Controlling a DC-DC Converter by using the power MOSFET as a voltage controlled resistor

Author

Smith, TA, Dimitrijev, S, Harrison, HB

Published

2000

Journal Title

IEEE Transactions on Circuits and Systems Part 1: Fundamental Theory and Applications

DOI

https://doi.org/10.1109/81.841918

Controlling a DC–DC Converter by Using the Power MOSFET as a Voltage Controlled Resistor

Trevor A. Smith, *Member, IEEE*, Sima Dimitrijev, *Member, IEEE*, and H. Barry Harrison

***Abstract*—Most converter designs assume that a closed power switch has zero volts across it. In general, this is a valid assumption that reduces the design complexity. However, the fact that a power switch does have a finite resistance means that there will be a nonzero voltage across it during its *on time.* This voltage can be taken advantage of. This paper proposes a simple control technique that utilizes the variable resistance of the power MOSFET in a dc–dc converter. This is the first switched mode power supply that uses the power switch in more than two states or operating points. It is also the first switched mode power supply that uses the power switch as a variable control device as well as a power device. A 48-5-V 20-W forward converter is implemented to confirm the theory and demonstrate its practicality. The proposed technique provides self oscillation, self overload protection, zero voltage switching (ZVS), input voltage *feedforward,* and a reduced component count and cost.**

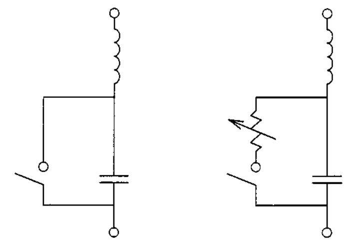
***Index Terms*—Current control, MOSFET circuits, oscillators, resonant power conversion.**

I. INTRODUCTION

ZERO voltage switching (ZVS) and zero current switching (ZCS) techniques are used to improve the efficiency of dc–dc converters [1]–[3]. Resonant components can allow either the switch voltage or switch current to be zero before the control system changes the state of the switch. A switch with its associated resonant components is called a resonant switch.

There are many variations of resonant switches that allow for ZVS or ZCS [1]–[3]. Generally, a resonant switch is modeled as an ideal switch with reactive components. A typical ZVS resonant switch is presented in [1]–[3] and shown in Fig. 1(a). Parasitic inductances and capacitances can be utilized to increase a converter's efficiency and reduce the component count [1]–[3]. What is not so common, however, is the use of parasitic resistances. Whether it is important to include this resistance in the model is totally dependent on the application.

MOSFET's are commonly used in the resonant switch of a converter. A MOSFET displays the characteristics of a voltage controlled resistor for small drain to source voltages ($v_{\rm DS}$). Fig. 1(b) shows the proposed resonant switch model that includes the variable resistance of the MOSFET. This paper presents a novel technique for self oscillation and regulation of a converter by varying the MOSFET's on resistance ($R_{\rm ON}$). Self oscillation is achieved by switching the MOSFET on and off, based on the comparison of the switch voltage $v_{\rm DS}$ to a constant threshold voltage of an inverter ($V_{\rm TI}$). Regulation is achieved by varying the total on time ($t_{\rm on}$) of the MOSFET, by only partially switching it on. A smaller gate-to-source voltage ($v_{\rm GS}$) means a larger $R_{\rm ON}$, which further means that $v_{\rm DS}$ reaches $V_{\rm TI}$ sooner, thereby decreasing $t_{\rm on}$.



(a) (b)

Fig. 1. Resonant switch models for converter design. (a) Typically used ideal switch. (b) Switch with a variable resistance, used for the first time in this paper to provide simple self oscillation and regulation for a converter.

The main advantage of using the power MOSFET as a voltage controlled resistor is that the control of the converter is simplified. This results in a reduction of the parts count, complexity and cost of the converter. This is because no external oscillator, timing circuitry, or control IC are required.

II. THEORY OF OPERATION

Fig. 2 shows a block diagram of the proposed converter. The most important difference between this circuit and conventional forward converters [4] is in the inverter. The inverter is common to the self oscillator and the regulator. The inverter's output is either zero volts or the positive supply of the inverter (V_+), depending on $v_{\rm DS}$ being greater than or less than $V_{\rm TI}$, respectively. The inverter is designed to tolerate the resonant peak voltages and have sufficient drive capability for the chosen MOSFET.

A. Self Oscillation

Many power supplies were self oscillating before the advent of integrated circuits. A typical resonant converter has variable on and off times. A zero volts detecting circuit must be included to ensure ZVS for variables such as input voltage and load. Self-oscillating converters have an inner control loop which increases the reliability of ZVS. Higher switching frequency can be achieved if this loop delay is minimized. An added advantage of self oscillation is that the converter generally has fewer components.

