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Stability Analysis for Model Predictive Voltage Control of Buck Converter

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***Abstract*—Model Predictive Control is a modern method of nonlinear control which gives superior dynamic performance with the cost of the increased computational burden. In this paper, a continuous control set model predictive control (CCS-MPC) strategy for a buck converter is proposed, which also limits inductor current within specified limits. The control strategy achieves a constant switching frequency with the PWM modulator, and its stability is analyzed.**

The proposed algorithm is based on the sampled data model of the buck converter, which is simplified using a polynomial approximation to reduce the computational complexity. The paper also shows how this approximation is valid for most practically designed buck converters and analyzes the computational requirement. The strategy is improved by limiting the duty ratio within the stability region. The computation is implemented within a short time with the help of hardware accelerators like floating-point unit (FPU) and trigonometric math unit (TMU) in the Delfino microcontroller from Texas instruments. The proposed control strategy is verified in simulation and compared with experimental results, and it shows good performance for both reference tracking and disturbance rejection. It is about five times faster than conventional PI with a lead controller.

***Index Terms*—CCS MPC, Continuous control set, Model Predictive Control, Computation analysis, Stability analysis, fixed switching frequency, Inductor current limit, fast control.**

I. Introduction

DC-DC buck converters are extensively used power converters in various applications. There are various control techniques for this converter, frequency domain classical control being the most widely used [1], [2]. However, conventional techniques do not fully use the converter's operational capability [3]. There are various fixed switching frequency, and variable switching frequency control techniques with particular advantages in specific applications [4]–[6]. Some strategies like current mode control (CMC), hysteresis control, and constant on-time control require analog circuits which are more susceptible to noise and difficult to change the parameters [7]. The control methods are shifting to digital implementations as it is easy to tune and change control algorithms. With improvements in the computational capability of present-day microcontrollers, various computationally challenging control techniques like model predictive control (MPC) and sliding mode control are adopted because of their advantages, [8], [9].

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TABLE I: Comparison of Transient response with literature

Reference	Settling time		Number of switching cycles	
	Load change	Ref change	Load change	Ref change
[10]	25 μs	15 μs	15	25
[11]	600 μs	-	30	-
[12]	2 ms	-	40	-
[13]	5 ms	-	100	-
PI+Lead	1.8 ms	2 ms	36	40
Proposed	400 μs	500 μs	8	10

Current improvements in computational capability have brought MPC to control power converters also [14]. The finite control set MPC (FCS-MPC) is widely explored in literature due to its simplicity. However, most FCS-MPC algorithms use different switching states as inputs in a cycle, resulting in variable switching frequency. Since the converters are usually designed to operate at a fixed switching frequency considering losses and filter design, this is a disadvantage.

It is required to consider modulating signal or duty ratio of switches as input to achieve constant switching frequency. This leads to a continuous control set of input. In [11], a linear combination of inductor current and the output voltage is controlled to a reference value by computing the optimal duty ratio. However, the model is modified considering the ESR of filter capacitance to make the calculation simpler. The paper also discusses the instability caused by this and provides the region of stability based on the weighing factor of current and voltage. The Linear time-varying model of the converter is considered with a cost function involving an error in state variables and an error in input from steady-state value. The output voltage of the buck converter can also be controlled by controlling the inductor current through MPC as the computation is easy, [15]. This strategy gives good steady-state results but slower transient response as it is an indirect control method. Above mentioned papers compute the optimal input online; there are also methods in which the optimization is done offline [10], [16], [17]. The category of MPC is called explicit model predictive control, and the optimal input is

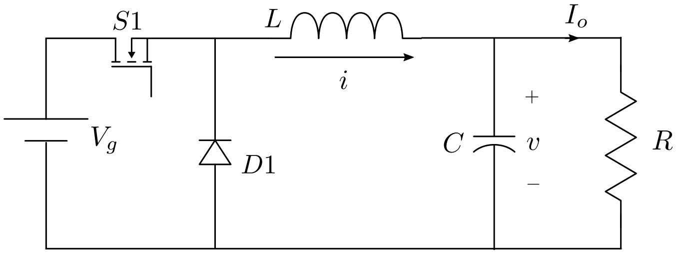


Fig. 1: Buck converter

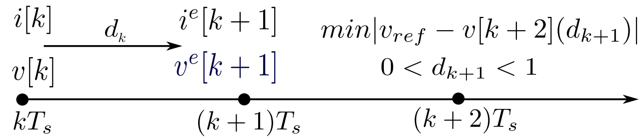


Fig. 2: MPC problem formulation for voltage control [18]

found from a look-up table that was constructed earlier.

