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# FAN501A

## Offline DCM / CCM Flyback PWM Controller for Charger Applications

### Features

- WSaver® Technology Provides Ultra-Low Standby Power Consumption for Energy Star's 5-Star Level (<30 mW)
- Constant-Current (CC) Control without Secondary-Side Feedback Circuitry for Discontinuous Conduction Mode (DCM) and Continuous Conduction Mode (CCM)
- Dual-Frequency Function Changes Switching Frequency (140 kHz / 85 kHz) According to Input Voltage to Maximize Transformer Utilization and Improve Efficiency
- High Power Density and High Conversion Efficiency in CCM Compact Charger Applications
- Frequency Hopping to Reduce EMI Noise
- High-Voltage Startup
- Precise Maximum Output Power Limit by CC Regulation through External Resistor Adjustment
- Peak-Current-Mode Control with Slope Compensation to Avoid Sub-Harmonic Oscillation
- Programmable Over-Temperature Protection with Latch Mode through External NTC Resistor
- Two-Level UVLO Reduces Input Power in Output Short Situation
- $V_S$  Over-Voltage Protection with Latch Mode
- $V_{DD}$  Over-Voltage Protection with Auto Restart
- Available in MLP 4 X 3 Package

### Applications

- Battery Chargers for Smart Phones, Feature Phones, and Tablet PCs
- AC-DC Adapters for Portable Devices or Battery Chargers that Require CV / CC Control

### Description

The advanced PWM controller, FAN501A, simplifies isolated power supply design that requires CC regulation of the output. The output current is precisely estimated with only the information in the primary side of the transformer and controlled with an internal compensation circuit, removing the output current-sensing loss and eliminating external CC control circuitry. With an extremely low operating current (250  $\mu$ A), Burst Mode maximizes light-load efficiency, allowing conformance to worldwide Standby Mode efficiency guidelines.

Compared with a conventional approach using external control circuit in the secondary side for CC regulation, the FAN501A can reduce total cost, component count, size, and weight; while increasing efficiency, productivity, and system reliability.

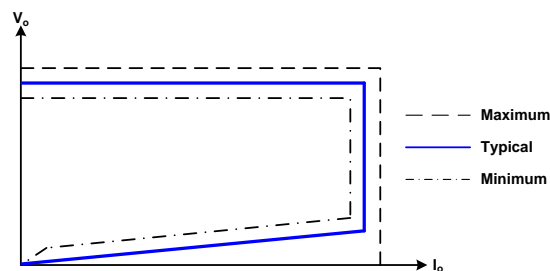


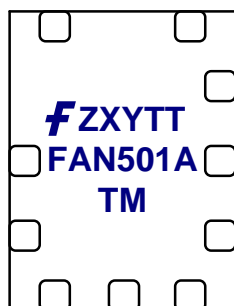
Figure 1. Typical Output V-I Characteristic

### Ordering Information

Part Number	Operating Temperature Range	Package	Packing Method
FAN501AMPX	-40°C to +125°C	10-Lead, MLP, QUAD, JEDEC MO-220 4 mm x 3 mm, 0.8 mm Pitch, Single DAP	Tape & Reel



## Marking Information



**F:** Fairchild Logo  
**Z:** Assembly Plant Code  
**X:** Year Code  
**Y:** Week Code  
**TT:** Die Run Code  
**T:** Package Type (MP=MLP)  
**M:** Manufacture Flow Code

Figure 4. Top Mark

## Pin Configuration

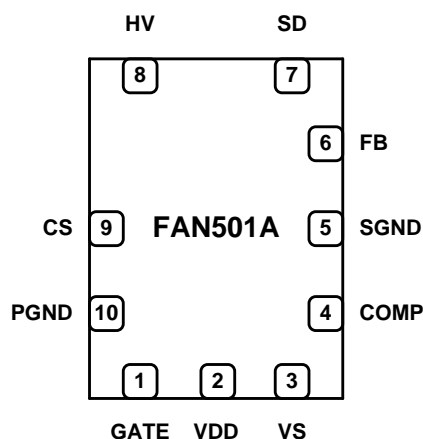


Figure 5. Pin Assignments

## Pin Definitions

Pin #	Name	Description
1	GATE	PWM Signal Output. This pin has an internal totem-pole output driver to drive the power MOSFET. The gate driving voltage is internally clamped at 7.5 V.
2	VDD	Power Supply. IC operating current and MOSFET driving current are supplied through this pin. This pin is typically connected to an external capacitor.
3	VS	Voltage Sense. This pin detects the output voltage information and diode current discharge time based on the voltage of the auxiliary winding. It also senses sink current through the auxiliary winding to detect input voltage information.
4	COMP	CC Control Correction. This pin connects to external resistor to program the CC control correction weighting.
5	SGND	Signal Ground
6	FB	Feedback. An opto-coupler is typically connected to this pin to provide feedback information to the internal PWM comparator. This feedback is used to control the duty cycle in Constant-Voltage (CV) regulation.
7	SD	Shut Down. This pin is implemented for external over-temperature protection by connecting to an NTC thermistor.
8	HV	High Voltage. This pin connects to a DC bus for high-voltage startup.
9	CS	Current Sense. This pin connects to a current-sense resistor to detect the MOSFET current for Peak-Current-Mode control for output regulation. The current-sense information is also used to estimate the output current for CC regulation.
10	PGND	Power Ground

## Absolute Maximum Ratings

Stresses exceeding the absolute maximum ratings may damage the device. The device may not function or be operable above the recommended operating conditions and stressing the parts to these levels is not recommended. In addition, extended exposure to stresses above the recommended operating conditions may affect device reliability. The absolute maximum ratings are stress ratings only.

Symbol	Parameter		Min.	Max.	Unit
V <sub>HV</sub>	HV Pin Input Voltage			500	V
V <sub>VDD</sub>	DC Supply Voltage			30	V
V <sub>VS</sub>	VS Pin Input Voltage		-0.3	6.0	V
V <sub>CS</sub>	CS Pin Input Voltage		-0.3	6.0	V
V <sub>FB</sub>	FB Pin Input Voltage		-0.3	6.0	V
V <sub>COMP</sub>	COMP Pin Input Voltage		-0.3	6.0	V
V <sub>SD</sub>	SD Pin Input Voltage		-0.3	6.0	V
P <sub>D</sub>	Power Dissipation (T <sub>A</sub> =25°C)			850	mW
θ <sub>JA</sub>	Thermal Resistance (Junction-to-Air)			150	°C/W
θ <sub>JC</sub>	Thermal Resistance (Junction-to-Case)			10	°C/W
T <sub>J</sub>	Operating Junction Temperature		-40	+150	°C
T <sub>STG</sub>	Storage Temperature Range		-40	+150	°C
T <sub>L</sub>	Lead Temperature (Wave soldering or IR, 10 Seconds)			+260	°C
ESD	Electrostatic Discharge Capability <sup>(3)</sup>	Human Body Model, ANSI/ESDA/JEDEC JS-001-2012 (Except HV Pin)		5.0	kV
		Charged Device Model, JEDEC:JESD22_C101 (Except HV Pin)		2.0	

### Notes:

1. All voltage values, except differential voltages, are given with respect to the GND pin.
2. Stresses beyond those listed under Absolute Maximum Ratings may cause permanent damage to the device.
3. ESD ratings including HV pin: HBM=3.5 kV, CDM=1.25 kV.

## Electrical Characteristics

$V_{DD}=15\text{ V}$  and  $T_J=-40\sim 125^\circ\text{C}$  unless noted.

