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NCP3121

Dual 3.0 A, Step-Down DC/DC Switching Regulator

The NCP3121 is a dual buck converter designed for low voltage applications requiring high efficiency. This device is capable of producing an output voltage as low as 0.8 V. The NCP3121 provides dual 3.0 A switching regulators with an adjustable 200 kHz − 750 kHz switching frequency. The switching frequency is set by an external resistor. The NCP3121 also incorporates an auto−tracking and sequencing feature. Protection features include cycle−by−cycle current limit and undervoltage lockout (UVLO). The NCP3121 comes in a 32−pin QFN package.

Features

- Input Voltage Range from 4.5 V to 13.2 V
- 12 V_{in} to 5.0 V_{out} = 85% Efficiency Min @ 3.0 A
- 200−750 kHz Operation
- Stable with Low ESR Ceramic Output Capacitor
- 0.8 ±1.5% FB Reference Voltage
- External Soft−Start
- Out of Phase Operation of OUT1 & OUT2
- Auto−Tracking and Sequencing
- Enable/Disable Capability
- Hiccup Overload Protection
- Low Shutdown Power (I_q < 100 μA)

Typical Applications

- Set−Top Boxes, Portable Applications, Networking and Telecommunications
- DSP/μP/FPGA Core

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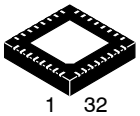

QFN32
CASE 488AM

MARKING DIAGRAM

NCP3121 = Specific Device Code
A = Assembly Location
WL = Wafer Lot
YY = Year
WW = Work Week
▪ = Pb−Free Package
(Note: Microdot may be in either location)

ORDERING INFORMATION

See detailed ordering and shipping information in the package dimensions section on page 40 of this data sheet.

