Modulated Ramp PWM Generator for Linear Control of the Boost Converter’s Power Stage

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Abstract — In this paper, the concept of the modulated-ramp PWM generator is presented. The specific nonlinear behavior of this generator enables an exact linear control-to-output static conversion characteristic of DC-DC switching converters, such as the boost (step-up), buck/boost, Ćuk or Sepic, operating in Continuous Conduction Mode (CCM). The resulting linearization leads to a significant improvement of the control accuracy, and simplification of the converters, operating mainly in the open-loop architecture. In this paper, the concept of the modulated-ramp PWM generator is described, followed by a detailed steady-state and small-signal AC analysis. The key aspects for practical implementation are also highlighted. All results and benefits of the modulated-ramp PWM generator are demonstrated here on the boost converter power stage.

I. INTRODUCTION

The conventional realization of switched-mode power converters such as boost, buck-boost, Ćuk or SEPIC suffers from significant nonlinearity of the control-to-output transfer function [1], [2]. The nonlinearity is manifested mainly by a significant variation of the control gain GC. As a consequence, specific nonlinear control schemes are frequently used for boost open loop or closed loop converter architectures.

The basic scheme of a boost converter power stage is shown in Fig. 1. It consists of a voltage source VIN, a switch SW including parasitic resistances RSW, an LC filter with parasitic elements RL and RSESR, a load resistance R, and a conventional (linear) pulse-width modulator (PWM). In Continuous Conduction Mode (CCM), the static input-to-output conversion ratio VOUT/VIN for the ideal converter (RL = RSW = 0) is usually determined from equal positive and negative inductor current increments ΔI1,ON = ΔI1,OFF occurring during the ON and OFF switching phases [1], [2], [3] as:

\[ \frac{V_{OUT}}{V_{IN}} = \frac{1}{1 - D} \]

(1)

where D is the duty-cycle ratio. The small-signal control gain GC relating to the incremental gain at the operating point (steady-state output voltage) can be obtained from (1) as the partial derivative:

\[ G_C = \frac{\partial V_{OUT}}{\partial D} \Bigg|_{V_{IN} = \text{const}} = \frac{V_{IN}}{(1 - D)^2} \]

(2)

This expression is highly non-linear. As we can observe from the graphical interpretation of (2) shown in Fig. 2, the usual duty-cycle range (D < 0.9) causes high variation of the small-signal control gain GC, even up to 40dB. This means that an incremental ΔD at high duty cycle ratio will cause significantly higher (up to 100-times) variation of VOUT, compared to the identical increment ΔD applied at low duty-cycle ratio. In the open loop architecture, this nonlinearity decreases...
the conversion ratio accuracy (1) due to increased sensitivity, whereas in the closed loop architecture, it can affect the stability of the converter. In addition, another nonlinearity caused by the non-ideal components is encountered in the analysis of the converter. As we will show later, this involves mainly the parasitic resistances of the inductor and switch, and is manifested by the negative control gain $G_c$ at high duty-cycle (see Fig. 9) [1].

In the domain of the feedback control, a lot has been done to address the problem of nonlinear control-to-output conversion. The closed-loop transfer function design in voltage control mode is usually based on the linearization around an operating point [3], [4], [5], or by using feedback linearization [6], [7].

A more suitable way is the predistortion technique which allows linearization of the static conversion ratio control through the specific PWM modulator schemes. Basically, predistortion can either be applied to the control signal [8] or to the voltage ramp [9]. However, common problems of these techniques are the increased modulator complexity, and low accuracy of generated duty-cycle. We can refer for example to [8], where the error signal used for the predistortion is obtained by the filtering of the output PWM signal waveform. An analysis of this technique reveals a high sensitivity to the power-supply voltage (PWM signal peak amplitude), and to the delay caused by the filtering of the PWM signal, where several clock periods are required to obtain properly filtered error signal.