Continuous Control Set MPC algorithm is not unique and becomes complex for non-linear systems. It is crucial to consider the non-linear nature of the switching converter to achieve better transient response [3]. A novel model predictive control strategy for a buck converter is proposed in [18] with superior transient response. However, it is not stable for the entire duty ratio from 0 to 1. This paper does analyse the stability of the CCS MPC and suggests modifications in the strategy to make it stable. Table I compares the proposed strategy with other recent MPC strategies in terms of transient response, and it can be seen that the proposed strategy is a promising solution with superior transient performance.

The paper is organized into four sections. The first section gives an introduction to model predictive control in power converters and the relevance of the proposed strategy. The second section analyses the stability of voltage control and proposes improved algorithm for the control of buck converter. Simulation and experimental results are presented in third section, fourth section discuss about the computational requirement of proposed strategy and the fifth section concludes the paper.

II. Model Predictive control for buck converter

For buck converter shown in Fig. 1 sample data model is given in (1), which is derived in [18]. The terms A, B, C, D, E and F are functions of T_s (switching time period), L (filter inductance), C (filter capacitance) and R (load resistance). Due to digital implementation delay, the duty cycle to be applied in the current cycle, d_k, is computed in the previous cycle, [19]. This implies, in the current cycle, we can only compute d_{k+1}. The objective in k^{th} cycle will be to compute d_{k+1} that minimizes the difference between the reference voltage v_{ref} and $v[k+2]$ as shown in Fig. 2.

$$\begin{bmatrix} i[k+1] \\ v[k+1] \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} i[k] \\ v[k] \end{bmatrix} + \begin{bmatrix} E(d_k) \\ F(d_k) \end{bmatrix} V_g \tag{1}$$

At the beginning of k^{th} cycle, we have sensed values of $i[k]$ and $v[k]$, we also know d_k (computed in previous cycle). So, using (1) we can estimate values of $i^e[k+1]$ and $v^e[k+1]$. From (1), $v[k+2]$ can be written as given below.

$$v[k+2] = Ci^e[k+1] + Dv^e[k+1] + F(d_{k+1}) * V_g \tag{2}$$

It turns out that $F(d_{k+1})$ is an increasing function, so if $v[k+2](d_{k+1}=0) < v_{ref} < v[k+2](d_{k+1}=1)$ then an optimal duty ratio d_{opt} can be found that will make the error zero. Otherwise, d_{opt} can be clamped to zero or one as given in (3), which will take $v[k+2]$ closer to v_{ref}.

$$d_{opt} = \begin{cases} 0, \quad v_{ref} \leq v_{k+2}^{min} = v[k+2](d_{k+1}=0) \\ 1, \quad v_{ref} \geq v_{k+2}^{max} = v[k+2](d_{k+1}=1) \\ \text{solve } v_{ref} - v[k+2](d_{k+1}) = 0, \text{ o/w} \end{cases} \tag{3}$$

Polynomial approximation for exponential and sinusoidal functions can be used to solve the minimization problem given in (3). Polynomial approximation up to second order is considered for the d dependent part of $F^{\zeta<1}$ and $F^{\zeta>1}$ denoted as $F_1(d)$. This is shown in (4), for $\zeta < 1$ and $\zeta > 1$ where $\omega = T_s/\sqrt{LC}$, $\zeta = (\sqrt{L/C})/(2R)$, damping ratio of RLC network of buck converter.