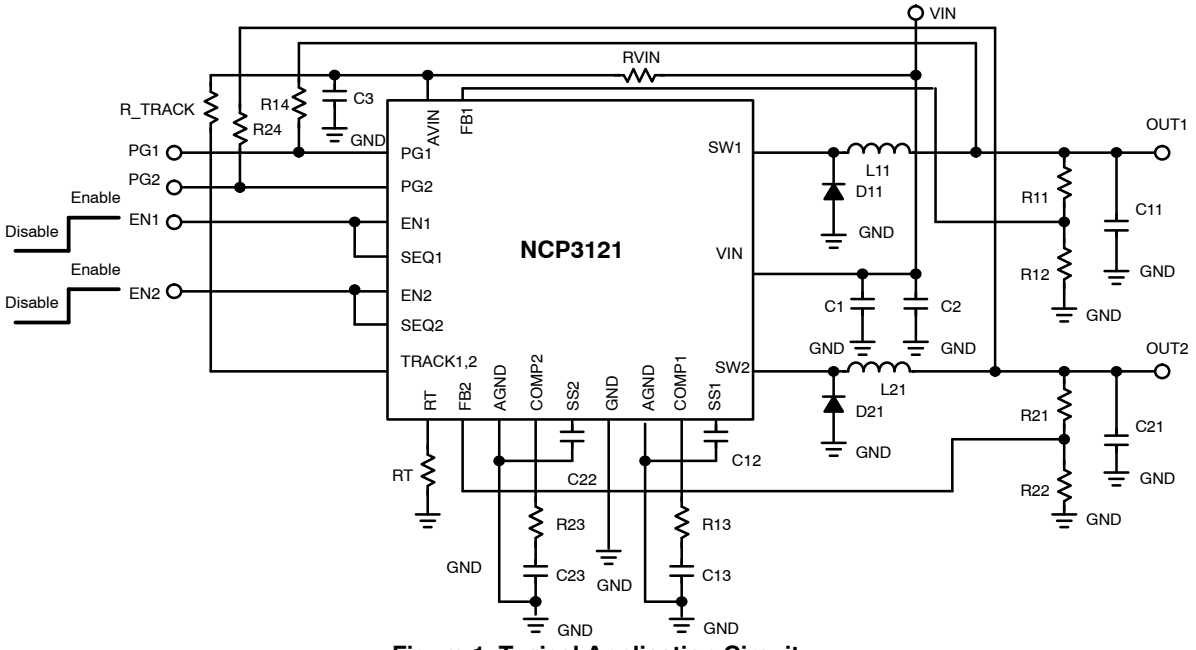

Figure 1. Typical Application Circuit

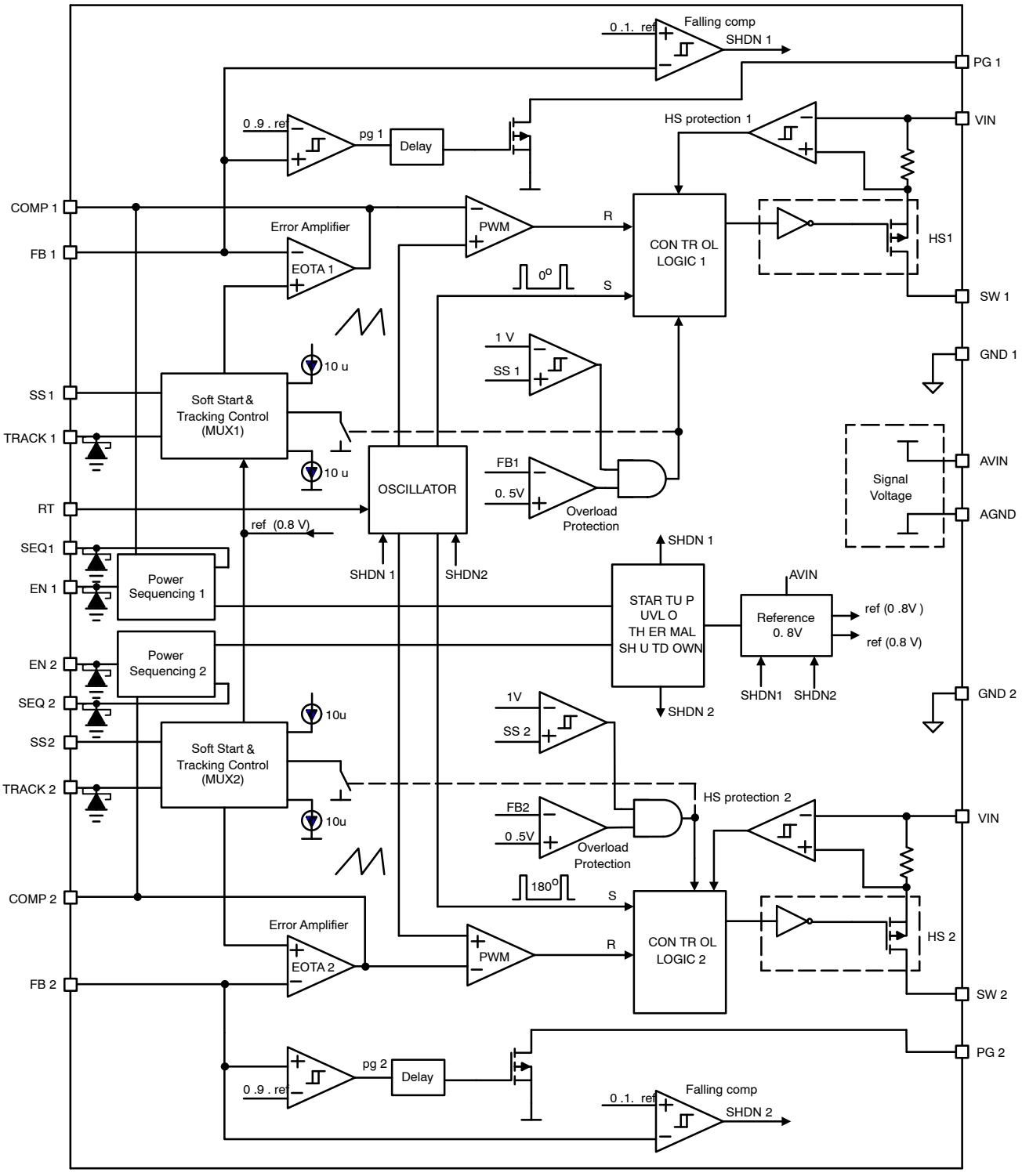

Figure 2. Block Diagram

PIN DESCRIPTION

Pin	Symbol	Description
1, 31, 32	SW1	Switch node of Channel 1. Connect an inductor between SW1 and the regulator output.
2 – 7	V_{IN}	Input power supply voltage pins. These pins should be connected together to the input signal supply voltage pin.
8 – 10	SW2	Switch node of Channel 2. Connect an inductor between SW2 and the regulator output.
11	GND2	Power ground for Channel 2
12	SS2	Soft−start control input for Channel 2. An internal current source charges an external capacitor connected to this pin to set the soft−start time.
13	COMP2	Compensation pin of Channel 2. This is the output of the error amplifier and inverting input of the PWM comparator.
14	AGND	Analog ground; connect to GND1 and GND2.
15	FB2	Feedback Pin. Used to set the output voltage of Channel 2 with a resistive divider from the output.
16	RT	Resistor select for the oscillator frequency. Connect a resistor from the RT pin to AGND to set the frequency of the master oscillator. Leave this pin floating, for 200 kHz operation.
17	TRACK 2	Tracking input for Channel 2. This pin allows the user to control the rise time of the second output. This pin must be tied high in the normal operation (except in the tracking mode).
18	TRACK 1	Tracking input for Channel 1. This pin allows the user to control the rise time of the first output. This pin must be tied high in the normal operation (except in the tracking mode).
19	SEQ2	Sequence pin for Channel 2. I/O used in power sequencing. Connect SEQ to EN for normal operation of a standalone device.
20	EN2	Enable input for Channel 2.
21	SEQ1	Sequence pin for Channel 1. I/O used in power sequencing. Connect SEQ to EN for normal operation of a standalone device.
22	EN1	Enable input for Channel 1.
23	PG2	Power good, open−drain output of Channel 2. Output logic is pulled to ground when the output is less than 90% of the desired output voltage. Tied to an external pull−up resistor. Leave this pin floating, if not used.
24	PG1	Power good, open−drain output of Channel 1. Output logic is pulled to ground when the output is less than 90% of the desired output voltage. Tied to an external pull−up resistor. Leave this pin floating, if not used.
25	AV_{IN}	Input signal supply voltage pin.
26	FB1	Feedback Pin. Used to set the output voltage of Channel 1 with a resistive divider from the output.
27	AGND	Analog ground. Connect to GND1 and GND2.
28	COMP1	Compensation pin of Channel 1. This is the output of the error amplifier and inverting input of the PWM comparator.
29	SS1	Soft−start/stop control input for Channel 1. An internal current source charges an external capacitor connected to this pin to set the soft−start time.
30	GND1	Power ground for Channel 1.
	Exposed Pad (GND)	The exposed pad at the bottom of the package is the electrical ground connection of the NCP3121. This node must be tied to ground.