In this paper, a concept based on the predistortion by modulated-ramp PWM generator is presented. The result shows that a linear relationship between control variable and output voltage is obtained. The achieved linear control characteristic yields considerably lower sensitivities, compared to the conventional linear PWM modulator, namely for high duty-cycle values ($D > 0.5$). This feature can allow to avoid the use of the closed-loop architecture, resulting in a simpler open-loop architecture of the power converter. This simplification can be done in cases where the constraints of the output voltage accuracy are low, such as the power supply of RF PA, LED drivers, battery chargers etc. Concept of modulated-ramp PWM generator can also be used in the closed loop architecture (current or voltage mode), where, as shown in the following, the nonlinear dynamic behavior of the power stage has to be considered.

This article is organized as follows: in section II, the concept of a current-modulated ramp PWM generator is presented, with an emphasis on the fundamental relationships between the input control value (control current $I_{CON}$), duty-cycle $D$, and boost converter output voltage $V_{OUT}$. Section III presents a detailed analysis of the modulated-ramp boost converter, while section IV deals with basic implementation constraints and shows an experimental result.

I. CURRENT-MODULATED RAMP PWM GENERATOR

The conventional “linear” PWM generator is depicted in Fig. 3. As we can see, it consists of a constant voltage ramp generator, comparator, and control input $V_{CON}(t)$. The constant voltage ramp is obtained thanks to a DC bias current $I_b$ injected into the periodically discharged capacitor $C$. This type of PWM generator yields a constant control-voltage to duty-cycle gain $1/V_{in}$ ($V_{in}$ is the voltage ramp amplitude and is further considered unitary). Hence, use of this generator preserves the nonlinear conversion characteristic $V_{OUT}/V_{IN}$ and DC control gain $G_C$, given by Eqs. (1) and (2).

![Fig. 3. Model of conventional (linear) PWM generator composed of a constant current source $I_b$, periodically discharged capacitor, and comparator.](image)

A. Current-modulated Ramp PWM Generator

The permutation of the control and reference inputs ($I_b$ and $V_{CON}$) in the original circuit from Fig. 3 results in a suitable nonlinear behavior of the generator. This modification creates the generator PWM with modulated voltage ramp, which is shown in Fig. 4. Here, the input variable is the control current $I_{CON}(t)$, whereas the comparator input voltage $V_b$ remains constant.

![Fig. 4. Current-modulated ramp PWM generator. $I_{CON}(t)$ is the input control variable and $V_b$ is the constant bias voltage.](image)

The control-to-duty-cycle conversion ratio of the modulator shown in Fig. 4 can be determined from a time analysis of the capacitor voltage $V_C(t)$. When considering, for convenience, the control current $I_{CON}$ constant during one clock period $T$, the capacitor voltage $V_C(t)$ results in a linear time function of $I_{CON}$:

$$V_C(t) = \frac{1}{C} \int I_{CON}(t) dt \approx \frac{I_{CON}}{C} t$$  \quad (3)

As shown in Fig. 4, capacitor voltage is compared with an arbitrary reference voltage $V_b$. While $V_C < V_b$, the output of the PWM modulator is low. $V_C(t)$ reaches the value $V_b$ at $t = V_b C / I_{CON}$, at which time the output of the modulator goes high. Considering the capacitor being discharged with period $T$, the duty-cycle ratio $D$ can be expressed as:

$$D = 1 - \frac{V_b C}{I_{CON} T} = 1 - \frac{\alpha}{I_{CON}}$$  \quad (4)

which is defined for $I_{CON,min} > V_b C/T$. A demonstration for two...
discrete values of control current $I_{\text{CON1}}$ and $I_{\text{CON2}}$ is shown in Fig. 5, where it can be seen how the slope of $V_c(t)$ varies with the control current.

Fig. 5. Capacitor voltage $V_c(t)$ for two different values of $I_{\text{CON1}}$ and $I_{\text{CON2}}$ ($I_{\text{CON1}}$ and $I_{\text{CON2}}$ are considered constant within the clock period).