$$F_1(d_{k+1}) \approx 1 - \frac{(\omega(1-d_{k+1}))^2}{2} \tag{4}$$

This expression for $F_1(d)$ is used to solve (3). Since (3) is a quadratic equation, it has two solutions. However, there is only one valid solution within the range $(0,1)$ for it since $F_1(d)$ is a monotonic function of d. However, this strategy is not stable for the entire range of duty ratio. Fig. 3 shows limit cycle oscillations in inductor current and output voltage for a reference value of 20V, this is analyzed in next section. The strategy described so far does not consider the other state variable, inductor current. In practical implementation, it will be necessary to limit the inductor current peak to a specified value, say, i_p. Using polynomial approximation for the solution

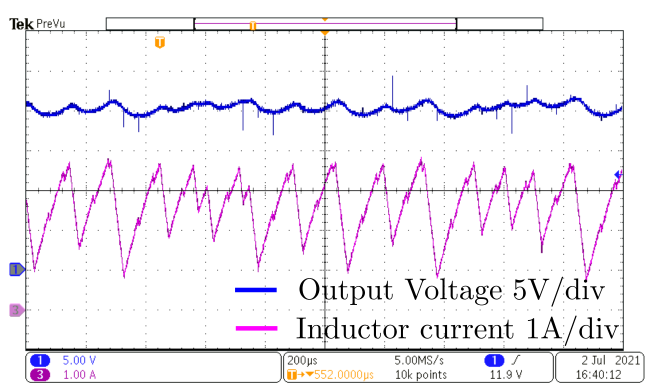


Fig. 3: Limit cycle oscillations for model predictive voltage control of buck converter at higher duty ratio ($V_g = 30V$, $v_{ref} = 20V$).

of inductor current during on state, peak inductor current at the end of on state (duration for which the switch S_1 is on in a switching time period) can be written as a function of d_{k+1} as given in (5). This relation can be rearranged with $i^e_{k+1+d} = i_p$, to find the duty ratio for limiting inductor current, as given in (6). Note, the computed value of d_{pk} using (6) can be greater than one; in that case, it must be clamped to unity. Now to limit the current, we must choose minimum of d_{pk} and d_{opt}.

$$i^e_{k+1+d} \approx i^e_{k+1} + \frac{(V_g - v^e_{k+1}) * d}{Lf_s} \tag{5}$$

$$d_{pk} = \begin{cases} 1, \text{ if } i_p \geq i^e_{k+1} + \dfrac{(V_g - v^e_{k+1})}{Lf_s} \\ \dfrac{(i_p - i^e_{k+1}) * Lf_s}{(V_g - v^e_{k+1})}, \quad \text{otherwise} \end{cases} \tag{6}$$

*1) **Stability analysis for voltage control**:* For simplicity of the analysis, let us neglect the computation delay of one sampling cycle, assume the computed duty cyce using (3) is a proper fraction. Following analysis shows how a perturbation in inductor current from a steady-state operating point is propagated while correcting the output voltage. For a steady-state operating point with state variables i^* and v^* and input d^*, the following equations hold.

$$i^* = Ai^* + Bv^* + E(d^*)V_g \tag{7}$$

$$v^* = Ci^* + Dv^* + F(d^*)V_g \tag{8}$$

If there is a perturbation in inductor current, then the instantaneous state variables in the current switching cycle are:

$$i[k] = i^* + \tilde{i_k}, \; v[k] = v^* \tag{9}$$

From this instant, control algorithm computes optimal duty ratio d_k to take $v[k+1]$ to v^*.

$$v[k+1] = v^* = C(i^* + \tilde{i_k}) + Dv^* + F(d_k)V_g \tag{10}$$

$$C\tilde{i_k} = [F(d^*) - F(d_k)]V_g \tag{11}$$

With the optimal input d_k, inductor current at the start of next cycle is taken to a different value $i[k+1]$.

$$i[k+1] = i^* + \tilde{i_{k+1}} = A(i^* + \tilde{i_k}) + Bv^* + E(d_k)V_g \tag{12}$$

$$\tilde{i_{k+1}} = A\tilde{i_k} + \{E(d_k) - E(d^*)\}V_g \tag{13}$$

From (11) and (13) ratio of inductor current perturbation is obtained as shown below.

$$\frac{\tilde{i_{k+1}}}{\tilde{i_k}} = A - \frac{C[E(d_k) - E(d^*)]}{[F(d_k) - F(d^*)]} \tag{14}$$