Manuscript received July 20, 1998; revised May 12, 1999. This work was supported in part by a grant from the Australian Research Council. This paper was recommended by Associate Editor K. Halonen.

T. A. Smith is with CSIRO Telecommunications and Industrial Physics, Lindfield, NSW, Australia, 2070 (e-mail: trevor.smith@tip.csiro.au).

S. Dimitrijev and H. B. Harrison are with the School of Microelectronic Engineering, Griffith University, Nathan, Brisbane, Qld Australia, 4111 (e-mail: s.dimitrijev@me.gu.edu.au).

Publisher Item Identifier S 1057-7122(00)02315-1.

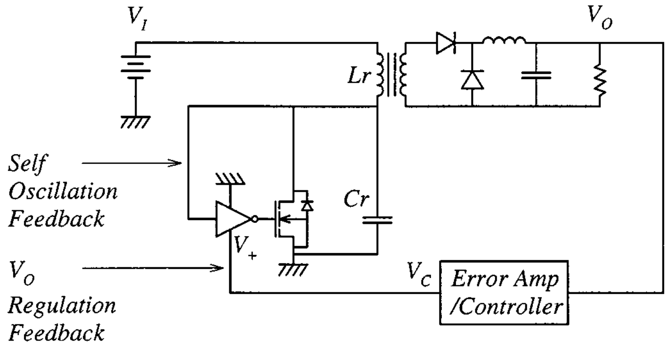


Fig. 2. A forward converter that uses the variable resistance of the power MOSFET (R_{ON}) to provide self oscillation and regulation. The resonant switch of Fig. 1(b) consists of the MOSFET, Cr, and L_r.

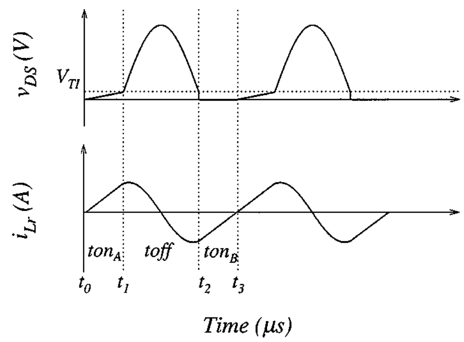


Fig. 3. Two periods of the ideal steady-state waveforms for v_{DS} and i_{L_r} (not to scale). There are three distinct time regions: t_{on_A} $(t_0 - t_1)$, t_{off} $(t_1 - t_2)$, and $t_{\mathrm{on}_B}(t_2 - t_3)$. The relative magnitude of V_{TI} is exaggerated for clarity.

Fig. 3 shows two periods of the ideal steady-state waveforms for v_{DS} and the current through the effective resonant transformer primary inductance (i_{L_r}). Self oscillation is achieved by switching the MOSFET off when v_{DS} increases to V_{TI}. The effective resonant transformer primary inductance (L_r) and the effective resonant capacitance (C_r) resonate and v_{DS} decreases back to V_{TI}, where the MOSFET is switched back on and L_r and C_r are then energized for the next cycle.

At t_0, $i_{L_r} = v_{\mathrm{DS}} = 0$ and the input of the inverter is low. The inverter output will be V_+ and the MOSFET will be on. L_r, C_r and R_{ON} are designed to be heavily overdamped and the input voltage (V_I) is much larger than V_{TI}. Therefore, i_{L_r} and v_{DS} will increase linearly until v_{DS} reaches V_{TI} at t_1. The time from t_0 to t_1 (t_{on_A}) is when the MOSFET is on and conducting.

At t_1, the inverter changes state and the MOSFET is turned off. At this point, the approximate initial conditions for C_r and L_r are, respectively

$$v_{Cr}(t_1) = V_{\mathrm{TI}} \tag{1}$$

$$i_{L_r}(t_1) = \frac{V_{\mathrm{TI}}}{R_{\mathrm{ON}}}. \tag{2}$$

These initial conditions have to be large enough so that v_{DS} can resonate back to V_{TI} at t_2 and still supply the required energy to the load. When $i_{L_r} = 0$ (positive gradient), then v_{DS} is a minimum. Therefore, in order to have sustained oscillations v_{DS} must resonate back to V_{TI} before i_{L_r} passes through zero. The time from t_1 to t_2 (t_{off}) is the resonant stage and the MOSFET is off.