Symbol	Parameter	Conditions	Min.	Typ.	Max.	Unit
<b>HV Section</b>						
$V_{HV-MIN}$	Minimum Startup Voltage on HV Pin				30	V
$I_{HV}$	Supply Current Drawn from HV Pin	$V_{HV}=120\text{ V}$ , $V_{DD}=0\text{ V}$	1.2	2.0	5.0	mA
$I_{HV-LC}$	Leakage Current Drawn from HV Pin	$V_{HV}=500\text{ V}$ , $V_{DD}=V_{DD-OFF}+1\text{ V}$		0.8	10.0	$\mu\text{A}$
<b>V<sub>DD</sub> Section</b>						
$V_{DD-ON}$	Turn-On Threshold Voltage	$V_{DD}$ Rising	16.0	17.5	18.5	V
$V_{DD-OFF}$	Turn-Off Threshold Voltage	$V_{DD}$ Falling	5.5	6.0	6.5	V
$V_{DD-HVON}$	Threshold Voltage for HV Startup		3.4	4.4	5.1	V
$V_{DD-DLH}$	Threshold Voltage for Latch Release			2.50		V
$I_{DD-ST}$	Startup Current	$V_{DD}=V_{DD-ON}-0.16\text{ V}$		150	250	$\mu\text{A}$
$I_{DD-OP}$	Operating Supply Current	$V_{CS}=5.0\text{ V}$ , $V_S=3\text{ V}$ , $V_{FB}=3\text{ V}$ , $V_{DD}=15\text{ V}$ , $C_{GATE}=1\text{ nF}$		3.5	4.0	mA
$I_{DD-Burst}$	Burst Mode Operating Supply Current	$V_{CS}=0.3\text{ V}$ , $V_S=0\text{ V}$ , $V_{FB}=0\text{ V}$ , $V_{DD}=V_{DD-ON}\rightarrow V_{DD-OVP}\rightarrow 10\text{ V}$ , $C_{GATE}=1\text{ nF}$		250	300	$\mu\text{A}$
$V_{DD-OVP}$	$V_{DD}$ Over-Voltage Protection Level		26.5	28.0	29.5	V
<b>Oscillator Section</b>						
$f_{OSC-H}$	Operating Frequency, $I_{VS}$ Below Threshold $I_{VS-L}$ (Low Line) <sup>(4)</sup>	$V_{CS}=5\text{ V}$ , $V_S=2.5\text{ V}$ , $V_{FB}=6\text{ V}$	133	140	147	kHz
$f_{OSC-L}$	Operating Frequency, $I_{VS}$ Over Threshold $I_{VS-H}$ (High Line) <sup>(4)</sup>	$V_{CS}=5\text{ V}$ , $V_S=2.5\text{ V}$ , $V_{FB}=4\text{ V}$	79	85	91	kHz
$\Delta f_{Hopping-H}$	Frequency Hopping Range, High Line	$V_{CS}=0.5\text{ V}$ , $V_S=0.7\text{ V}$ , $V_{FB}=3\text{ V}$	$\pm 5.5$	$\pm 7.0$	$\pm 8.5$	kHz
$\Delta f_{Hopping-L}$	Frequency Hopping Range, Low Line	$V_{CS}=0.5\text{ V}$ , $V_S=0.0\text{ V}$ , $V_{FB}=3\text{ V}$	$\pm 2.5$	$\pm 4.0$	$\pm 5.5$	kHz
$\Delta t_{Hopping}$	Frequency Hopping Period			2.54		ms
<b>Feedback Input Section</b>						
$Z_{FB}$	FB Pin Input Impedance		36	41	48	k $\Omega$
$A_V$	Internal Voltage Attenuator of FB Pin			1/2.5		V/V
$V_{FB-Open}$	FB Pin Pull-Up Voltage	FB Pin Open	5.00	5.50	5.90	V
$V_{FB-Burst-H}$	FB Threshold to Enable Gate Drive in Burst Mode <sup>(4)</sup>	$V_{FB}$ Rising with $V_{CS}=0.3\text{ V}$ , $V_S=0\text{ V}$	1.60	1.70	1.80	V
$V_{FB-Burst-L}$	FB Threshold to Disable Gate Drive in Burst Mode <sup>(4)</sup>	$V_{FB}$ Falling with $V_{CS}=0.3\text{ V}$ , $V_S=0\text{ V}$	1.55	1.65	1.75	V
<b>Over-Temperature Protection Section</b>						
$T_{OTP}$	Threshold Temperature for Over-Temperature Protection			140		$^\circ\text{C}$
<b>Shutdown Function Section</b>						
$I_{SD}$	SD Pin Source Current	$V_{CS}=0.3\text{ V}$	85	100	115	$\mu\text{A}$
$V_{SD-TH}$	Threshold Voltage for Shutdown Function Enable	$V_{CS}=0.3\text{ V}$	0.85	1.00	1.15	V

Continued on the following page...

## Electrical Characteristics

$V_{DD}=15\text{ V}$  and  $T_J=-40\sim 125^\circ\text{C}$  unless noted.

Symbol	Parameter	Conditions	Min.	Typ.	Max.	Unit
<b>Voltage-Sense Section</b>						
$I_{TC}$	Temperature-Independent Bias Current	$V_{CS}=5\text{ V}$ , $V_{FB}=3\text{ V}$	8.75	10.00	11.25	$\mu\text{A}$
$I_{VS-H}$	$V_S$ Source Current Threshold to $f_{OSC-L}$ Operation			750		$\mu\text{A}$
$I_{VS-L}$	$V_S$ Source Current Threshold to $f_{OSC-H}$ Operation			680		$\mu\text{A}$
$I_{VS-Brownout}$	$V_S$ Source Current Threshold to Enable Brownout			160		$\mu\text{A}$
$V_{VS-OVP}$	Output Over-Voltage Protection with $V_S$ Sampling Voltage <sup>(4)</sup>		3.10	3.20	3.30	V
$N_{VS-OVP}$	Output Over-Voltage Protection Debounce Cycle Counts <sup>(4)</sup>			8		Cycle
<b>Current-Sense Section</b>						
$V_{VR}$	Internal Reference Voltage for CC Regulation		2.460	2.500	2.540	V
$V_{CCR}$	Variation Test Voltage on CS Pin for CC Regulation <sup>(4)</sup>	$V_{CS}=0.375\text{ V}$ , $V_{COMP}=1.59\text{ V}$ , $V_S=6\text{ V}$	2.405	2.430	2.455	V
$K_{CCM}$	Design Parameter in CC Regulation			12.0		V/V
$V_{CS-LIM}$	Current Limit Threshold Voltage		0.80	0.85	0.90	V
$t_{PD}$	GATE Output Turn-Off Delay			100	200	ns
$t_{LEB}$	Leading-Edge Blanking Time			150	200	ns
$V_{Slope}$	Slope Compensation	Maximum Duty Cycle		66.6		mV/ $\mu\text{s}$
<b>Constant Current Correction</b>						
$I_{COMP-H}$	COMP Pin Source Current as $V_S=0.3\text{ V}$	$V_{CS}=0.3\text{ V}$ , $V_{FB}=2.5\text{ V}$ , $V_S=0.3\text{ V}$	25	35	45	$\mu\text{A}$
<b>GATE Section</b>						
$t_{ON-MIN}$	Minimum On Time	$V_{CS}=0.6\text{ V}$ , $V_S=0.3\text{ V}$ , $V_{FB}=1.7\text{ V}$	450	550	650	ns
$t_{ON-MIN-Limit}$	Limited Minimum On Time	$V_{CS}=0.6\text{ V}$ , $V_S=0.5\text{ V}$ , $V_{FB}=1.7\text{ V}$	0.95	1.20	1.45	$\mu\text{s}$
$D_{CYMAX}$	Maximum Duty Cycle	$V_{CS}=0.6\text{ V}$ , $V_S=0\text{ V}$ , $V_{FB}=4\text{ V}$	60.0	68.5	77.0	%
$V_{GATE-L}$	Gate Output Voltage Low		0		1.5	V
$V_{DD-PMOS-ON}$	Internal Gate PMOS Driver ON		7.0	7.5	8.0	V
$V_{DD-PMOS-OFF}$	Internal Gate PMOS Driver OFF		9.0	9.5	10.0	V
$t_r$	Rising Time	$V_{CS}=0\text{ V}$ , $V_S=0\text{ V}$ , $C_{GATE}=1\text{ nF}$	100	140	180	ns
$t_f$	Falling Time	$V_{CS}=0\text{ V}$ , $V_S=0\text{ V}$ , $C_{GATE}=1\text{ nF}$	30	50	70	ns
$V_{GATE-CLAMP}$	Gate Output Clamping Voltage	$V_{DD}=25\text{ V}$	7.0	7.5	8.0	V

### Notes:

4.  $T_J$  guaranteed range at  $25^\circ\text{C}$ .