B. Demonstration on the Step-Up Converter

The obtained relationship (4) between duty-cycle $D$ and control current $I_{\text{CON}}$ can be substituted in the boost converter conversion ratio Eq. (1). The resulting dependency is a linear function of $I_{\text{CON}}$:

$$\frac{V_{\text{OUT}}}{V_{\text{IN}}} = \frac{I_{\text{CON}} T}{V_b C}$$

and is defined for $I_{\text{CON}} > I_{\text{CON(min)}}$. Small-signal control gain $G_C$ can be obtained from (5) as a constant term:

$$G_C = \frac{\partial V_{\text{OUT}}}{\partial I_{\text{CON}}} \Bigg|_{I_{\text{CON}}=\text{const}} = \frac{V_b T}{V_b C}$$

This means that, contrary to (2), an arbitrarily small step $\Delta I_{\text{CON}}$ causes an identical step $\Delta V_{\text{OUT}}$, independent on the steady-state output voltage. Similar to the boost-converter power stage, conversion characteristics and small-signal control gain for other mentioned converters can be obtained. As we can see in Table 1 below, the boost, buck-boost, Cuk, and SEPIC DC/DC converters with modulated-ramp PWM generator exhibit linear $V_{\text{OUT}}/V_{\text{IN}}$ conversion characteristic and a constant small-signal gain $G_C$.

C. Switched-simulation Example

The technique presented above for the linearization of the static conversion characteristic is demonstrated here by the slow (i.e. quasi-static) switched transient simulation of an ideal boost converter power stage. The simulation setup is shown in Fig. 6. In this model, the excitations (control values $V_{\text{CON}}$ and $I_{\text{CON}}$) are realised: i) for the linear PWM generator by a 40ms voltage ramp from 0 to 1V, and ii) for the modulated ramp PWM by a 40ms current ramp from 1.6µA to 16µA. The resulting time responses are shown in Fig. 7, where the linear control to output conversion for modulated ramp PWM generator can be observed.

Fig. 6. Switched-simulation scheme of boost converter with linear (Fig. 3) and modulated ramp (Fig. 4) PWM generators.

Fig. 7. Switched transient simulation of Fig. 6 boost converter with linear (Fig. 3) and modulated-ramp (Fig. 4) PWM generators (duty-cycle is linearly increasing with time).

II. DC AND AC CHARACTERISTICS OF THE MODULATED-RAMP BOOST SWITCHED CONVERTER

The boost power converter with modulated-ramp PWM generator exhibits constant $V_{\text{OUT}}/I_{\text{CON}}$ conversion gain. However, the parasitic elements in the circuit (mainly the inductor’s series resistance $R_L$, see Fig. 1) introduce extra higher-order terms to transfer characteristics. Due consideration for these parasitics is important for the proper design of the system, and is the matter of the following paragraph. Here, the comparison of non-ideal boost converter DC and AC characteristics is provided between the conventional linear (Fig. 3), and modulated-ramp PWM generators (Fig. 4).

TABLE I

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Linear PWM</th>
<th>Modulated-ramp PWM</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\frac{V_{\text{OUT}}}{V_{\text{IN}}}$</td>
<td>$\frac{I_{\text{CON}} T}{V_b C}$</td>
<td>$\frac{V_b T}{V_b C}$</td>
</tr>
<tr>
<td>$G_C$</td>
<td>$\frac{V_b T}{V_b C}$</td>
<td>$\frac{V_b T}{V_b C}$</td>
</tr>
</tbody>
</table>
CONVERSION RATIO $V_{OUT}/V_{IN}$ AND CONTROL GAIN $G_C$ OF BASIC POWER STAGES OPERATING IN CCM

<table>
<thead>
<tr>
<th>Power stage</th>
<th>Linear PWM</th>
<th>Modulated-ramp PWM</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>conversion ratio $V_{OUT}/V_{IN}$</td>
<td>conversion ratio $V_{OUT}/V_{IN}$</td>
</tr>
<tr>
<td>Boost</td>
<td>$1/(1 − D)$</td>
<td>$I_{CON}T/V_C$</td>
</tr>
<tr>
<td>Buck/Boost</td>
<td>$1/(1 − D)$</td>
<td>$I_{CON}T/V_C$</td>
</tr>
<tr>
<td>Sepic</td>
<td>$D/(1 − D)$</td>
<td>$I_{CON}T/V_C + 1$</td>
</tr>
<tr>
<td>Buck</td>
<td>$D/(1 − D)$</td>
<td>$I_{CON}T/V_C + 1$</td>
</tr>
</tbody>
</table>

A. Small-signal Model of the Boost Converter

The inherent nonlinear behaviour of the Boost converter is treated in widely available literature, providing methods which allow linear (small-signal) representation of switched converters.

