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Constant Switching Frequency Self-Oscillating Controlled Class-D Amplifiers

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\(^1\)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 ripple voltage, 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 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. \(K_p\) 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

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of the output 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
time voltage and output filter are attenuated [12].

\[ f_{sw} = \frac{K_{sw} \cdot D(1 - D)}{\varepsilon} \]

where \( \varepsilon \) is the height of the hysteresis threshold (hysteresis
window), \( K_{sw} \) is a constant dependent 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

\[ M = \frac{|V_o - V_{DC}/2|}{V_{DC}/2} = |D - 1|. \]

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, normalized 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.

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],
[10]. 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
[10], 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 [10] 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 [10] 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 \( \varepsilon \) 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
The specifications of the converter to be examined are listed in Table I.

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

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 $t = 70 \mu s$, $t = 140 \mu s$, $t = 210 \mu s$, 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     |
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| Nominal load    | 26        |
| Nominal switching frequency | 1 MHz |
| $\tau_{MMV}$   | 100 ns    |

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.](image)

![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 µs.](image)

![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.](image)

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.](image)

![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 µs.](image)

![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.](image)

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. 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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