The algorithm is stable if the perturbation is attenuated in every cycle, ie. $|\tilde{i_{k+1}}| < |\tilde{i_k}|$. This translates to:

$$-1 < A - \frac{C[E(d_k) - E(d^*)]}{[F(d_k) - F(d^*)]} < 1 \tag{15}$$

Since $A < 1$, and the second term is positive as $E(d)$ and $F(d)$ are monotonic increasing functions of d and C is positive, the overall term is less than 1.

$$A + 1 > \frac{C[E(d_k) - E(d^*)]}{[F(d_k) - F(d^*)]} \tag{16}$$

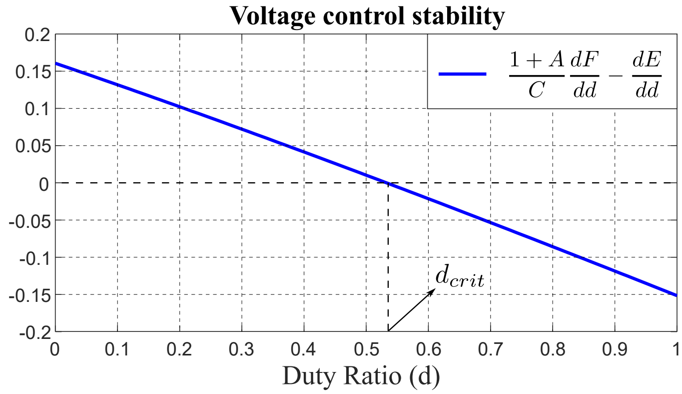


Fig. 4: Stability limit for voltage control ($d_{crit} = 0.53$)

When the perturbation is small, d_k is close to d^*; therefore the following approximations can be made:

$$E(d_k) - E(d^*) \approx \frac{dE}{dd}(d_k - d^*) \tag{17}$$

$$F(d_k) - F(d^*) \approx \frac{dF}{dd}(d_k - d^*) \tag{18}$$

From (16), (17) and (18), the algorithm is stable if:

$$\frac{dE}{dd} < \frac{1+A}{C} * \frac{dF}{dd} \tag{19}$$

$(1 + A)/C * dF/dd - dE/dd$ is a function of converter parameters. Note that this analysis is valid if we have a perturbation in output voltage alone or in both state variables because, in the next cycle, the output voltage will be taken to v^* leaving a perturbation in inductor current. From that instant, the above stability analysis is valid.

Equation (19) is a function of converter parameters ω, ζ and d^*, where d^* is the steady state duty cycle. For given converter parameters, its a monotonically decreasing function of d^* and there is a d_{crit} where it becomes zero. d_{crit} is a function of converter parameters. So stability condition is met for all $d^* < d_{crit}$. This function is plotted in Fig. 4 against d for the converter parameters given in Table. II and d_{crit} is found to be 0.53.

A. *Improved Strategy*

The critical duty ratio, d_{crit} can be found by solving the following equation given in (20).

$$\frac{dE}{dd} = \frac{1+A}{C} * \frac{dF}{dd} \tag{20}$$

The above equation is a transcendental equation, so finding the exact solution is quite difficult. An approximate expression for d_{crit} is obtained using polynomial approximation for the terms dE/dd and dF/dd. This is given below in (21). The expression is same for both the cases $\zeta < 1$ and $\zeta > 1$. The approximate solution is compared with the exact solution for a wide range of omega and zeta, giving a close match.