## Typical Performance Characteristics

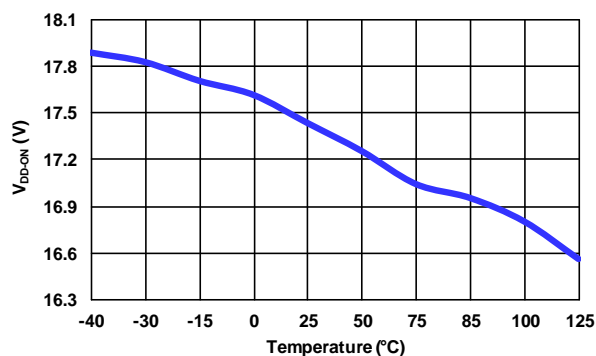


Figure 6. V<sub>DD</sub> Turn-On Threshold Voltage (V<sub>DD-ON</sub>) vs. Temperature

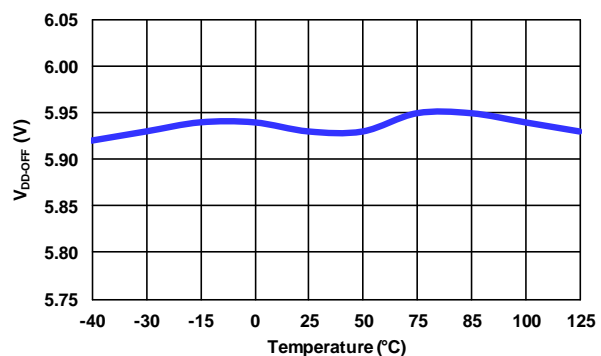


Figure 7. V<sub>DD</sub> Turn-Off Threshold Voltage (V<sub>DD-OFF</sub>) vs. Temperature

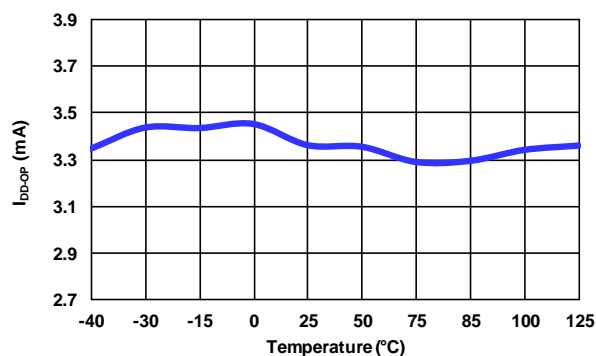


Figure 8. Operating Supply Current (I<sub>DD-OP</sub>) vs. Temperature

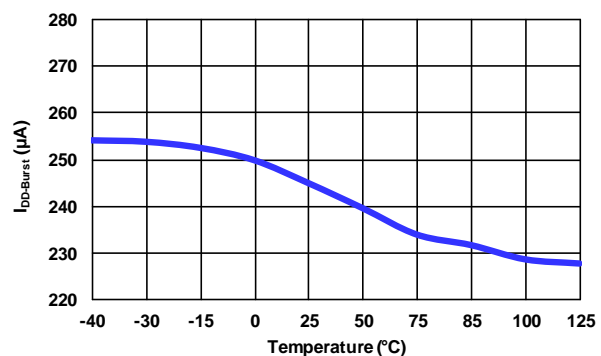


Figure 9. Burst Mode Operating Supply Current (I<sub>DD-Burst</sub>) vs. Temperature

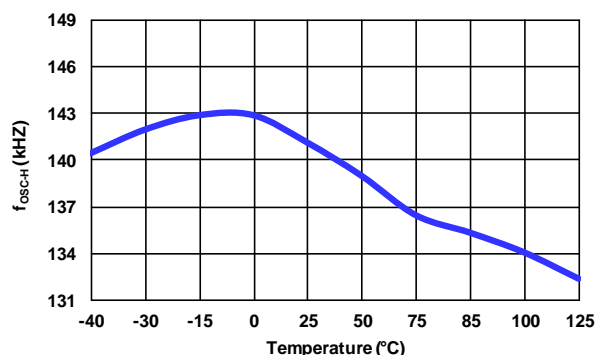


Figure 10. Operating Frequency, I<sub>VS</sub> < I<sub>VS-L</sub> Threshold (f<sub>OSC-H</sub>) vs. Temperature

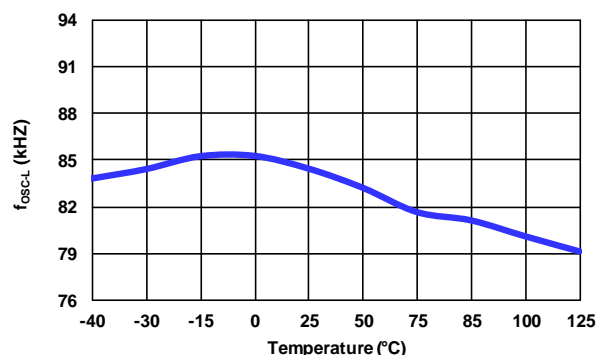


Figure 11. Operating Frequency while I<sub>VS</sub> < I<sub>VS-H</sub> Threshold (f<sub>OSC-L</sub>) vs. Temperature



## Typical Performance Characteristics (Continued)

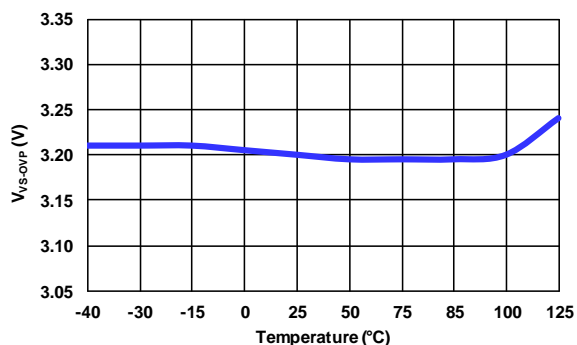


Figure 12. Output OVP with  $V_S$  Sampling Voltage ( $V_{VS-OVP}$ ) vs. Temperature

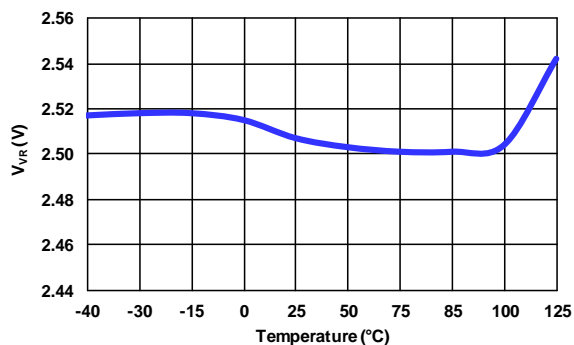


Figure 13. Internal Reference Voltage for CC Regulation ( $V_{VR}$ ) vs. Temperature

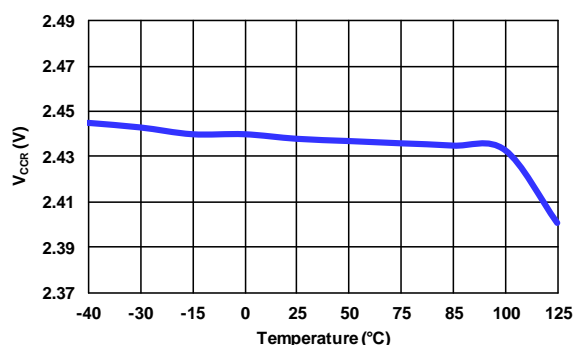


Figure 14. Variation Test Voltage on CS Pin for CC Regulation ( $V_{CCR}$ ) vs. Temperature