![Diagram of Boost Converter](image)

**Fig. 8.** Linear model of boost converter power stage. $V_{DC}$ and $I_{IN}$ forms the “DC transformer” [2]. In following, we consider: $V_{IN} = 10V$, $L = 5\mu H$, $C_1 = 22\mu F$, $R_L = 150m\Omega$, $R_{ESR} = 20m\Omega$, $R = 100\Omega$, $V_S = 0.5V$, $C = 1pF$, $T = 0.3125\mu s$ ($\alpha = 1.6\mu A$), $R_{SW,D} = 0$.

Generally, approaches to the linear representation are based on averaging, provided either in the state-space (State Space Averaging - SSA) [2], [10] or through the averaged switch model [12] (sometimes presented as “DC transformer” [2]). The advantage of the averaged switch model is its compatibility with circuit simulation environments. The model of the boost DC/DC converter shown in Fig. 8, which is based on this concept, is described e.g. in [2]. Here, the dependent sources $V_{DC}$ and $I_{IN}$ represent the average voltage and current of the switch terminals (averaged over one clock period), variable $D$ is the duty-cycle, and the rectangle representing $E_{(4)}$ embodies the modulated ramp PWM generator. The switches’ resistance $R_{SW}$ in Fig. 8 model are shown for convenience and are not considered in the following. (An efficient method to embody the nonlinear resistance of semiconductor switches is presented in [14]). Generally, the accuracy of AC simulation with the given averaged models decreases at frequencies approaching the switching frequency [1], [11]. This is to be considered in the evaluation of the following AC characteristics, where the switching frequency is not specified.

B. Static $V_{OUT}/V_{IN}$ Conversion Characteristics

The steady-state analysis of Fig. 8’s circuit model leads to an improvement in the accuracy of $V_{OUT}/V_{IN}$ conversion ratio (5). We focus on higher order terms which cause the spurious drop of $V_{OUT}$ towards zero at high duty-cycle [1]. Beyond the excessive increase of the inductor and switch currents, the high duty cycle value (above $D_{MAX}$, see Fig. 9) causes the inversion of the control gain polarity, which can affect the stability of the converter operating in closed-loop. In practical applications, a positive control gain $G_C$ is usually maintained by employing an advanced control technique, such as current-mode control [1], [10], [13]. In such techniques, the inductor current $I_L$ is measured and limited to a certain value (below $D_{MAX}$) by an additional current-sensing circuit. However, an accurate current sensing circuit may be difficult to integrate at high switching frequencies, which limits the maximum switching frequency of the converter [15]. Compared to linear PWM schemes (Fig. 3), the modulated-ramp PWM generator shown in Fig. 4 provides a method to maintain a positive small-signal control gain $G_C$, and limits the maximal current $I_L$ by the natural limitation of the duty-cycle, via the maximal value of $I_{CON}$.

The $V_{OUT}/V_{IN}$ conversion ratio of a boost converter power stage with linear PWM modulator containing parasitic resistance $R_L$ can be derived from Fig. 8 model as:

$$V_{OUT} = \frac{V_{IN}}{1 − D + \frac{R_L}{R (1 − D)}}$$

![Graph of Output Voltage vs Duty Cycle](image)

**Fig. 9.** Comparison of real ($R_L = 150 m\Omega$) and ideal ($R_L = 0$) conversion characteristics of boost converter with linear PWM generator and load resistance $R = 100 \Omega$.

which, for the ideal case ($R_L \rightarrow 0$) converges to (1). An example of the conversion characteristic (7) for real ($R_L > 0$) and ideal ($R_L = 0$) power stages are shown in Fig. 9. Here we notice the mentioned point $D_{MAX}$ ($dV_{OUT}/dD = 0$) where the control gain $G_C$ changes the polarity.