MAXIMUM RATINGS

Characteristics	Symbol	Min	Max	Unit
Power Supply Voltage Input	V_{VIN}	−0.3	15	V
Signal Supply Voltage Input	V_{AVIN}	−0.3	15	V
SW Pin Voltage	V_{SW}	−0.7 −5V for < 50 ns	V_{VIN}	V
EN Pin Voltage Input	V_{EN}	−0.3	8.0	V
SEQ Pin Voltage Output	V_{SEQ}	−0.3	8.0	
PG Pin Voltage	V_{PG}	−0.3	5.5	V
All Other Pins	–	−0.3	5.5	°V
Thermal Resistance, Junction−to−Ambient (Note 1)	$R_{\theta JA}$	50		°C/W
Storage Temperature Range	T_{STG}	−55 to +150		°C
Junction Operating Temperature (Note 2)	T_J	−40 to +150		°C

Stresses exceeding Maximum Ratings may damage the device. Maximum Ratings are stress ratings only. Functional operation above the Recommended Operating Conditions is not implied. Extended exposure to stresses above the Recommended Operating Conditions may affect device reliability.

1. $R_{\theta JA}$ on a 100 x 100 mm PCB with two solid 1 oz ground planes.
2. The maximum package power dissipation limit must not be exceeded

$$P_D = \frac{T_{J\ (max)} - T_A}{R_{\theta JA}}$$

ELECTRICAL CHARACTERISTICS (−40°C < T_J < 125°C, T_J = 25°C for typical values, V_{AVIN} =12 V, V_{VIN} =12 V, unless otherwise noted. R_T = open kΩ)

Characteristic	Conditions	Min	Typ	Max	Unit
RECOMMENDED OPERATING CONDITIONS					
Input Voltage Range		4.5		13.2	V
SUPPLY CURRENT					
Quiescent Supply Current	V_{EN} = H, V_{FB} = 1.0 V No Switching, PG open		5.0	7.0	mA
Shutdown Supply Current	V_{EN} = 0 V, PG open			100	μA
UNDERVOLTAGE LOCKOUT					
UVLO Threshold	V_{IN} Rising Edge V_{IN} Falling Edge	 3.9	4.3 4.1	4.5	V
UVLO Hysteresis		0.15	0.20	0.25	V
SWITCHING REGULATOR					
Minimum Duty Cycle	Comp = 0.6 V			0	%
Maximum Duty Cycle	Comp = 2.6 V	90			%
High Side MOSFET $R_{DS(on)}$	I_{SW} = 0.5 A, T_J = 25°C		250		mΩ
High Side Leakage Current	V_{EN} = 0V, V_{SW} = 0V			10	μA
High Side Switch Current Limit Set Point	(Note 3)	3.5	4.15	4.8	A
Current Loop Transient Response	(Note 4)		100		nsec
FB					
V_{FB} Feedback Voltage	T_J = 25°C T_J = −40 to 125°C, 4.5 V < V_{IN} < 13.2V	0.788 0.784	0.8 –	0.812 0.816	V
OSC					
Oscillator Frequency	T_J = 25°C, T_J = −40 to 125°C T_J = 25°C, T_J = −40 to 125°C (RT = 52.3 kΩ)	180 170 635	200 200 750	220 230 865	kHz kHz kHz
Standard Oscillator Frequency Range	T_J = 25°C	200		750	kHz
TRANSCONDUCTANCE ERROR AMPLIFIER (GM)					
Transconductance	(Note 4)	0.9	1.0	1.1	mS
DC Gain	(Note 4)	50	55	60	dB
Unity Gain Bandwidth	(Note 4)		4.0		MHz
Output Sink Current	V_{FB} = 1.0 V, Vcomp = 1.5 V	80	100		μA
Output Source Current	V_{FB} = 0.6 V, Vcomp = 1.5 V	80	100		μA
Input Bias Current	V_{FB} = 0.8 V		100	500	nA
Comp Pin Operating Voltage Range	(Note 4)	0.6		2.6	V
SOFT−START					
Soft−Start Period	V_{FB} < 0.8 V, C_S = 0.1 μF		10		ms
Soft−Start Voltage Range		0		V_{FB}	V
Soft−Start Current Source	Charging, V_{SS} = 1 V Discharging, V_{SS} = 1 V	6.0 6.0	8.0 8.0	12 12	μA μA

ELECTRICAL CHARACTERISTICS (−40°C < T_J < 125°C, T_J = 25°C for typical values, V_{AVIN} =12 V, V_{VIN} =12 V, unless otherwise noted. R_T = open kΩ)

Characteristic	Conditions	Min	Typ	Max	Unit
TRACK					
Tracking Voltage Range		0		V_{FB}	V
Tracking Voltage Offset	V_{TRACK} = 0.6 V			15	mV
Track Bias Current	V_{TRACK} = 0.6 V		100	500	nA
POWER GOOD					
PG Threshold	Feedback Voltage Rising, EN Tied to SEQ, V_{PG} = 3.3 V	90% V_{FB}			V
PG Shutdown Mode	Feedback Voltage Falling, EN Tied to SEQ, $V_{EN,SEQ}$ = 0V, V_{PG} = 3.3V	10% V_{FB}	15% V_{FB}	20% V_{FB}	V
PG Delay	Rising Edge of V_{out} Falling Edge of V_{out}		50 10		μs μs
PG Low Level Voltage	$I_{(PG)}$ = 1 mA			0.3	V
PG Hysteresis			45		mV
PG Leakage Current	V_{PG} = 5.5 V			1.0	μA
ENABLE/POWER SEQUENCING					
Enable Internal Pullup Current			4.0		μA
Sequence Internal Pulldown Current			16		μA
Enable Threshold High	EN Tied to SEQ	2.0			V
Sequence Threshold Low	EN Tied to SEQ			0.8	V
THERMAL SHUTDOWN					
Overtemperature Trip Point	(Note 4)		160		°C
Hysteresis			15		°C