At t_2, the inverter changes state once again and the MOSFET is turned on. The body diode of the MOSFET clamps v_{DS} at approximately zero volts and the circuit is operating in voltage half wave mode. The body diode supplies the linear i_{L_r} until this current passes through zero at t_3. The time from t_2 to t_3 (t_{on_B}) is when the MOSFET is on but the body diode is conducting. The cycle then repeats.

B. Inherent Overload Protection

Most converter applications require some form of overload protection. This is usually implemented with a discrete circuit or built into the control IC. These practices can contribute to the overall size and cost of the converter.

This converter has inherent overload protection. At the designed full power limit, t_{on_B} is zero. This is because v_{DS} is a minimum when it equals V_{TI} and therefore $i_{L_r} = 0$. If the load is increased beyond its designed limits, then too much current will be drawn and i_{L_r} will cross zero before v_{DS} reaches V_{TI} and oscillations will cease. This mechanism shuts the converter down into a stable state where $v_{\mathrm{DS}} = V_I$ and no current is drawn. A simple start-up circuit can be included if the converter is to automatically restart after an overload condition.

C. Regulation

Regulation is needed to keep the output voltage (V_O) constant for variables such as load and input voltage (V_I). In switching converters this is generally achieved by adjusting the duty cycle (D). In voltage mode control [4], a variable D is generated by comparing the control voltage (V_C) to an externally generated sawtooth wave. Only one control loop is present. In current mode control [4] V_C is compared to an internally generated current, usually the current in an inductor or the switch. This technique introduces two control loops and the presence of the inner control loop provides feedforward for faster response.

The new technique presented in this paper is best described by current mode control or, more accurately, peak current mode control. It has an inner control loop where i_{L_r} is compared to V_{TI} and V_C is used to adjust D and the peak current. During t_{on_A}, $V_C = V_+ = v_{\mathrm{GS}}$, i_{L_r}, and v_{DS} are approximately linear with gradients of V_I/L_r and $R_{\mathrm{ON}}V_I/L_r$, respectively. If, for example, V_O is too high then D and, hence, t_{on} needs to be decreased. Therefore, V_C needs to be reduced to lower v_{GS}, which in turn increases R_{ON} and the gradient of v_{DS}. This causes v_{DS} to reach V_{TI} sooner and thereby reduces t_{on} and V_O. The MOSFET is being used as a voltage controlled resistor.

With no regulation V_O would decrease when V_I decreased. The proposed control technique provides a degree of inherent regulation for changes in V_I, that is, input voltage feedforward. During t_{on_A}, v_{DS} can be expressed as

$$v_{\mathrm{DS}} \approx i_{L_r} R_{\mathrm{ON}} \approx \frac{V_I}{L_r} R_{\mathrm{ON}} \tag{3}$$

$$\therefore\ t \approx \frac{v_{\mathrm{DS}}}{R_{\mathrm{ON}}} \frac{L_r}{V_I} \tag{4}$$

$$\therefore t_{\mathrm{on}_A} \approx \frac{V_{\mathrm{TI}}}{R_{\mathrm{ON}}} \frac{L_r}{V_I}. \tag{5}$$

Equation (5) shows that t_{on_A} increases as V_I decreases. Therefore, t_{on}, D and, consequently, V_O increases to compensate for the decrease in V_I.

III. Experimental Results

A multiresonant forward converter that uses a transformer coupling of 0.9 [5], [6] was chosen to demonstrate the new control technique. This type of converter was previously controlled by PWM of the switch and provided ZVS for all loads.

A 20-W 48-5-V forward converter was designed and constructed to confirm the theory that a MOSFET's variable resistance can provide self oscillation and regulate the output voltage for a varying load and input voltage. The MOSFET used was an IRF620 which has a specified maximum R_{ON} of 0.8 Ω at 25°C and with a v_{GS} of 10 V [7]. The transformer primary was 12 μH, the secondary was 0.85 μH, the coupling was 0.9, and 1300 pF was added in parallel with the IRF620.

The new control technique was implemented on a converter [5], [6] where D was designed to be 50% at the maximum output power (P_O) and the switching frequency (fs) was at 1 MHz. Fig. 4(a) shows v_{DS}, Fig. 4(b) shows v_{GS}, while Fig. 4(c) shows i_{L_r} and a zoomed in view of v_{DS} for the maximum P_O of 20 W. Fig. 4(c) shows that 20 W is the maximum power limit because v_{DS} and the positive gradient of i_{L_r} cross zero at approximately the same time and t_{on_B} is zero. Fig. 4(c) also shows that V_{TI} is approximately 2.3 V. Since i_{L_r} is also the MOSFET drain current during t_{on_A}, then an approximate value for R_{ON} can be found

$$R_{\mathrm{ON}} = \frac{v_{\mathrm{DS}}}{i_{\mathrm{L_r}}} \qquad \text{during } t_{\mathrm{on}_A}. \tag{6}$$

From Fig. 4(c), R_{ON} is approximately equal to 1 Ω. Fig. 4(b) shows that a v_{GS} of approximately 5.8 V is required for R_{ON} to equal 1 Ω.