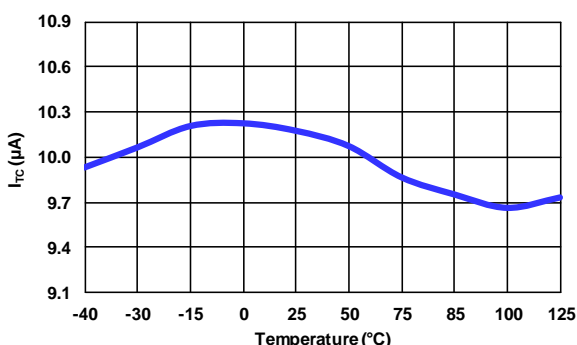


Figure 15. Temperature-Independent Bias Current ( $I_{TC}$ ) vs. Temperature

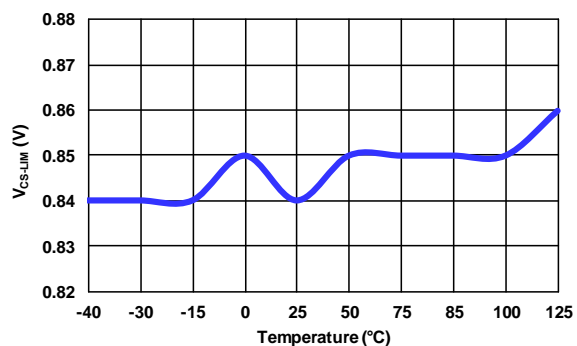


Figure 16. Current Limit Threshold Voltage ( $V_{CS-LIM}$ ) vs. Temperature

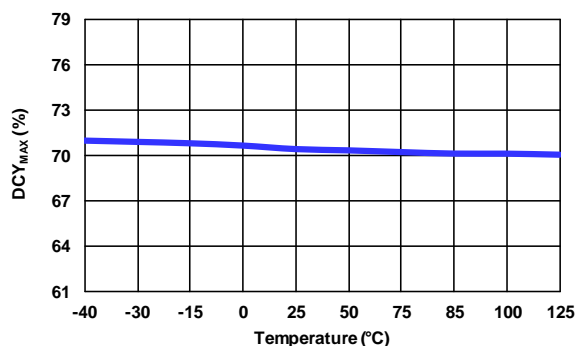


Figure 17. Maximum Duty Cycle ( $DCY_{MAX}$ ) vs. Temperature

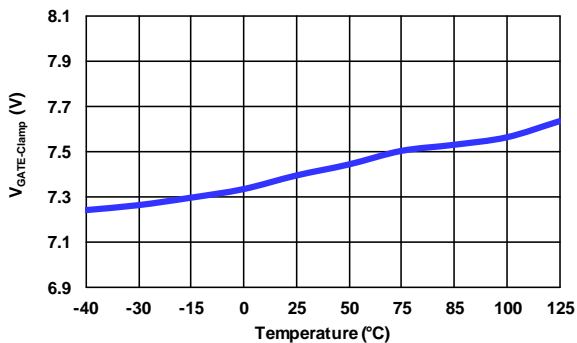


Figure 18. Gate Output Clamping Voltage ( $V_{GATE-Clamp}$ ) vs. Temperature

## Functional Description

FAN501A is an offline flyback converter controller that offers constant output voltage (CV) regulation through opto-coupler feedback circuitry and constant output current (CC) regulation with primary-side control. Advanced output current estimation technology allows stable CC regulation regardless of the power stage operation mode: Continuous Conduction Mode (CCM) or Discontinuous Conduction Mode (DCM).

Dual-switching-frequency operation adaptively selects the operational frequency between 85 kHz and 140 kHz according to the line voltage. As a result, the transformer can be fully utilized and high efficiency is maintained over entire line range. A frequency-hopping function is incorporated to reduce EMI noise.

Line voltage information through transformer auxiliary winding is used for dual-switching frequency selection and line voltage CC correction.

mWSaver® technology, including high-voltage startup and ultra-low operating current in Burst Mode, enables system compliance with Energy Star's 5-star requirement of <30 mW standby power consumption.

Protections such as  $V_{DD}$  Over-Voltage Protection ( $V_{DD}$  OVP),  $V_S$  Over-Voltage Protection ( $V_S$  OVP), internal Over-Temperature Protection (OTP), and brownout protection improve reliability.

All these innovative technologies allow the FAN501A to offer low total cost, reduced component counts, small size / weight, high conversion efficiency, and high power density for compact charger / adapter applications requiring CV / CC control.

## CV / CC PWM Operation Principle

Figure 19 shows a simplified CV / CC PWM control circuit of the FAN501A. The Constant Voltage (CV) regulation is implemented in the same manner as the conventional isolated power supply, where the output voltage is sensed using a voltage divider and compared with the internal reference of the shunt regulator to generate a compensation signal. The compensation signal is transferred to the primary side through an opto-coupler and scaled down by attenuator  $A_V$  to generate a COMV signal. This COMV signal is applied to the PWM comparator to determine the duty cycle.

The Constant Current (CC) regulation is implemented internally with primary-side control. The output current estimator calculates the output current using the transformer primary-side current and diode current discharge time. By comparing the estimated output current with internal reference signal, a COMI signal is generated to determine the duty cycle.

These two control signals, COMV and COMI, are compared with an internal sawtooth waveform ( $V_{SAW}$ ) by two PWM comparators to determine the duty cycle. Figure 20 illustrates the outputs of two comparators, combined with an OR gate, to determine the MOSFET turn-off instant. Of COMV and COMI, the lower signal determines the duty cycle. As shown in Figure 20, during CV regulation, COMV determines the duty cycle

while COMI is saturated to HIGH level. During CC regulation, COMI determines the duty cycle while COMV is saturated to HIGH level.

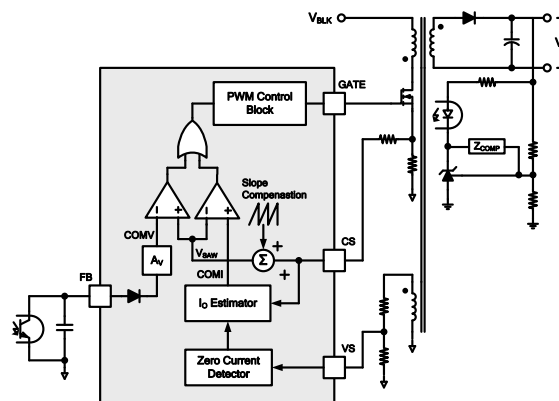


Figure 19. Simplified CV / CC PWM Control Circuit

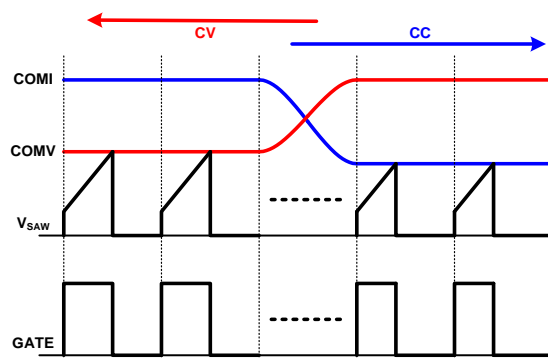


Figure 20. PWM Operation for CV / CC Regulation

## Primary-Side Constant Current Operation

Figure 21 and Figure 22 show the key waveforms of a flyback converter operating in DCM and CCM, respectively. The output current of each mode is estimated by calculating the average of output diode current over one switching cycle:

$$I_O = \langle I_D \rangle_{t_s} = \frac{\int_0^{t_s} I_D(t) dt}{t_s} = \frac{[I_D]_{AREA}}{t_s} \quad (1)$$

The area of output diode current in both DCM and CCM operation can be expressed in a same form, as a product of diode current discharge time ( $t_{DIS}$ ) and diode current at the middle of diode discharge ( $I_{D-Mid}$ ), such as:

$$[I_D]_{AREA} = I_{D-Mid} \cdot t_{DIS} \quad (2)$$

In steady state,  $I_{D-Mid}$  can be expressed as:

$$I_{D-Mid} = I_{DS-Mid} \cdot \frac{N_p}{N_s} \quad (3)$$

where  $I_{DS-Mid}$  is primary-side current at the middle of MOSFET conduction time and  $N_p/N_s$  is primary-to-secondary turn ratio.