$$d_{crit} \approx 1 - \frac{C}{(1+A)2\omega\zeta R} \tag{21}$$

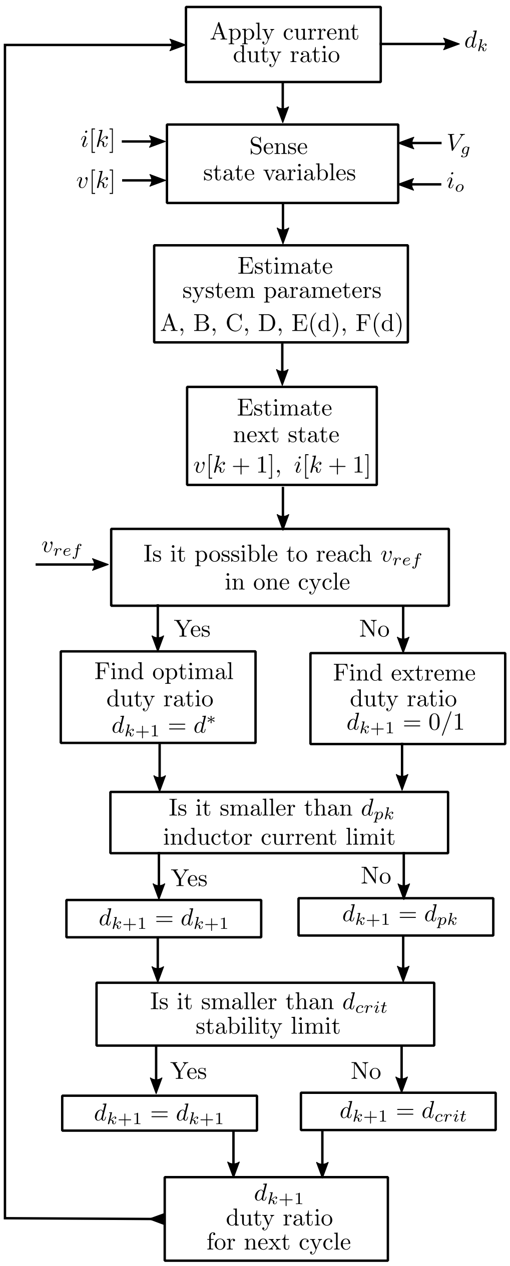


Fig. 5: Flow chart for the proposed algorithm

The modified strategy is d_{k+1} = min(d_{opt}, d_{pk}, d_{crit}). The flow chart for the algorithm is given in Fig. 5. If the reference voltage v_{ref} is such that $v_{ref}/V_g < d_{crit}$ then we can get to a stable steady state operation with d_{k+1} = min(d_{opt}, d_{pk}, d_{crit}). In the other case of $v_{ref} > d_{crit} * V_g$ inductor current is controlled to i_{ref} which is a stable strategy. With this, a stable control strategy is proposed for control of buck converter with superior transient performance.

B. *Offset free operation*

Steady-state error is a common problem seen in predictive control due to model mismatch [12]. There are many existing techniques to overcome this problem; one such solution is the addition of a PI controller, which is very slow. The additional PI controller contributes a part of the input, which doesn't significantly affect the transient response of the controller but eliminates offset in the output variable. Another existing solution in the literature is to add a slow outer PI/PID controller, which will feed reference to the MPC controller. This reference will vary very slowly, not contributing to the transient response of MPC but only affecting in steady state so that the output is offset free [11]. There are also more complex techniques involving Kalman filter to achieve offset-free performance, [13]. For simplicity first technique is used in this paper as it serves the purpose.