In comparison, the boost converter power stage with modulated-
The considered dynamics of maximal control current 
conversion ratio

\[ V_{OUT} = \frac{V_{IN}}{\alpha + R L_1^2 I_{CON}} \]  

(8)

For low control current values, this approximates a linear function of 
I_{CON} (Fig. 10). The maximum allowed control current I_{CON(max)} which corresponds to the value where the PWM generator produces the duty-cycle value D_{MAX} results from (8) as:

\[ I_{CON(max)} = \frac{V_{OUT}}{\sqrt{R_L} \cdot \sqrt{V_C}} = \alpha \frac{R}{\sqrt{R_L}} \]  

(9)

As an illustration of Eq. (9), we can refer to a power stage with parameters listed in the caption of Fig. 8. This power stage exhibits the point \( dV_{OUT}/dI_{CON} = 0 \) at I_{CON(max)} = 41 µA. As we can see from the conversion characteristic shown in Fig. 10, this point is far from the considered dynamics of I_{CON}, being between 1.6 µA and 14 µA (conversion ratio V_{OUT}/V_{IN} from 1 to 8). For a safe choice of maximal control current value I_{MAX}, we can mention the expression of the maximal inductor current I_{max} [3] \( (D_{MAX}, I_{max} \) refers to the maximal values defined by design):

\[ I_{L_{max}} = \frac{V_{OUT}}{R \left(1 - D_{MAX}\right)} = \frac{V_{OUT} I_{max}}{\alpha R} \]  

(10)

The transfer function \( G_C(s) \) is obtained by the linearization of the converter around an operation point \( (I_{CON}, D, V_{IN}, I_L, \text{ and } V_{OUT}) \). Thus, AC analysis of the circuit in Fig. 8 requires first a steady-state analysis, in order to determine the gain of the dependent sources \( V_{PC} \) and \( I_{PA} \). For the purpose of AC analysis, the dependent sources \( V_{PC} \) and \( I_{PA} \) in Fig. 8 model are defined as follows (we neglect \( R_{SW,D} \)): 

\[ V_{PA} = V_{OUT} D + V_{OUT} d \text{ and } I_{PA} = I_L + I_{L} d \]  

Here \( V_{OUT}, I_L \) and \( D \) are the steady-state (constant) values, \( I_{L} \) and \( d \) are the AC values, and \( d \) is the AC control input. 

An analysis of the linear model of Fig. 8 results in the transfer function \( G_C(s) \) for a modulated ramp and linear PWM generator as \( V_{OUT}(s)/I_{CON}(s) \) and \( V_{OUT}(s)/D(s) \), respectively. Both transfer functions can be expressed in the following polynomial form:

\[ G_C(s) = G_C \left(1 + \frac{s}{z_1} \right) \left(1 - \frac{s}{z_2} \right) \]  

(11)

where \( z_1 \) is the left and \( z_2 \) right half-plane real zero frequencies, \( Q \) is the quality factor, and \( \Omega_0 \) the resonant frequency of the 2nd order denominator polynomial. The coefficients corresponding to the transfer function of power stages with both types of PWM modulators are listed in Tab. II.

**TABLE II**

<table>
<thead>
<tr>
<th>Coef</th>
<th>Linear PWM generator</th>
<th>Modulated-ramp PWM generator</th>
</tr>
</thead>
<tbody>
<tr>
<td>( z_1 )</td>
<td>( \frac{1}{R_{ESR} C_1} )</td>
<td>( \frac{1}{R_{ESR} C_1} )</td>
</tr>
<tr>
<td>( z_2 )</td>
<td>( \frac{R(1-D)^2-R_L}{L_1} )</td>
<td>( \frac{\alpha^2 R}{L_1} + \frac{R_L}{I_{CON}} )</td>
</tr>
<tr>
<td>( \Omega_0 )</td>
<td>( \Omega_0 L_1 C_1 (R + R_{ESR}) )</td>
<td>( \Omega_0 L_1 C_1 (R + R_{ESR}) )</td>
</tr>
<tr>
<td>( Q )</td>
<td>( C_1 R (R_{ESR} (1-D)^2 + R_L) + L_1 )</td>
<td>( C_1 R (R_{ESR} (1-D)^2 + R_L) + L_1 )</td>
</tr>
</tbody>
</table>