The unified output current equation both for DCM and CCM operation is obtained as:

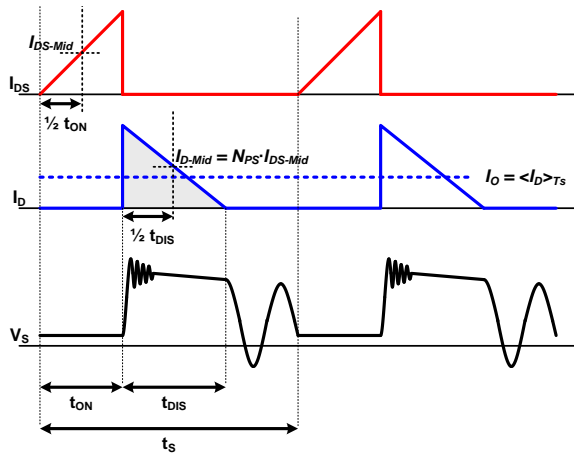
$$I_O = \frac{N_P}{N_S} \cdot I_{DS\_Mid} \cdot \frac{t_{DIS}}{t_S} = \frac{N_P}{N_S} \cdot \frac{V_{CS\_Mid}}{R_{CS}} \cdot \frac{t_{DIS}}{t_S} \quad (4)$$

$V_{CS\_Mid}$  is obtained by sampling the current-sense voltage at the middle of the MOSFET conduction time. The diode current discharge time is obtained by detecting the diode current zero-crossing instant. Since the diode current cannot be sensed directly in the primary side, Zero-Crossing Detection (ZCD) is accomplished indirectly by monitoring the auxiliary winding voltage in the primary side. When the diode current reaches zero, the transformer winding voltage begins to drop sharply. To detect the corner voltage, the  $V_S$  is sampled, called  $V_{SH}$ , at 85% of diode current discharge time ( $t_{DIS}$ ) of the previous switching cycle and compared with the instantaneous  $V_S$  voltage. When instantaneous voltage of the  $V_S$  pin drops below  $V_{SH}$  by more than 200 mV, the ZCD of diode current is obtained, as shown in Figure 23.

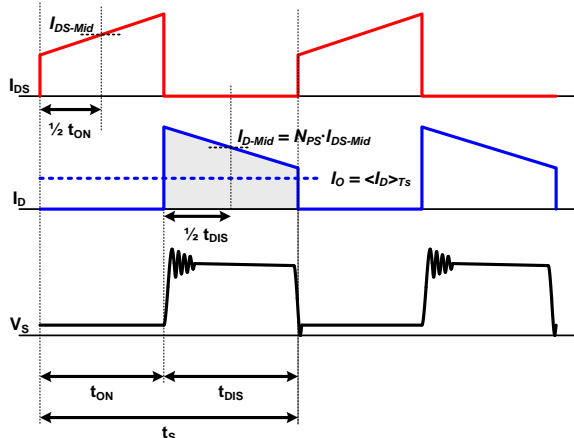
The output current can be programmable by setting current sensing resistor as:

$$R_{CS} = \frac{1}{I_O} \cdot \frac{N_P}{N_S} \cdot \frac{V_{CCR}}{K_{CC}} \quad (5)$$

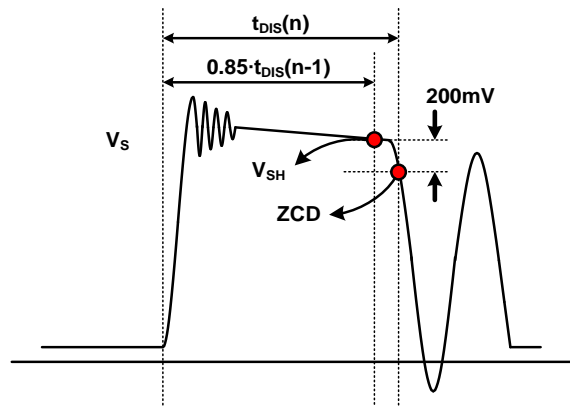
where  $V_{CCR}$  is the internal voltage for CC control and  $K_{CC}$  is the IC design parameter, 12 for the FAN501A.



**Figure 21. Waveforms of DCM Flyback Converter**



**Figure 22. Waveforms of CCM Flyback Converter**



**Figure 23. Operation Waveform for ZCD Function**

### Line Voltage Detection and its Utilization

The FAN501A indirectly senses line voltage using the current flowing out of the  $V_S$  pin while the MOSFET is turned on, as illustrated in Figure 25 and Figure 26. During the MOSFET turn-on period, auxiliary winding voltage,  $V_{Aux}$ , reflects input bulk capacitor voltage,  $V_{BLK}$ , by the transformer coupling between primary and auxiliary. During MOSFET conduction time, the line voltage detector clamps the  $V_S$  pin voltage  $\sim 0.5$  V and the current,  $I_{VS}$ , flowing from the  $V_S$  pin is expressed as:

$$I_{VS} = \frac{N_A / N_P \cdot V_{BLK}}{R_{VS1}} + \frac{0.5}{R_{VS1} // R_{VS2}} \quad (6)$$

Typically, the second term in Equation (6) can be ignored because it is much smaller than the first term. The current,  $I_{VS}$ , is approximately proportional to the line voltage, calculated as:

$$I_{VS} \cong \frac{N_A / N_P}{R_{VS1}} \cdot V_{BLK} \quad (7)$$

The  $I_{VS}$  current, reflecting the line voltage information, is used for dual switching frequency operation, CC control correction weighting, and brownout protection; as illustrated in Figure 25.

### Dual Switching Frequency

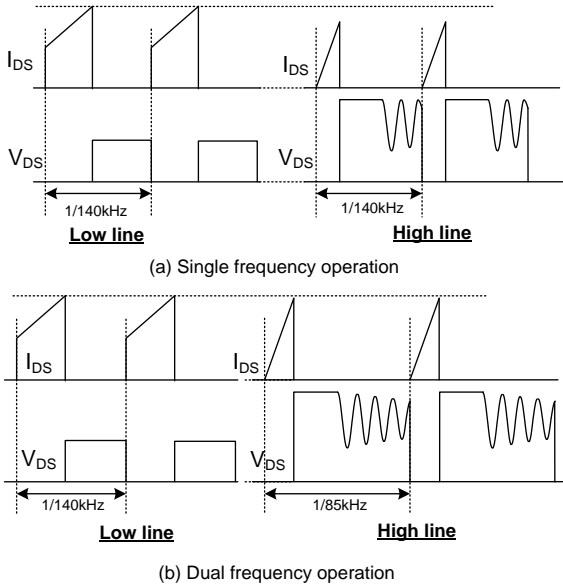
The FAN501A changes the switching frequency between 85 kHz and 140 kHz according to the line voltage. It is typical to design the flyback converter to operate in CCM for low line and DCM in high line. Therefore, the peak transformer current decreases as the operation mode changes from CCM to DCM, as shown in Figure 24(a), for single-frequency operation. The transformer is not fully utilized at high line when a single switching frequency is used. The peak transformer current can be maintained almost constant when the flyback converter operates at lower frequency at high line, as illustrated in Figure 24(b). This allows full transformer utilization and improves the efficiency by decreasing the switching losses at high line.

When  $I_{VS}$  is larger than  $I_{VS-H}$  (750  $\mu$ A), the switching frequency is set at  $f_{OSC-L}$  (85 kHz) in CV Mode. When  $I_{VS}$  is less than  $I_{VS-L}$  (680  $\mu$ A), the switching frequency is set at  $f_{OSC-H}$  (140 kHz) in CV Mode. For the universal line range, the frequency change should occur between 132  $\sim$  180  $V_{AC}$  to avoid the transition within the actual

operation range. It is typical to design the voltage divider for the VS pin such that frequency change occurs at  $170 V_{AC}$  ( $V_{DC} - 170 V_{AC} = 240 V$ ); calculated as:

$$R_{VS1} = \frac{N_A / N_P}{I_{VS-H}} \cdot 240 \quad (8)$$

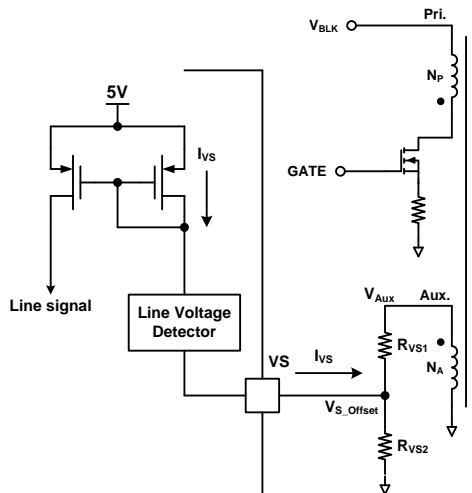
With the value of  $R_{VS1}$  determined from Equation (8), the switching frequency drops to 85 kHz as line voltage increases above  $170 V_{AC}$ , while switching frequency increases to 140 kHz, as line voltage drops  $<155 V_{AC}$ .