Fig. 5 shows v_{DS}, v_{GS}, and i_{L_r} for an intermediate P_O of 4 W. Fig. 5(c) shows that t_{on_B} now exists and R_{ON} is approximately 2.8 Ω. Fig. 5(b) shows that as P_O decreases, t_{on} decreases, fs increases, and a v_{GS} of approximately 5 V is required for R_{ON} to equal 2.8 Ω.

Fig. 6 shows v_{DS}, v_{GS}, and i_{L_r} for the minimum power of 50 mW. Fig. 6(c) shows that t_{on_B} is much larger than t_{on_A} and R_{ON} is approximately 4 Ω. Fig. 6(b) shows that a v_{GS} of approximately 4.4 V is required for R_{ON} to equal 4 Ω. From Figs. 4–6 it can be seen that to maintain V_O at 5 V from 20 W to 50 mW requires R_{ON} to be varied from 1 to 4 Ω by varying v_{GS} from 5.8 to 4.4 V, respectively.

With V_I at 48 V and V_O at 5 V, the output power ranged from 50 mW to 20 W. As the load was reduced, v_{GS} and, consequently, t_{on}, had to be reduced to keep the voltage constant at 5 V. There was, however, a minimum t_{on} below which oscillations will cease since a minimum amount of energy had to be stored in L_r and C_r to ensure v_{DS} resonated back to V_{TI}. In this case, P_O was at 50 mW when t_{on} was a minimum.

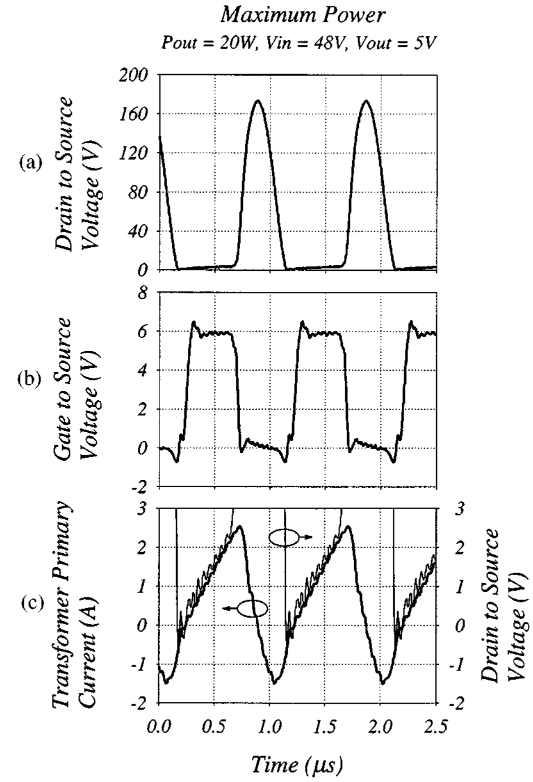


Fig. 4. Some relevant waveforms for the maximum output power of 20 W. (a) v_{DS} shows the maximum voltage stress on the MOSFET and the resonant frequency. (b) v_{GS} shows the switching frequency, duty cycle and the magnitude of v_{GS} determines R_{ON}. (c) i_{L_r} shows the peak current stress on the MOSFET and a close up of v_{DS} is used with i_{L_r} to determine an approximate value of R_{ON}. In this case it is approximately 1 Ω.

Fig. 7 shows the efficiency (η), fs, and v_{GS} for P_O ranging from 50 mW to 20 W. As expected, the efficiency falls off as P_O decreases. fs and v_{GS} are approximately linear functions of P_O where fs increases by approximately 25% from full load to no load and v_{GS} decreases by approximately 25%.

The ability of the controller to regulate for changing V_I was then investigated. The original converter [5], [6] was designed to have a maximum V_I of 48 V and a maximum P_O of 20 W. Therefore, V_I could only be reduced and P_O was held constant at 20 W. The minimum V_I was 38 V where V_O and P_O were at 5 V and 20 W, respectively. Fig. 8 shows v_{DS}, v_{GS}, and i_{L_r} with V_I reduced to 38 V. From Fig. 8(c) it can be seen that R_{ON} is approximately 0.8 Ω. Fig. 8(b) shows that t_{on} has approximately doubled, compared with a V_I of 48 V, and therefore fs decreased to approximately two thirds. For all changes in load and input voltage the converter had a near constant t_{off} of 0.5 μs. Fig. 8(b) also shows that a v_{GS} of approximately 11.2 V is required for R_{ON} to equal 0.8 Ω.