3. DC value.
4. Guaranteed by design.

TYPICAL OPERATING CHARACTERISTICS

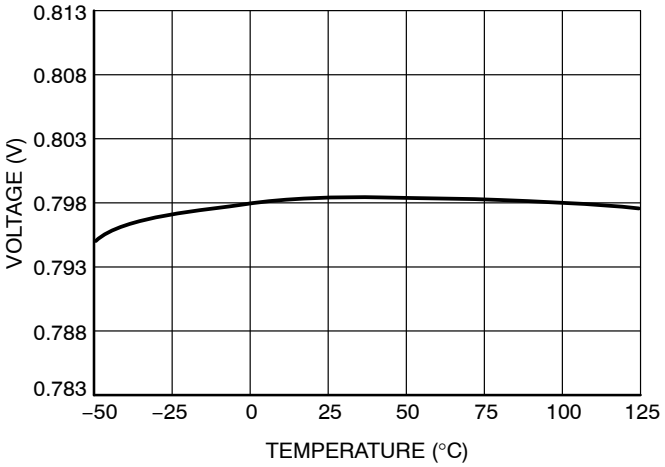

Figure 3. Feedback Voltage vs. Temperature

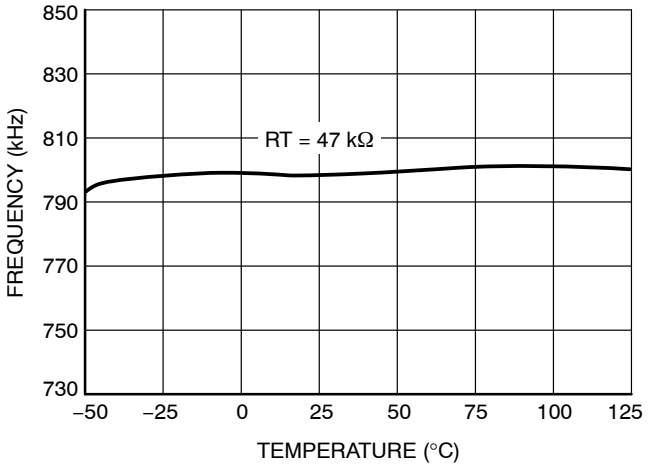

Figure 4. High Switching Frequency vs. Temperature

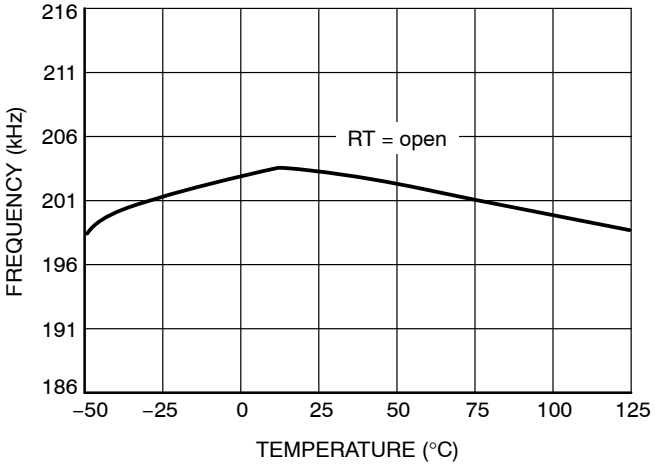

Figure 5. Low Switching Frequency vs. Temperature

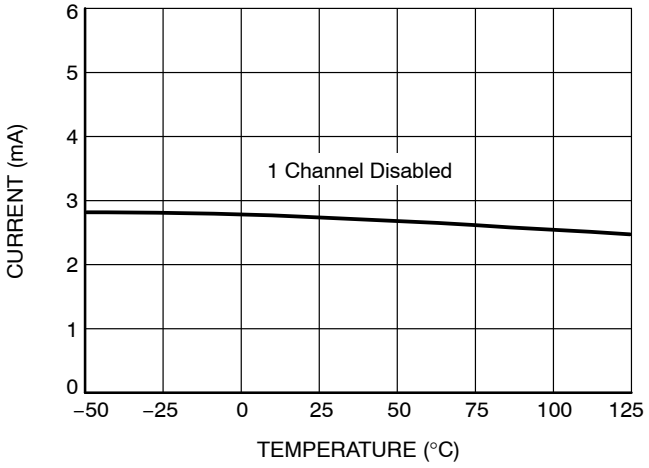

Figure 6. Quiescent Supply Current vs. Temperature

TYPICAL OPERATING CHARACTERISTICS

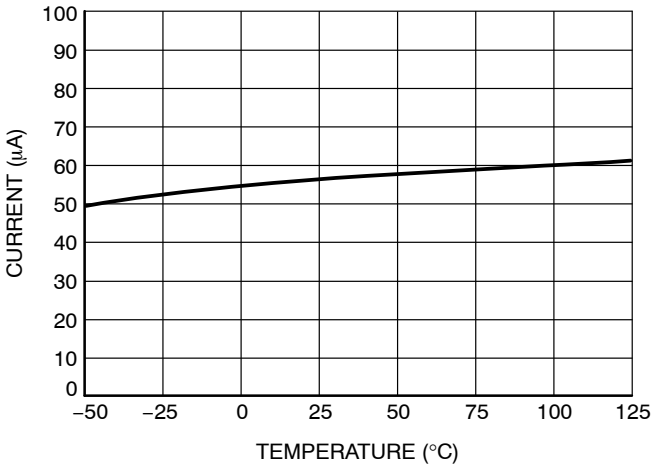