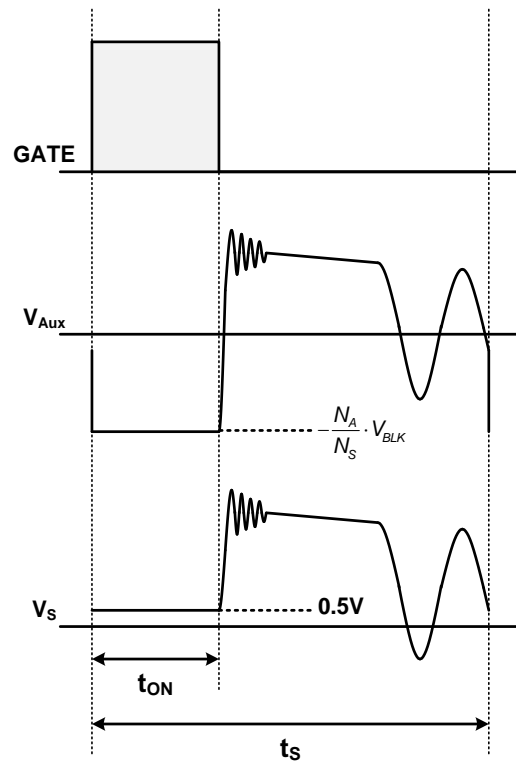
**Figure 24. Peak Switch Current, Single- and Dual-Frequency Operation**

### Brownout Protection

Line voltage information is also used for brownout protection. When the  $I_{VS}$  current out of the VS pin during the MOSFET conduction time is less than  $160 \mu A$  for longer than 30 ms, the brownout protection is triggered. When setting  $R_{VS1}$  as calculated in Equation (8), the brownout level is set at  $30 V_{AC}$ .



**Figure 25. Line Voltage Detection Circuit**



**Figure 26. Waveforms for Line Voltage Detection**

### Maximum Power Limit by Precision CC Control

Primary-side current-sensing voltage is used to estimate the output current for CC regulation. However, the actual output current regulation is also affected by the turn-off delay of the MOSFET, as illustrated in Figure 27. While FAN501A samples the CS pin voltage at the half on-time of gate drive signal, the actual turn-off is delayed by the MOSFET gate charge and driving current resulting in peak current detection error as:

$$\Delta I_{DS}^{PK} = \frac{V_{DL}}{L_m} t_{OFF.DLY} \quad (9)$$

where  $L_m$  is the primary side magnetic inductance.

As can be seen, the error is proportional to the line voltage. FAN501A has an internal correction function to improve CC regulation, as shown in Figure 28. Line information is obtained through the line voltage detector as shown in Figure 25 and Figure 26 and this information is used for the CC regulation correction. The correction gain can be programmed using external resistor  $R_{COMP}$  on the COMP pin. This correction current,  $I_{LVF}$ , flows through internal resistor,  $R_{LVF}$ , and external resistor,  $R_{CSF}$ , to introduce offset voltage on current sensing voltage. Thus, the primary current detection error affected by line voltage and turn-off delay is corrected for better CC regulation. The  $R_{COMP}$  resistor can be calculated as:

$$R_{COMP} = \frac{N_P}{N_A} \cdot \frac{R_{CS}}{R_{LVF} + R_{CSF}} \cdot R_{VS1} \cdot \frac{t_{OFF.DLY}}{L_m} \cdot K_{COMP} \quad (10)$$

where  $R_{LVF}$  is the internal resistor on the IC, which is  $2.0 k\Omega$ , and  $K_{COMP}$  is the design factor of the IC, which is  $3.745 M\Omega$ .

The turn-off delay should be obtained by measuring the time between the falling edge and actual turn-off instant of MOSFET, as illustrated in Figure 27.

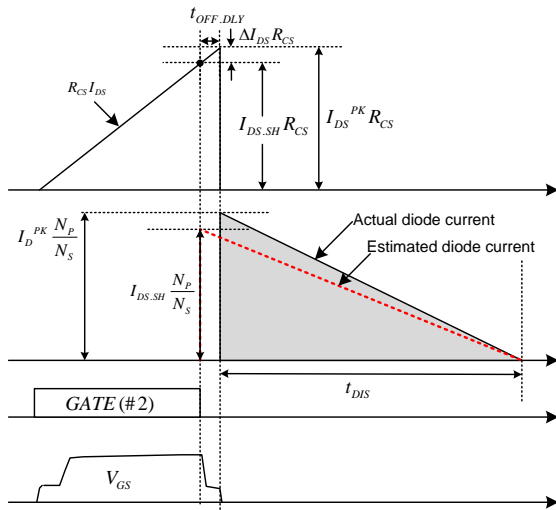


Figure 27. CC Control Correction Concept

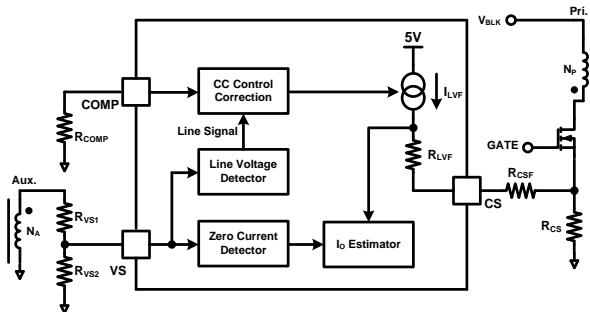


Figure 28. CC Correction Circuit

### Pulse-by-Pulse Current Limit

Since the peak transformer current is controlled by a feedback loop, the peak transformer current is not properly controlled when the feedback loop is saturated to HIGH, which typically occurs under startup or overload conditions. To limit the current, a pulse-by-pulse current limit forces the gate drive signal to turn off when the CS pin voltage reaches the current-limit threshold ( $V_{CS-LIM}$ ) in normal operation.

### Burst Mode Operation

The power supply enters Burst Mode at no-load or extremely light-load condition. As shown in Figure 29, when  $V_{FB}$  drops below  $V_{FB-Burst-L}$ , the PWM output shuts off and the output voltage drops at a rate dependent on load current. This causes the feedback voltage to rise. Once  $V_{FB}$  exceeds  $V_{FB-Burst-H}$ , the internal circuit starts to provide a switching pulse. The feedback voltage then falls and the process repeats. In this manner, Burst Mode alternately enables and disables switching of the MOSFET to reduce the switching losses in Standby Mode. In Burst Mode, the operating current is reduced from 3.5 mA to 250  $\mu$ A to minimize power consumption.

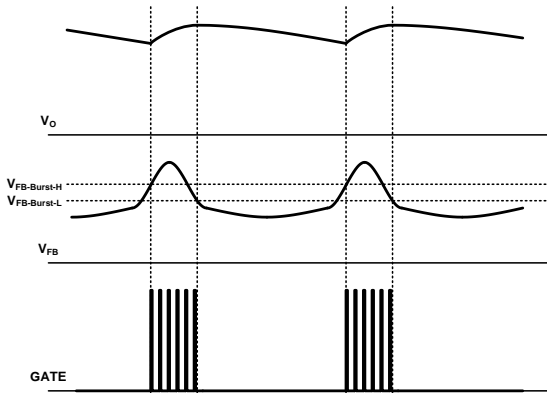


Figure 29. Burst-Mode Operation

### Frequency Hopping

EMI reduction is accomplished by frequency hopping, which spreads the energy over a wider frequency range than the bandwidth of the EMI test equipment, allowing compliance with EMI limitations.