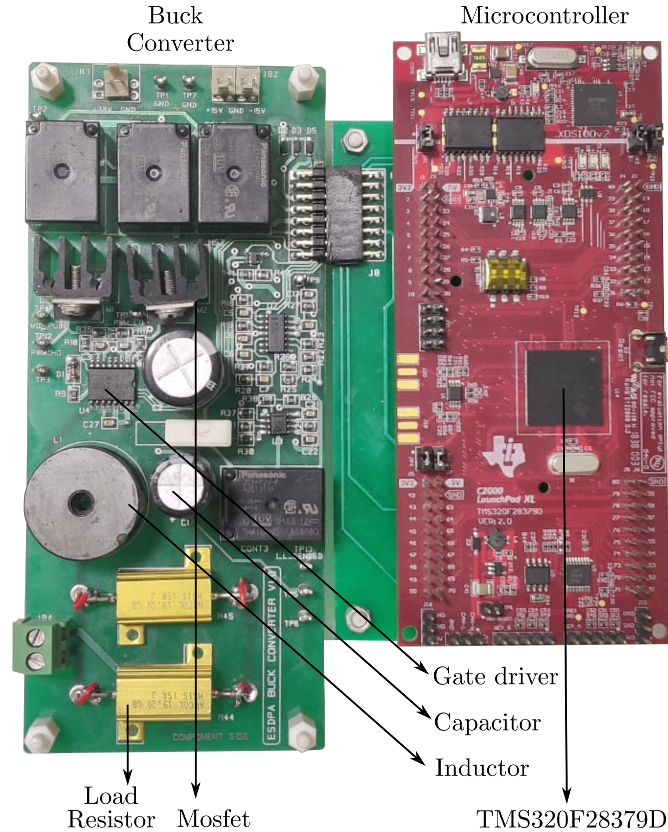


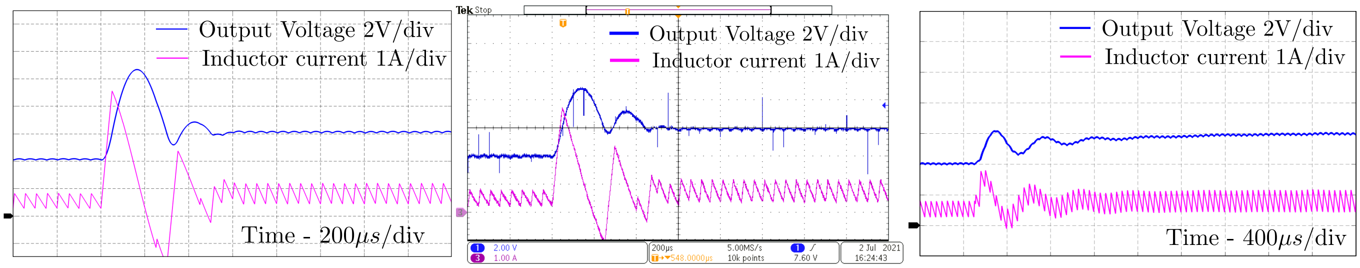
Fig. 6: Experimental set up

III. Simulation and Experimental Results

TABLE II: Converter Parameters

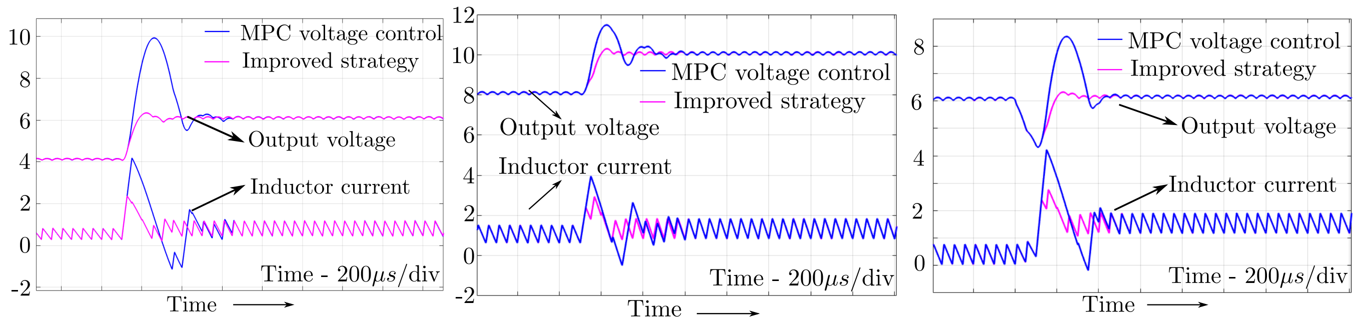
Parameter	Symbol	Value
Input Voltage	(V_g)	30 V
Filter Inductance	(L)	330 μH
Filter Capacitor	(C)	47 μF
Load Resistance	(R)	7.5 Ω
Switching frequency	(f_s)	20 kHz
Update period	(T_s)	50 μs

Texas Instruments' microcontroller $TMS320F28379D$ (Delfino) is used to control the buck converter. Experimental setup for buck converter and controller is shown in Fig. 6



(a) MPC Simulation results (b) MPC Experimental results (c) PI with lead controller

Fig. 7: Response for a step change in reference from 4V to 6V



(a) Step change in ref from 4V to 6V (b) Step change in ref from 8V to 10V (c) Load change from 15Ω to 4 Ω

Fig. 8: Comparison of transient response between MPC voltage control and Improved strategy strategy (Simulation)

and parameter values are given in Table II. It includes Buck converter hardware and a microcontroller launchpad. $IR2100$ gate driver IC is used to drive $IRF840$ mosfet, and a body diode of $IRF840$ is used in place of a diode in the converter. The states, inductor current, and output voltage after signal conditioning are passed to the ADC of the microcontroller. Computation is done in the microcontroller, and it generates a PWM signal corresponding to the MPC algorithm. This signal is then passed to the gate driver of the mosfet in the buck converter.