Figure 7. Shutdown Supply Current vs. Temperature

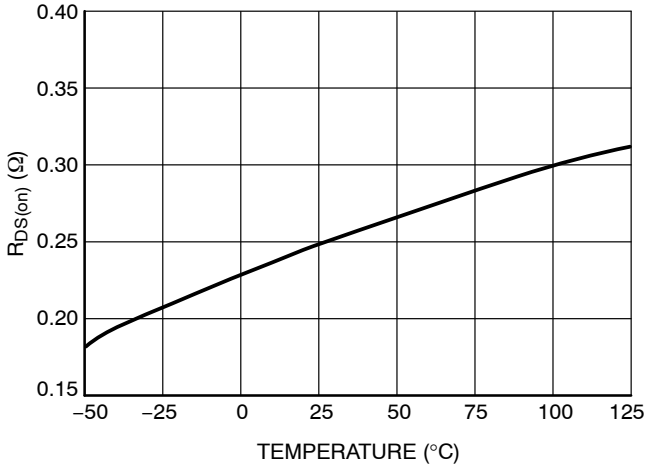

Figure 8. $R_{DS(on)}$ vs. Temperature

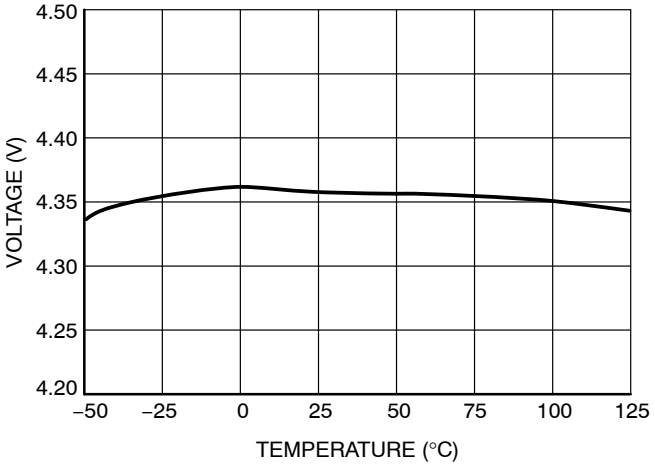

Figure 9. UVLO − Rising Threshold vs. Temperature

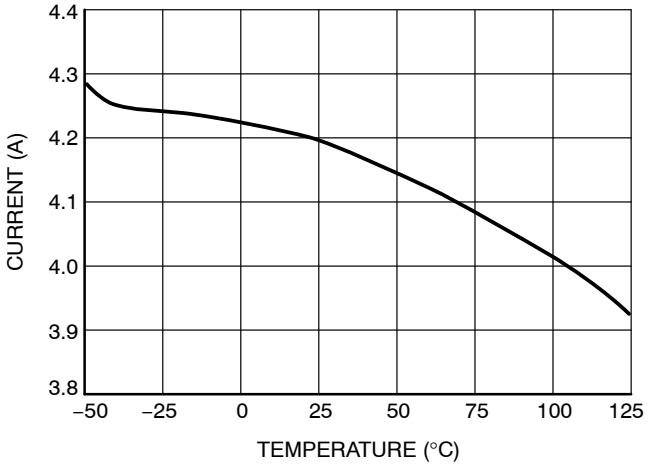

Figure 10. Current Limit vs. Temperature

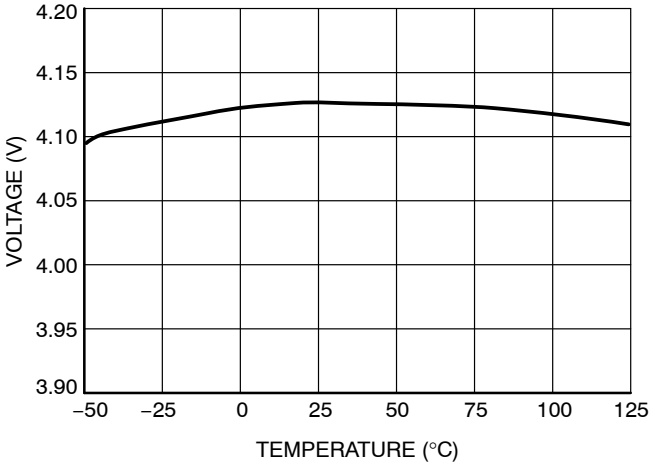

Figure 11. UVLO − Falling Threshold vs. Temperature

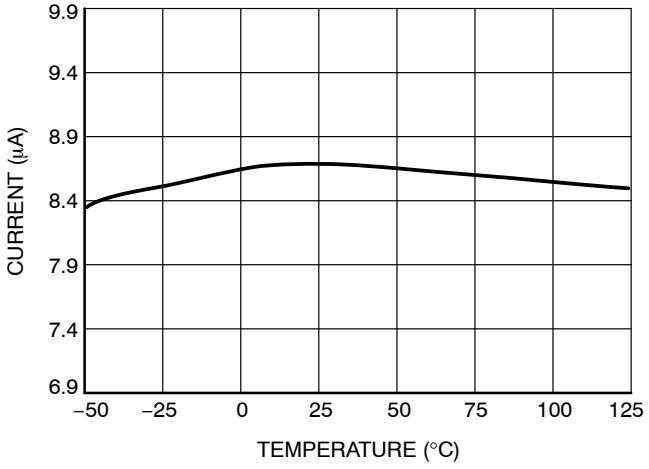

Figure 12. Soft−Start Charge Current vs. Temperature