### Slope Compensation

The sensed voltage across the current-sense resistor is used for current-mode control and pulse-by-pulse current limiting. A synchronized ramp signal with a positive slope is added to the current-sense information at each switching cycle, improving noise immunity during current mode control and avoiding sub-harmonic oscillation during CCM operation.

### Leading-Edge Blanking (LEB)

Each time the power MOSFET is switched on, a turn-on spike occurs at the sense resistor. To avoid premature termination of the switching pulse by the spike, a 150 ns leading-edge blanking time is incorporated. Conventional RC filtering can therefore be omitted. During this blanking period, the current-limit comparator is disabled and it cannot switch off the gate driver.

### Noise Immunity

Noise from the current sense or the control signal can cause significant pulse-width jitter. Though slope compensation helps alleviate this problem, precautions should be taken to improve the noise immunity. Good placement and layout practices are important. Avoid long PCB traces and component leads and locate bypass capacitor as close to the PWM IC as possible.

### High Voltage (HV) Startup

Figure 30 shows the high-voltage (HV) startup circuit for FAN501A applications. The JFET is used to internally implement the high-voltage current source (see Figure 31 for characteristics). Technically, the HV pin can be directly connected to voltage ( $V_{BLK}$ ) on an input bulk capacitor. To improve reliability and surge immunity, however, it is typical to use a  $\sim 100$  k $\Omega$  resistor between the HV pin and bulk capacitor voltage. The actual HV current with a given bulk capacitor voltage and startup resistor is determined by the intersection of V-I characteristics line and load line, as shown in Figure 31.

During startup, the internal startup circuit is enabled and the bulk capacitor voltage supplies the current,  $I_{HV}$ , to charge the hold-up capacitor,  $C_{VDD}$ , through  $R_{HV}$ . When  $V_{DD}$  reaches  $V_{DD-ON}$ , the internal HV startup circuit is disabled and the IC starts PWM switching. Once the HV startup circuit is disabled, the energy stored in  $C_{VDD}$  should supply the IC operating current until the transformer auxiliary winding voltage reaches the nominal value. Therefore,  $C_{VDD}$  should be designed to prevent  $V_{DD}$  from dropping to  $V_{DD-OFF}$  before the auxiliary winding builds up enough voltage to supply  $V_{DD}$ .

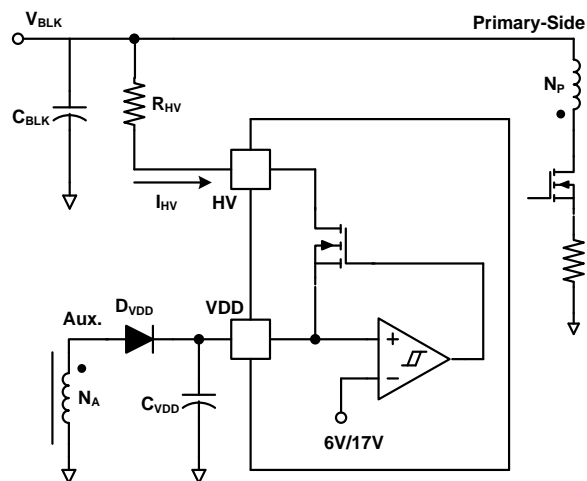


Figure 30. HV Startup Circuit

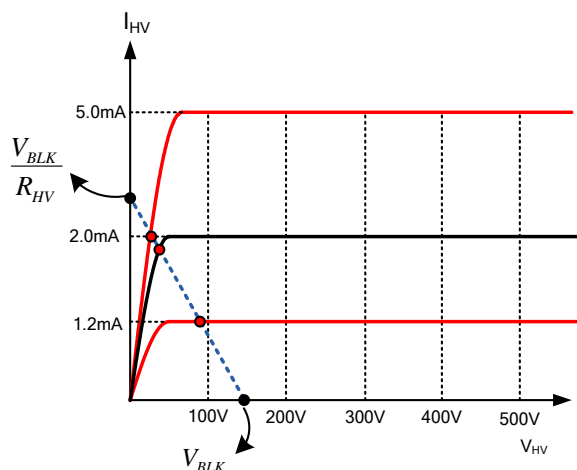


Figure 31. V-I Characteristic of HV Pin

## Protections

The protection functions include  $V_{DD}$  over-voltage protection ( $V_{DD}$  OVP), brownout protection,  $V_S$  over-voltage protection ( $V_S$  OVP), internal over-temperature protection (OTP), and externally triggered shutdown (SD) protection. The  $V_{DD}$  OVP and brownout protection are implemented as Auto-Restart Mode.  $V_S$  OVP, OTP, and SD protections are implemented as Latch Mode.

When an Auto-Restart Mode protection is triggered, switching is terminated and the MOSFET remains off, causing  $V_{DD}$  to drop. When  $V_{DD}$  drops to the  $V_{DD}$  turn-off voltage of 5.8 V, the protection is reset, and next step to reduced operation current until startup circuit is enabled.

The supply current drawn from the HV pin charges the hold-up capacitor. When  $V_{DD}$  reaches the turn-on voltage of 17.5 V, normal operation resumes. In this manner, Auto-Restart Mode alternately enables and disables MOSFET switching until the abnormal condition is eliminated, as shown in Figure 32.

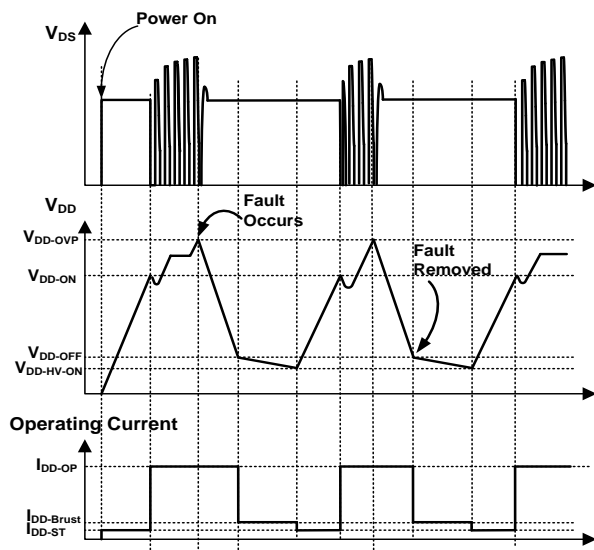


Figure 32. Auto-Restart Mode Operation

When a Latch Mode protection is triggered, PWM switching is terminated and the MOSFET remains off, causing  $V_{DD}$  to drop. When  $V_{DD}$  drops to the  $V_{DD}$  turn-off voltage of 5.8 V, the internal startup circuit is enabled without resetting the protection and the supply current drawn from HV pin charges the hold-up capacitor. Since the protection is not reset, the IC does not resume PWM switching even when  $V_{DD}$  reaches the turn-on voltage of 17.5 V, disabling HV startup circuit. Then  $V_{DD}$  drops again down to 5.8 V. In this manner, Latch Mode protection alternately charges and discharges  $V_{DD}$  until there is no more energy in DC link capacitor. The protection is reset when  $V_{DD}$  drops to 2.5 V, which is allowed only after power supply is unplugged from the AC line, as shown in Figure 33.

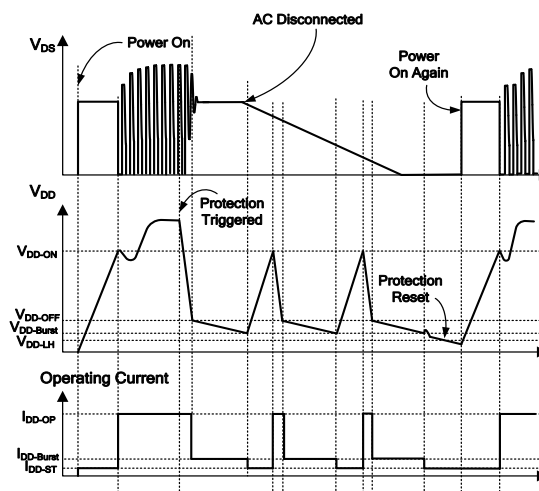


Figure 33. Latch Mode Operation



### V<sub>DD</sub> Over-Voltage-Protection

V<sub>DD</sub> over-voltage protection prevents damage from over-voltage exceeding the IC voltage rating. When V<sub>DD</sub> exceeds 28 V due to an abnormal condition, protection is triggered. This protection is typically caused by an open circuit in the secondary-side feedback network.