Fig. 9 shows η, fs, and v_{GS} for V_I ranging from 38 to 48 V with a constant P_O of 20 W. The efficiency falls off approximately linearly as V_I decreases. This is to be expected due to the increase in t_{on}, D and, hence, average power dissipated by R_{ON}. fs and v_{GS} are also approximately linear functions of V_I where fs decreases by approximately 40% from maximum to minimum V_I and v_{GS} increases by approximately 90%.

Fig. 10 shows the v_{DS}, v_{GS}, and i_{L_r} waveforms when the maximum power rating of the converter is exceeded. These

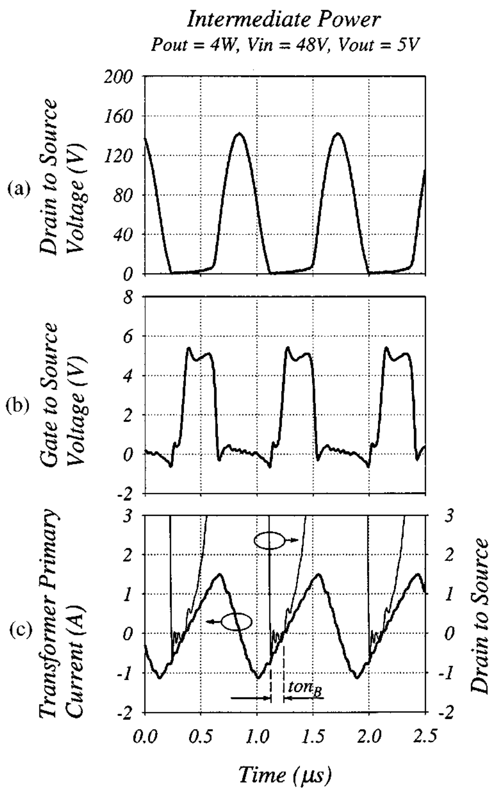


Fig. 5. Some relevant waveforms for an intermediate output power of 4 W. For a decreasing load. (a) $v_{\rm DS}$ shows that the voltage stress on the MOSFET reduces. (b) $v_{\rm GS}$ shows the switching frequency increases while the duty cycle and the magnitude of $v_{\rm GS}$ decrease. (c) i_{L_r} shows the peak current stress on the MOSFET decreases and because $v_{\rm GS}$ decreases then $R_{\rm ON}$ increases. In this case it is approximately $2.8\ \Omega$. $t_{{\rm on}_B}$ exists when the output power is less than maximum.

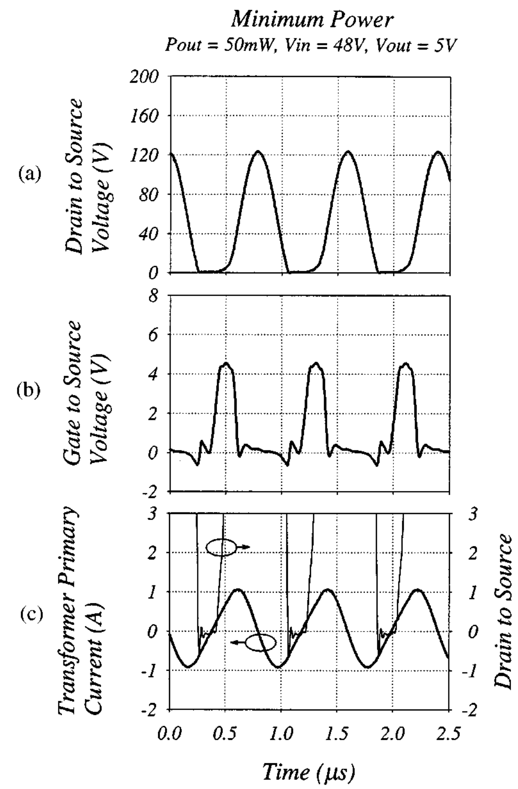


Fig. 6. Some relevant waveforms for the minimum output power of 50 mW. This is the lowest power for which the converter could be regulated at 5 V. $t_{{\rm on}_A}$ is much less than $t_{{\rm on}_B}$ and $R_{\rm ON}$ is approximately $4\ \Omega$.

waveforms are the same as those in Fig. 4 until the circuit is overloaded at time = 0. As previously explained, $v_{\rm DS}$ does not return past $V_{\rm TI}$ and oscillations cease. This provides automatic shut down of the converter for an overload condition.