TYPICAL OPERATING CHARACTERISTICS

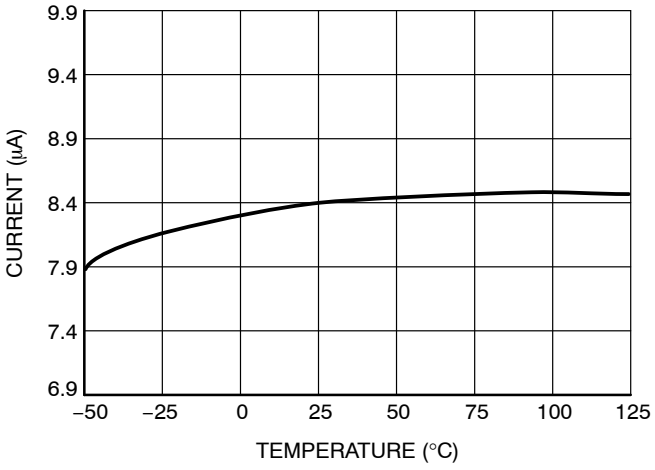

Figure 13. Soft−Start Discharge Current vs. Temperature

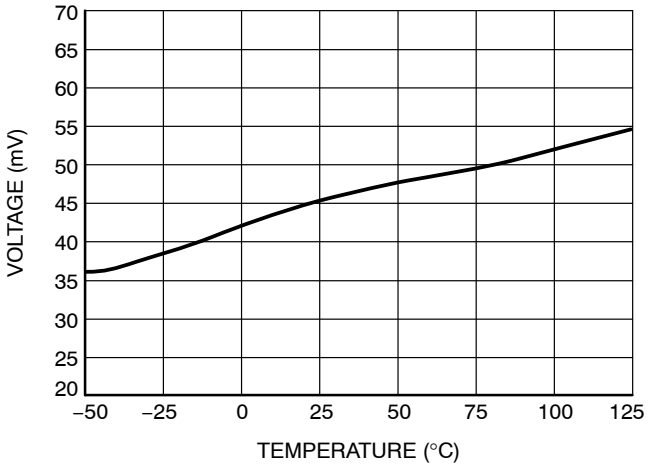

Figure 14. Power Good Hysteresis vs. Temperature

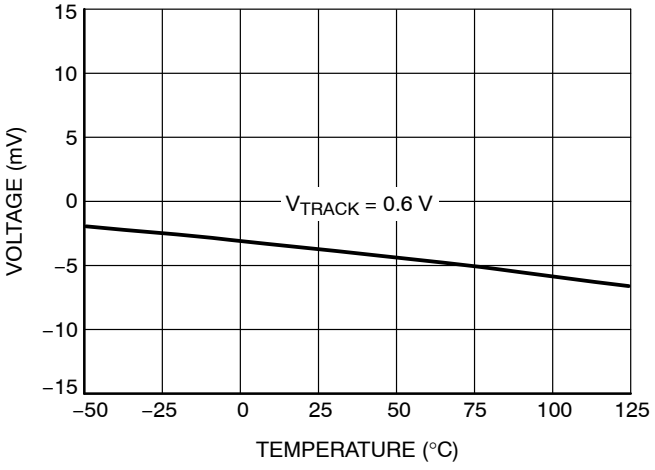

Figure 15. Tracking Voltage Offset vs. Temperature

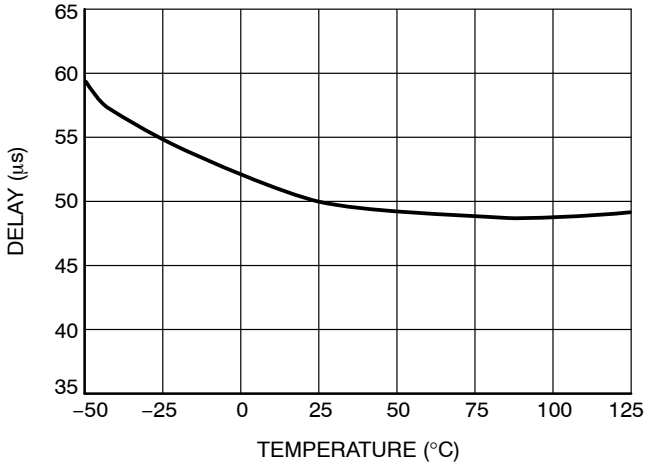

Figure 16. Power Good Rising Delay vs. Temperature

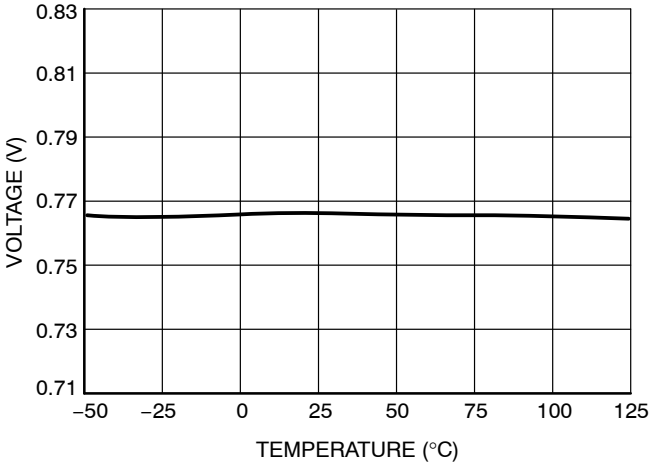

Figure 17. Power Good Feedback Threshold vs. Temperature