### Brownout Protection

Brownout protection is implemented through line voltage detection circuit using the auxiliary winding, as shown in Figure 25 and Figure 26. When the current flowing out of the VS pin during the MOSFET conduction time is smaller than 160 μA for longer than 30 ms, the brownout protection is triggered.

### Over-Temperature Protection (OTP)

If the junction temperature exceeds 140°C (T<sub>OTP</sub>), the internal temperature-sensing circuit shuts down PWM output and enters Latch Mode protection.

### Fold-Back Point and Over-Voltage Protection (V<sub>S</sub> OVP)

Generally, the fold-back point in CC regulation as output drops is determined by the V<sub>DD-OFF</sub> level. Thus, the fold-back level mainly depends on the characteristics of the V<sub>DD</sub> diode and transformer. For V<sub>S</sub> pin voltage divider design, R<sub>VS1</sub> is obtained from Equation (8), and R<sub>VS2</sub> is determined by the V<sub>SOVP</sub> function as:

$$R_{VS2} = R_{VS1} \cdot \left( \frac{V_{O-OVP}}{V_{VS-OVP}} \cdot \frac{N_A}{N_S} - 1 \right)^{-1} \quad (11)$$

where V<sub>O-OVP</sub> is the output over-voltage protection threshold level.

V<sub>S</sub> over-voltage protection prevents damage caused by output over-voltage condition. Figure 34 shows the internal circuit of V<sub>S</sub> OVP. When abnormal system conditions occur that cause V<sub>S</sub> sampling voltage to exceed V<sub>VS-OVP</sub> (3.2 V) for more than debounce switching cycles (N<sub>VS-OVP</sub>), PWM pulses are disabled and the FAN501A enters Latch Mode protection. V<sub>S</sub> over-voltage conditions are usually caused by an open circuit in the secondary-side feedback network or a fault condition in the V<sub>S</sub> pin voltage divider resistors.

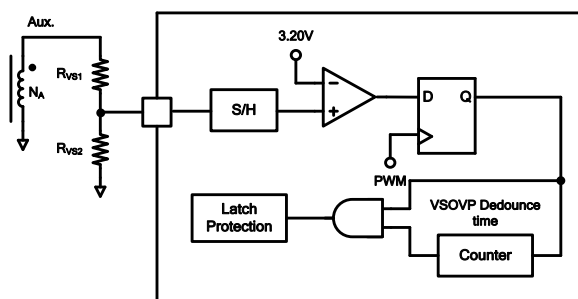


Figure 34. V<sub>S</sub> OVP Protection Circuit

### Externally Triggered Shutdown

By pulling the SD pin voltage below threshold voltage, V<sub>SD-TH</sub> (1.0 V); shutdown can be externally triggered and the FAN501A enters Latch Mode protection. It can be also used for external OTP protection by connecting an NTC thermistor between the shutdown (SD) programming pin and ground. An internal constant current source, I<sub>SD</sub> (100 μA), introduces voltage drop across the thermistor. Resistance of the NTC thermistor becomes smaller as the ambient temperature increases, which reduces the voltage drops across the thermistor. When the voltage of the SD pin is less than threshold voltage V<sub>SD-TH</sub> (1.0 V), OTP protection is triggered.

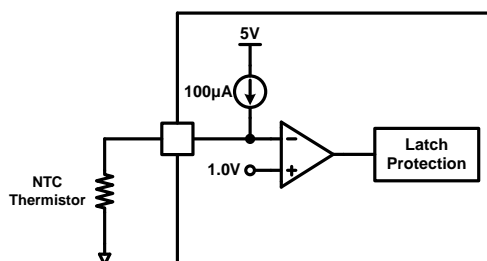
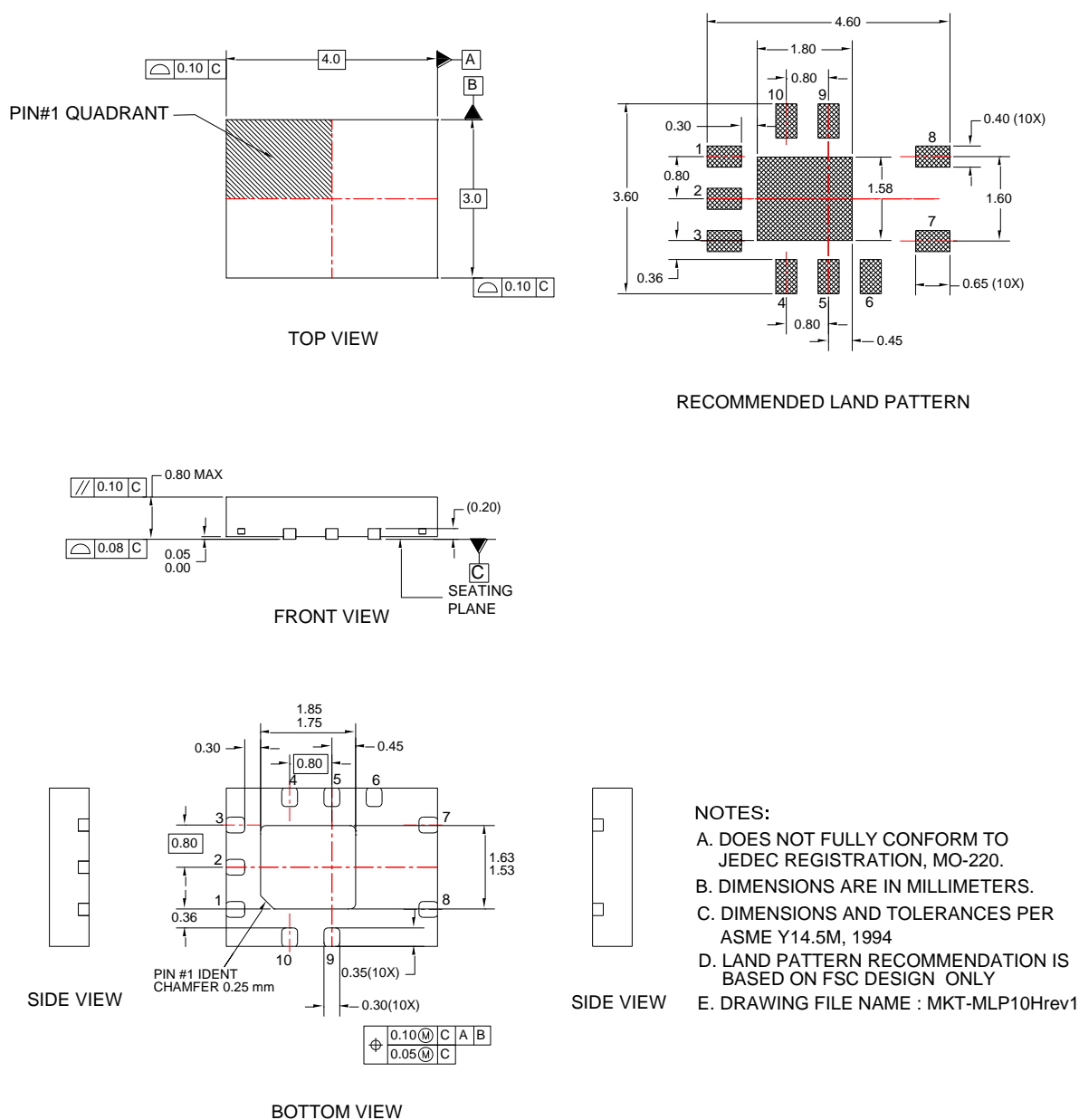


Figure 35. Thermal Shutdown Using SD Pin

## Physical Dimensions



**Figure 36. 10-Lead, MLP, QUAD, JEDEC MO-220 4 mm X 3 mm, 0.8 mm Pitch, Single DAP**

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