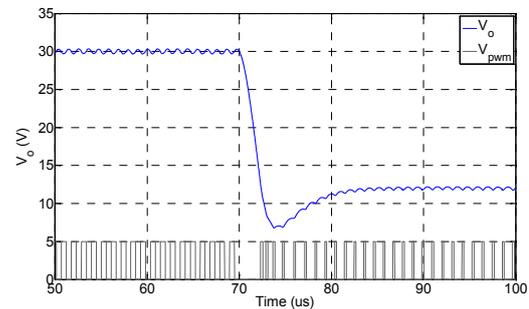


Fig. 13. Partial zoom of the step response of output voltage with the proposed frequency compensator. Top: V_o (V), bottom: V_{pwm} (V), the duty cycle is shifting from 0.5 to 0.2 at time 70 μ s.

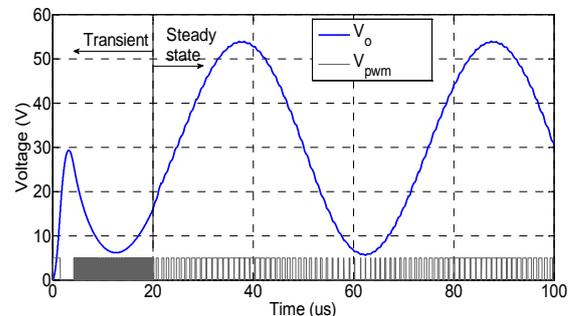


Fig. 14. Output response to a 20-kHz audio reference with the proposed frequency compensator. Top: V_o (V), bottom: V_{pwm} (V), the duty cycle is continuously varying from 0.1 to 0.9 and vice versa.

Converter with the proposed frequency compensator is shown in Fig. 12–Fig. 14.

IV. DISCUSSION

While most existing solutions for fixing the switching frequency have worked with varying the hysteresis window when the duty cycle changes, the approach presented here fixes the hysteresis window and generates a compensating gain based on the variation in frequency.

In [6], the hysteresis window is kept constant and the regulation of frequency is achieved by changing the loop gain. The added integrator is implemented by a passive RC filter. Therefore, the low frequency gain is preserved, but the high frequency components are attenuated. The extra comparator differentiates the output of the added integrator, yielding clamped output signals. Therefore, the gain of the

low frequency signal is preserved at the rate of -20 dB/dec, ensuring a linear carrier signal to the main comparator, while the high frequency signals are approximately attenuated with a slope of -40 dB/dec. As a result, a drop of gain due to a change of modulation index will result in half of the drop in the switching frequency compared to the uncompensated system. In fact, that method is only able to reduce the drop of the switching frequency by approximately half compared to an uncompensated system. Finally, the article also states that only “close to constant switching frequency” is achieved. The advantage of this method is its simplicity and low-cost compared to the method proposed in this paper.

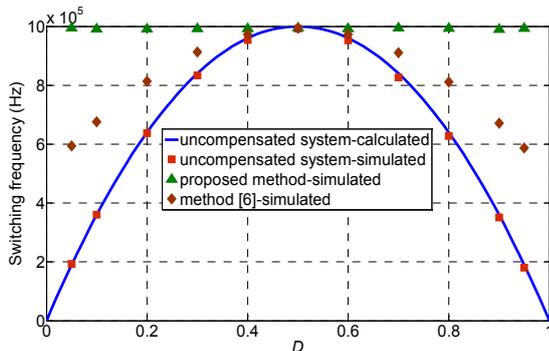


Fig. 15. Switching frequency versus variation of duty cycle: calculated results without frequency compensation, simulated results without frequency compensation, simulated results with proposed frequency compensation, and simulated results with method proposed in [6].

In [7], the comparator is configured with positive feedback to yield hysteresis. The novelty of this method is to use feedback impedance, which consists of frequency dependant components such as an RC network instead of passive resistors. The network is used to filter out the PWM signals generated by the comparators. When the duty cycle diverges from the centre value, the filtered signal has a smaller value, providing a higher gain to compensate the drop of term $D(1-D)$ in equation (1). However, the compensated gain is only dependant on the impedance magnitude of the RC network, which proves to vary differently from the desired parabolic gain of $D(1-D)$. The method is only valid at a specific value of the output duty cycle, and cannot guarantee a constant switching frequency over the whole output range. In fact, the frequency is only raised for different modulation indexes, but it fails to be raised at very high modulation index. As a result, only “close to constant switching frequency” is claimed. The advantage is simple and low-cost implementation by means of insertion of passive resistors and capacitors.

The methods in [8] and [13], due to their direct compensation of either frequency errors or phase errors, have the advantage of precise control of the desired switching frequency. The penalties of those methods are high complexity and component count if implemented by analogue components.

Compared to the known existing methods analysed above, the proposed method has high precision frequency control for similar reasons as in [8] and [13]. However, the significant differences are the compensation mechanism and

the method of implementation. The proposed method focuses on compensating the drop of open loop gain by providing a correcting gain and injecting it to the self-oscillating loop. Conversely, the methods in [8] and [13] adopt the classical approach of changing the hysteresis threshold. Implementation of adjusting the hysteresis threshold must involve extra effort to design the comparator circuit with fast online tuning capability of the hysteresis threshold. On the other hand, implementation of the proposed compensator can be less complex, involving a good dynamic multiplication unit while allowing the use of standard comparators.

V. CONCLUSIONS

In this paper, the authors have proposed a new approach to maintain constant switching frequency of self-oscillating controlled converters. It has been proven by extensive simulation studies that the proposed idea is effective. The proposed methods not only preserve the fast transient response characteristic of a self-oscillating controlled system, but also improve the steady state response by lowering the output voltage ripple. The feasibility of the idea can be verified with experiments in a future work.

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