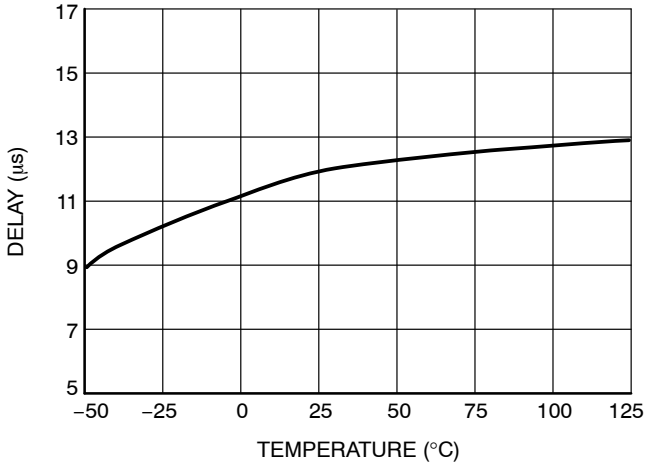

Figure 18. Power Good Falling Delay vs. Temperature

TYPICAL OPERATING CHARACTERISTICS

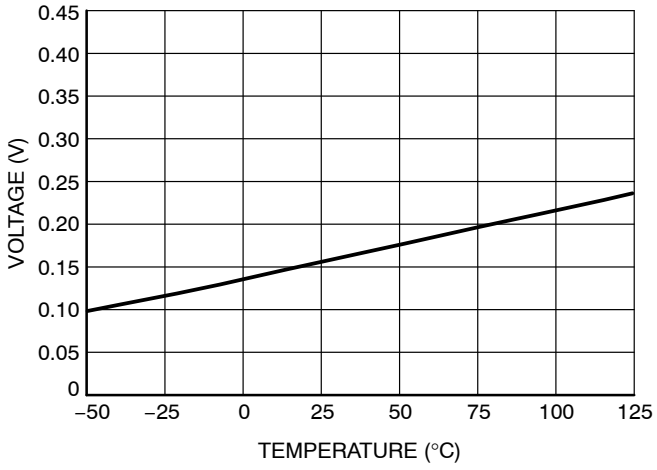

Figure 19. Power Good Saturation Voltage vs. Temperature

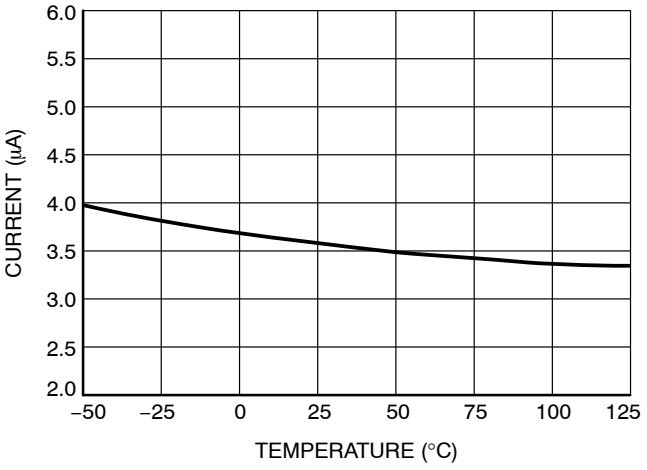

Figure 20. EN Internal Pull−up Current vs. Temperature

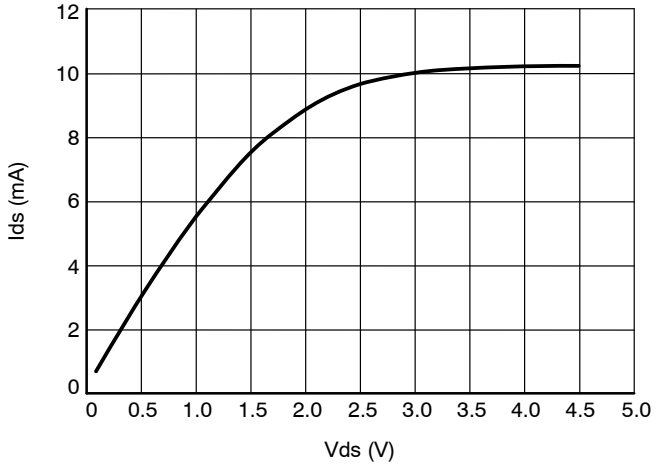

Figure 21. Power Good Current vs. Drain−to−Source Voltage

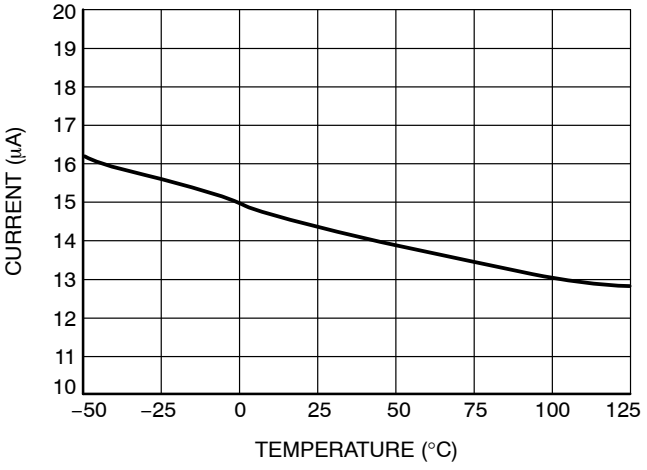

Figure 22. SEQ Internal Pull−down Current vs. Temperature