The DC control gain \( G_C \) in Eq. (11), which in the ideal case was given by Eqs. (2) and (6), can now be derived for the linear PWM generator from the \( V_{OUT}/V_{IN} \) conversion characteristic (7) as:

\[ G_C = \frac{\partial V_{OUT}}{\partial D} \bigg|_{V_{IN \text{const}}} = \frac{(1-D)^2 - R_L/R}{(R(1-D)^2 + R_L/R)^2} V_{IN} \]  

(12)

and for the modulated ramp PWM generator from (8) as:

\[ 

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\[ G_c = \left. \frac{\partial V_{OUT}}{\partial I_{CON}} \right|_{I_{CON}=\text{const}} = \frac{V_{IN}}{\alpha} \left( 1 - \frac{R I_{CON}^2 \alpha}{\alpha^2 R} \right) \] (13)

Compared to this, the boost converter with conventional linear PWM generator exhibits a control gain variation in the order of 24dB (for identical conversion ratios \(V_{OUT}/V_{IN}\) as in Fig. 11). This is graphically shown in Fig. 12.

III. IMPLEMENTATION CONSIDERATIONS AND EXPERIMENTAL RESULTS

As shown in Fig. 3, the modulated-ramp PWM generator requires the integration of the controlled current source \(I_{CON}\), comparator, and a constant bias voltage source \(V_b\). As follows from Eq. (5), the bias voltage \(V_b\) can be used to adjust the control gain, if necessary.

For very low \(V_b\), the control current \(I_{CON}\) can be obtained directly from a control voltage by a large value series resistance (RC integrator). In other cases, an active current source may be used, in order to achieve linear conversion \(V_{CON} \rightarrow I_{CON}\). Two examples are the non-inverting Howland current source [16] and the current conveyor (CCII+) based circuit shown in Fig. 13 [17]. In the circuit shown in Fig. 13, the conversion transconductance is simply given by \(I_{CON} = V_{CON}/R\).

Fig. 13. Single ended current source based on the second-generation CC [17].

Alternatively, a very simple realization can be achieved by using the inverting integrator shown in Fig. 14. Except for the inversion property of this circuit (inverted \(V_{CON}\)), the benefit from the low output impedance of the output node is that it allows the elimination of parasitic load capacitance.

It should also be noted that most of today’s PWM generators use the triangular wave generated PWM [18] rather than this one, based on the sawtooth-wave ramp. This avoids excessive current during the discharging of the integrating capacitor. The presented technique of modulated ramp PWM can be adapted to the modulated triangle-wave amplitude, which extends the possible application of the presented linearization technique.

Fig. 14. Inverting modulated-ramp PWM generator. Generated ramp is available on the operation amplifier output.

In a closed-loop control scheme, the current \(I_{CON}\) can be generated
directly by the controller with the current output terminal. For example, an active current-output derivative-filtered PID controller is shown in Fig. 14, based on a grounded RLC circuit. The grounded inductor $L$ forms the integrator, $R_1$ is related to the proportional gain and $R_2 C_2$ to the filtered derivative component of the controller transfer function [19].

The inductorless circuit equivalent to Fig. 15 can be obtained by topological transformation. One realization of the active circuit based on an operational amplifier is shown in Fig. 16, where the lossy grounded inductor is realized by $R_1 C_2$ and operational amplifier [20], $C_2 R_2$ corresponds approximately to the derivative branch of Fig. 15 circuit (with identical values), and voltage divider $R_H$ and $R_2$ steps down the output voltage to the reference voltage level.

![Fig. 15. Behavioral model of the current-output derivative-filtered PID controller with CCCS (current-controlled current source).](image)

Fig. 15. Behavioral model of the current-output derivative-filtered PID controller with CCCS (current-controlled current source).

![Fig. 16. Active equivalent of Fig. 15 current-output PID controller. (R3 = 462kΩ, C3 = 185pF, R2 = 9.3kΩ C2 = 45pF, RH = 19 kΩ, R = 1 kΩ).](image)

Fig. 16. Active equivalent of Fig. 15 current-output PID controller. (R3 = 462kΩ, C3 = 185pF, R2 = 9.3kΩ C2 = 45pF, RH = 19 kΩ, R = 1 kΩ).