Both MPC strategies are verified and compared to PI with a lead controller. Simulation is done in MATLAB Simulink. This validates the proposed strategy.

The reference voltage is changed from 4V to 6V, and the results are shown in Fig. 7. The controller corrected the output voltage to the new reference value within 8 to 10 switching cycles or 500μs. It is inferred that the proposed MPC algorithm has a superior dynamic performance compared to classical PI with a lead controller in terms of the speed of response. For a step change in reference voltage from 4V to 6V, the response of PI with lead controller settled within 2.5ms, which is around five times that of the MPC controller.

From Fig.7 it is observed that for a step-change in reference from $4V$ to $6V$, there are overshoots in this strategy. A closer analysis highlights the peak in inductor current. It is because of the aggressive nature of model predictive control, to change output voltage in a single step, the controller demands a higher duty ratio. However, it settles down to the required value after a period of time.

A. *Results for improved Strategy*

Although the transient response is fast, the previous strategy has an overshoot in inductor current and output voltage. In this section, simulation results of imrpoved strategy is shown. From the figure 8 it can be seen that, since the duty ratio is limited within the stability range, the transient response settles quickly much faster than first strategy. The transient response for reference tracking and disturbance rejection happens very fast with very little overshoot, which is desirable.

IV. Computational Requirement

These advantages, however, come with the cost of an increased computational burden for MPC control. Microcontroller TMS320F28379D was used with a clock frequency of 100 MHz to compare control strategies. Computation time for the MPC algorithm was much higher than PI with the lead controller, which had very few calculations. It had only about six multiplications and a few additions to be carried out. This whole computation was executed within about 1.35 μs for PI with a lead controller.

However, the computational requirement of the MPC algorithm was significant. Following are the required computations to be carried out to implement MPC for the buck converter.

- 6 Trigonometric $\sin$ computations.
- 4 Trigonometric $\cos$ computations.
- 50 Floating point multiplications.
- 12 Floating point divisions.
- 3 Floating point square root.
- 1 Exponential computation.

One exponential computation alone took about $4\mu s$ using an exponential function from math.h library. Trigonometric and square root computations were done faster with the help of the hardware accelerator TMU, which is present in C2000 microcontrollers. The hardware accelerator improves the computation speed of complex math operations like trigonometric, square-root, and floating-point divisions. With 100 MHz clock frequency using microcontroller from texas instruments $TMS320F28379D$, the computation time for MPC algorithm was 12.5 μs, which was roughly ten times that is required for the classical controller, PI with lead compensator.

V. Conclusion

This paper proposes a continuous control set model predictive control (CCS MPC) strategy to control a buck converter's output voltage in continuous conduction mode. The prediction horizon is chosen to be unity with the update rate same as the switching frequency. This strategy employs the sample data model of a buck converter based on the exact solution of circuit dynamics. Plant variations because of load change were accounted for by sensing the load current and calculating load resistance. The strategy takes care of one cycle delay in applying the optimal duty cycle due to the finite computation time required to determine its value by the microcontroller. The controller could correct the error within a few switching cycles. The results show that the proposed controller achieves almost five times faster error correction rate when compared with a classical controller (PI with lead). In the improved strategy, the duty cycle was limited within the stable operating region to have a better transient response within the region of stability. The stability region was expanded to the full operating region by incorporating current control for higher duty ratio. The steady-state performance of the controller is similar to the conventional controller maintaining a constant duty ratio without any limit cycle oscillations. It is experimentally verified that it is possible to implement this computationally involved algorithm with the present-day high-performance micro-controllers quite effectively. The computation time for the proposed algorithm was about ten times that of the PI with a lead controller, indicating its computational complexity.

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