The experimental results from the simple circuit of Fig. 2, clearly demonstrate that the new control technique is practical and can provide regulation for a respectable range of output power and input voltage. It is important to note that ZVS is maintained throughout the entire output power range. This is important for reduced stress on the MOSFET and increased efficiency. In addition, the experiment verified another advantage of the new concept, namely the inherent overload protection.

IV. Discussion

The new control technique can simplify a converter to the extent that a control IC is not required. This can give advantages such as improved robustness and reliability, and decreased manufacturing and maintenance cost. Perhaps the most important advantage is that the self-oscillation loop delay can be minimized to permit an increase in the switching frequency.

Self oscillation requires the sensing of $v_{\rm DS}$. Ringing of $v_{\rm DS}$ when the MOSFET is turned on can create a stability or regulation problem if the ringing crosses the inverter threshold voltage. Therefore, attention to circuit layout and parasitics is required to minimize the ringing. The threshold voltage should then be

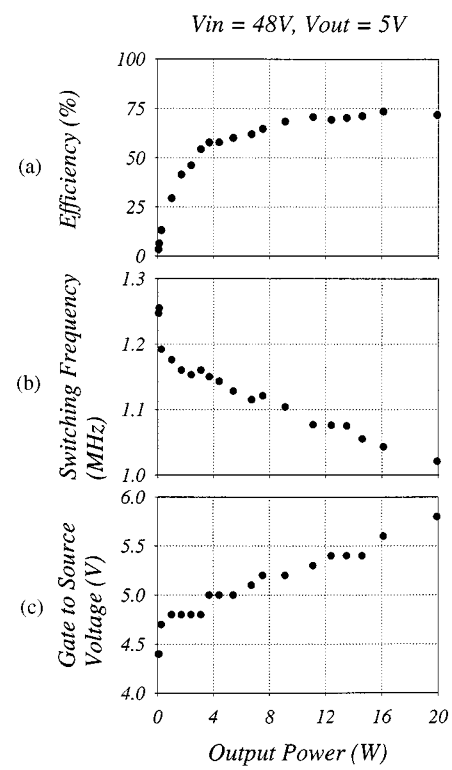


Fig. 7. (a) η, (b) fs, and (c) v_{GS} as the output power ranges from 50 mW to 20 W. ZVS occurs at all loads.

chosen to be greater than the ringing. It should also be mentioned that ringing is one of the reasons why a control based

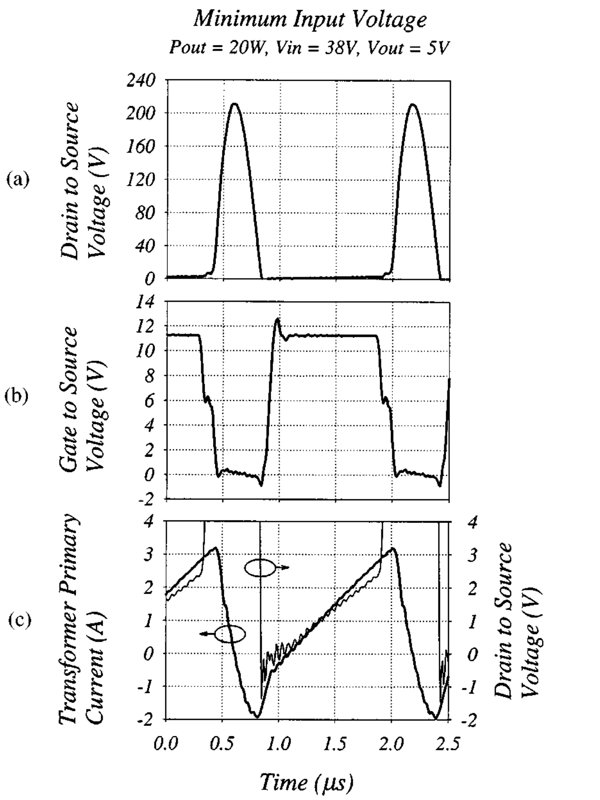


Fig. 8. Some relevant waveforms for the minimum input voltage of 38 V. (a) v_{DS} shows that the voltage stress on the MOSFET increases. (b) v_{GS} shows the switching frequency decreases while the duty cycle and the magnitude of v_{GS} increase. (c) i_{L_r} shows the peak current stress on the MOSFET increases and R_{ON} decreases to approximately $0.8\ \Omega$.

on variable threshold voltage and fixed resistance, which is an alternative control technique that is possible in self-oscillating dc–dc converters, is not considered in this paper. Such a controller would face the danger of decreasing the threshold voltage lower than the ringing voltage. This would be preventable, but the switching frequency would then become extremely limited.