The set of values shown in Fig. 16 were computed to stabilize the converter model from Fig. 8 with reference gain $V_{OUT}/V_{REF} = 20$ ($L = 9.85H$, $R_1 = 231kΩ$, $R_2 = 9.3kΩ$, $C_2 = 45pF$). The operational amplifier output current can be measured by sensing the bias current as shown e.g. in [17].

IV. MEASURED EXAMPLE

A practical benefit from the use of the modulated-ramp PWM generator is shown here by measurement of a boost converter power stage. In the experiments, the PWM modulator uses the scheme from Fig. 6 with $C = 71pF$ (including parasitic capacitances), and clock frequency $f_{sw} = 500kHz$. The power stage is built using an N-channel power MOSFET transistor with low RDSON < 100mΩ and a Schottky diode rectifier. The 0.5V bias voltage $V_b$ results in $\alpha = \frac{V_b C}{T} = 17.7x10^{-6}$ (6), which gives a control current range of 30 – 150µA (conversion ratio from 1 to 8).

![Fig. 17. Measured steady-state output voltage and conversion ratio $V_{OUT}/V_{IN}$ of the boost converter with modulated-ramp PWM generator, comparison with ideal characteristic Eq. (5) ($V_{IN} = 5V$, $I_{sw} = 500kHz$, $C = 71pF$, $L = 10μH$, $C_1 = 22μF$, $R = 120Ω$).](image)

Fig. 17. Measured steady-state output voltage and conversion ratio $V_{OUT}/V_{IN}$ of the boost converter with modulated-ramp PWM generator, comparison with ideal characteristic Eq. (5) ($V_{IN} = 5V$, $I_{sw} = 500kHz$, $C = 71pF$, $L = 10μH$, $C_1 = 22μF$, $R = 120Ω$).

The comparison of measured output voltage and output voltage calculated per Eq. (5) is shown in Fig. 17. We can see the low nonlinear conversion ratio error corresponding to Eq. (8), which is caused by the low parasitic resistance of $L$ (23mΩ). The small slope error is caused by the accuracy of capacitor $C$, and can be compensated by the $V_b$.

An example of the captured waveform of the converter operating with conversion ratio $V_{OUT}/V_{IN} = 6.5$ ($V_{IN} = 5V$, $V_{OUT} = 32V$) is shown in Fig. 18. In this figure, we can observe the voltage ramp $V_C$, bias voltage $V_{in}$, switch voltage $V_{LX}$ and output voltage $V_{OUT}$. Here, the dynamics of current source $I_s$ is designed to provide high output impedance (high linearity) in the range 0 – $V_b$ (i.e. 0 – 500mV). Above, the low supply voltage and parasitic resistance causes the ramp to be nonlinear, what however, has no more impact to the conversion accuracy.

![Fig. 18. Measured waveforms of the modulated-ramp PWM generator and power stage with conversion ratio 6.5 (the ramp-linearity error in the range of interest (0-500mV) is < 1%).](image)

Fig. 18. Measured waveforms of the modulated-ramp PWM generator and power stage with conversion ratio 6.5 (the ramp-linearity error in the range of interest (0-500mV) is < 1%).
CONCLUSION

The proposed modulated-ramp PWM generator aims to achieve a linear static conversion characteristic for specified switched converters operating in continuous conduction mode (CCM). The obtained linear conversion characteristic enhances the accuracy of the output voltage control. This is very important in the design of the open-loop architecture, where the linearization lowers the sensitivity of control gain at high duty-cycle values. In this paper, analytical models describing static and dynamic behavior of the modulated ramp PWM scheme were delivered, allowing its efficient evaluation, design, and optimization.

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REFERENCES


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He is involved in the research and design of switchedmode power converters for mobile phones and tablet platforms. Past and present research activities concern mainly circuit theory, analog signal processing with a particular focus on the analog frequency filters, optimization of switched DC/DC converters, and activity on cryogenic electronic as the ultralow temperature amplifiers and superconducting rapid single flux quantum logic. Dr. Vratislav Michal received several awards as an early-stage researcher from the Czech Ministry of Education, Brno University of Technology, and Czech Radio Club.