A multiresonant forward converter was chosen to experimentally verify the new concept. The proposed control technique is portable and should be equally well applied to other types of resonant converters. With a minimum of redesign effort and depending on the application, it should be possible to substitute an existing resonant switch and IC with the newly proposed resonant switch and inverter. It should be noted that the new technique is intended to only replace the switching times of existing resonant controllers and therefore should generally have no effect on the specifications and waveforms of the original converter. For example, the line and load regulation of a converter will not necessarily be improved or worsened. It is essential, however, that the MOSFET and inverter threshold voltage are chosen to give the appropriate range of switching times and duty cycle as required by the original converter.

ZVS improves the efficiency of a converter by minimizing the power dissipated in the switching device. Since any resistive component will dissipate power, it is desirable to use a MOSFET with the lowest possible R_{ON}. There is a limit, however, because the smaller the R_{ON} the larger the input capacitance and the more difficult it becomes to switch the MOSFET at high frequencies. When the output power is at its maximum, the power dissipated in the MOSFET is at its highest and is the same for

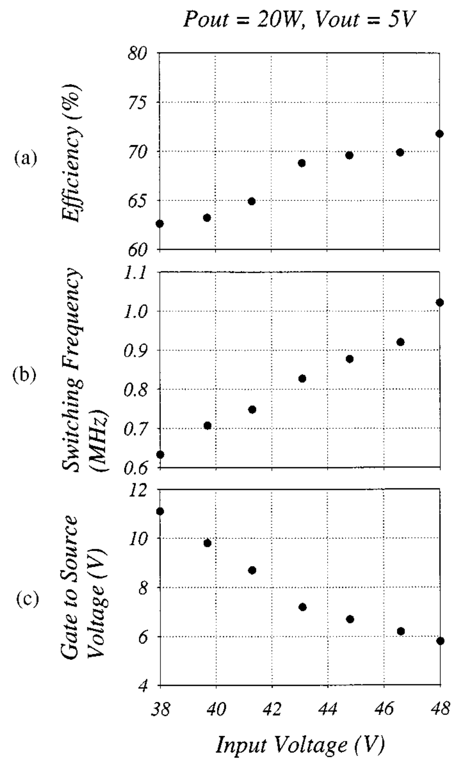


Fig. 9. (a) η, (b) fs, and (c) v_{GS} as the input voltage ranges from 38 to 48 V.

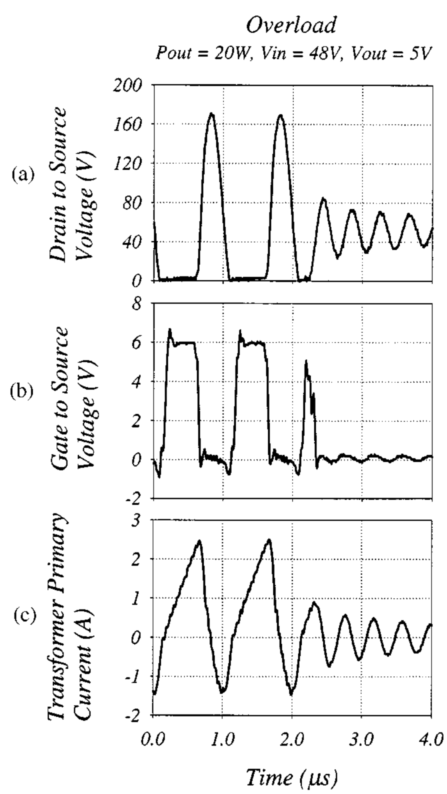


Fig. 10. Some relevant waveforms for when the load is increased beyond the design limit of 20 W. Before $0\ \mu$s the waveforms are the same as Fig. 4, at $0\ \mu$s the load is increased greater than 20 W, and after two periods v_{DS} never crosses V_{TI} (2.3 V) and oscillations cease. This provides inherent overload protection.

both the proposed converter and a conventional converter. This is because both R_{ON}'s are near their minimum. In the proposed control technique, regulation for lighter loads is accomplished by increasing R_{ON}. This characteristic does increase the instantaneous power dissipated in the MOSFET, however the conduction time also becomes reduced. Therefore, increasing R_{ON} for lighter loads has a negligible effect on the converter's total efficiency.

The proposed technique required the MOSFET to have its minimum R_{ON} for maximum output power and minimum input voltage. However, R_{ON} varies with temperature and a typical R_{ON} increases by 0.7%/°C [7]. For example, R_{ON} will increase by approximately 50% from 25 to 100°C. Therefore, the converter design should take into account the maximum expected operating temperature of the MOSFET.

V. Conclusion

This paper presented a simple control technique for dc–dc converters by using an improved resonant switch model. The model recognizes that a converter's power switch has a finite resistance in its on state. If a MOSFET is used then this resistance can be variable. A switch that has a finite resistance has a voltage across it when current passes through it. This voltage can be compared to a threshold voltage of an inverter which determines when the switch changes state. This method provides self oscillation and inherent overload protection for a converter. The variable resistance of a MOSFET can be utilized to change the time taken for the switch voltage v_{DS} to reach the inverter threshold voltage. This changes the on time and provides a regulation mechanism for variable output power and input voltage. This is the first switched mode power supply that uses the power switch in more than two states or operating points. It is also the first switched mode power supply that uses the power switch as a variable control device as well as a power device. A 48-5-V 20-W forward converter successfully demonstrated the proposed technique. The output power could be regulated from 20 W to 50 mW by changing the MOSFET's resistance from 1 to $4\ \Omega$, respectively. By decreasing the resistance to $0.8\ \Omega$ the converter could be regulated when the input voltage was reduced to 38 V. The proposed control technique is very simple and it can reduce the component count and cost of a converter.

References

[1] K. H. Liu, R. Oruganti, and F. C. Lee, "Resonant switches—Topologies and characteristics," in *IEEE 1985 PESC Rec.*, pp. 106–116.

[2] M. K. Kazimierczuk and J. Jozwik, "New topologies of high-efficiency high frequency zero-voltage-switching resonant dc/dc converters," in *Proc. 29th Midwest Symp. Circuits Systems*, Aug. 1986, pp. 474–477.

[3] K. H. Liu and F. C. Lee, "Zero-voltage switching technique in dc/dc converters," in *IEEE 1986 PESC Rec.*, June 1986, pp. 58–70.

[4] Mohan, Underland, and Robbins, *Power Electronics—Converters, Applications, and Design*, 2nd ed. New York, NY: Wiley, 1995.

[5] T. A. Smith and S. Dimitrijev, "A multi-resonant forward converter based on non-ideal coupling of the transformer," in *Proc. IEEE PEDES'98*, Nov. 1998, pp. 431–435.

[6] T. A. Smith and S. Dimitrijev, "Analysis of a multi-resonant forward converter based on non-ideal coupling of the transformer," *IEEE Trans. Power Electron.*, vol. 15, pp. 111–120, Jan. 2000.

[7] *HEXFET Power MOSFET Designer's Manual*, vol. III, 1993.

Trevor A. Smith (S'93–M'00) was born in Brisbane, Australia, in 1967. He received the B.Eng. degree in microelectronic engineering (First Class Hons.) from Griffith University, Brisbane, Australia, in 1996. He is currently working toward the Ph.D. degree at Griffith University

His research interests include resonant converters, transformer design, and electromagnetics.

Sima Dimitrijev (S'87–M'88) received the B.Eng., M.Sci., and Ph.D. degrees in electronic engineering from the University of Nis, Yugoslavia, in 1982, 1985, and 1989, respectively.

From 1982 to 1983, he was with the Semiconductor Factory of the Electronics Industry, Nis, where he worked on the development of CMOS technology. From 1983 to 1990, he was with the Faculty of Electronic Engineering at the University of Nis. In 1990, he joined Griffith University where he is currently an Associate Professor at the School of Microelectronic Engineering. He is an active researcher in the areas of semiconductor technology (including micromachining and integrated optics), MOSFET applications (in particular power electronics), as well as MOSFET design, modeling, fabrication and characterization (including SiC MOSFET's). He is the author of *Understanding Semiconductor Devices,* (NY: Oxford Univ. Press).

Dr. Dimitrijev is a member of the Editorial Board of Microelectronics Reliability, Elsevier.

H. Barry Harrison is currently Dean of the Faculty of Engineering at Griffith University and Professor of Microelectronics. He is a Director of the Microelectronic Research Centre (formerly the Space Centre for Microelectronic Technology) and is also a major contributor to the CRC for Microtechnology. He has published more than 200 papers, one book, and two book chapters. His research interests are in the areas of limits to scale of integration and contact modeling.