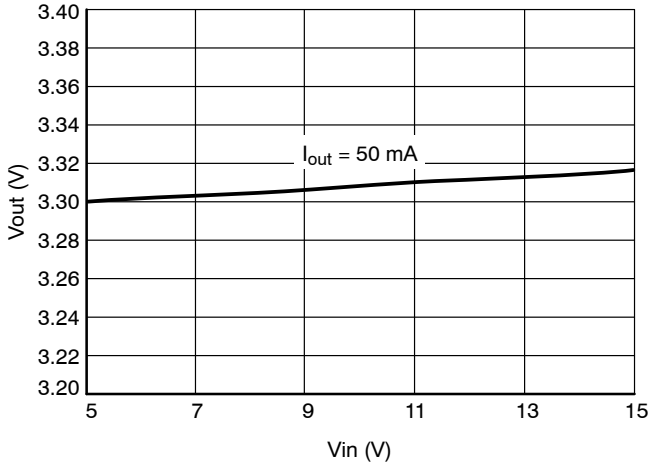

Figure 23. NCP3121 Line Regulation

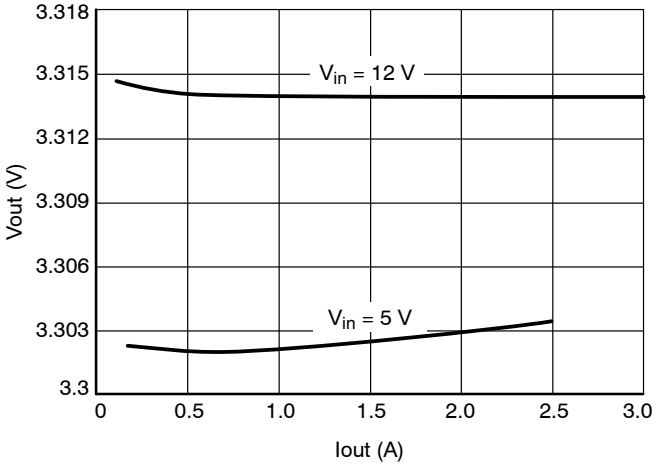

Figure 24. NCP3121 Load Regulation

TYPICAL OPERATING CHARACTERISTICS

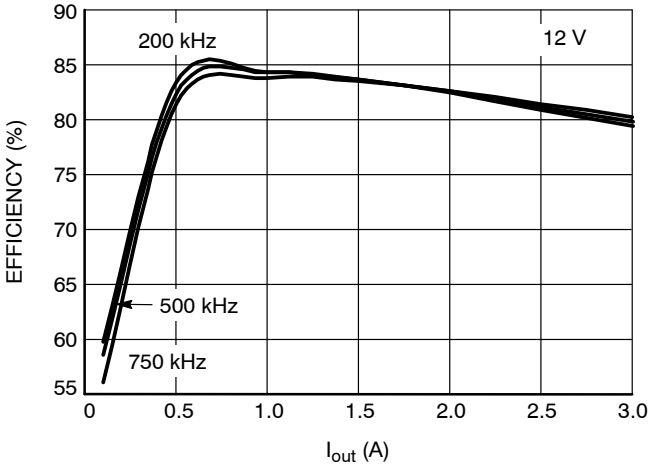

Figure 25. NCP3121 Efficiency, V_{in} = 12 V, V_{out} = 3.3 V, 25°C

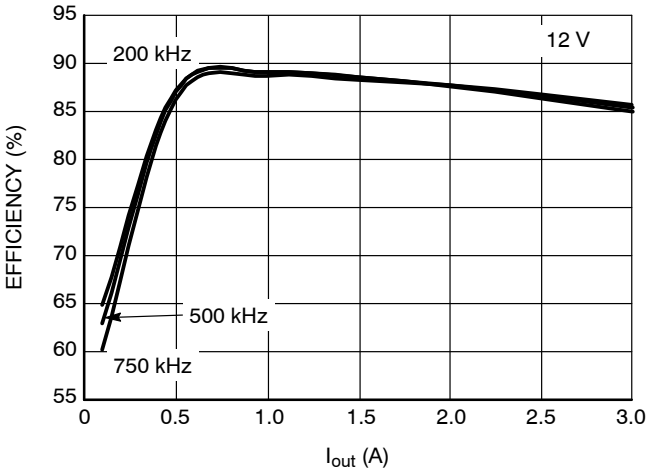

Figure 26. NCP3121 Efficiency, V_{in} = 12 V, V_{out} = 5 V, 25°C

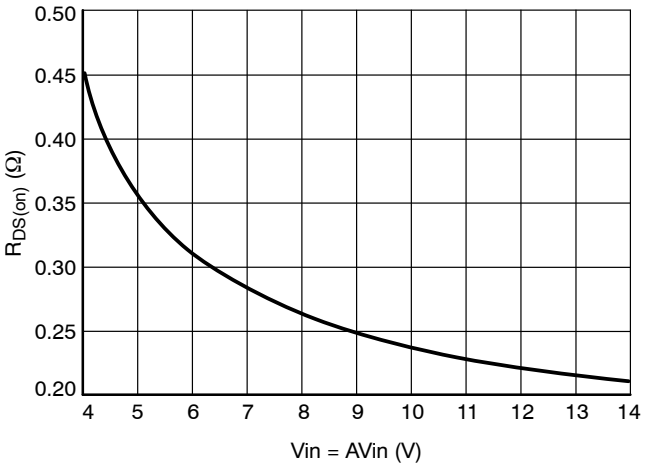

Figure 27. $R_{DS(on)}$ vs. Input Voltage

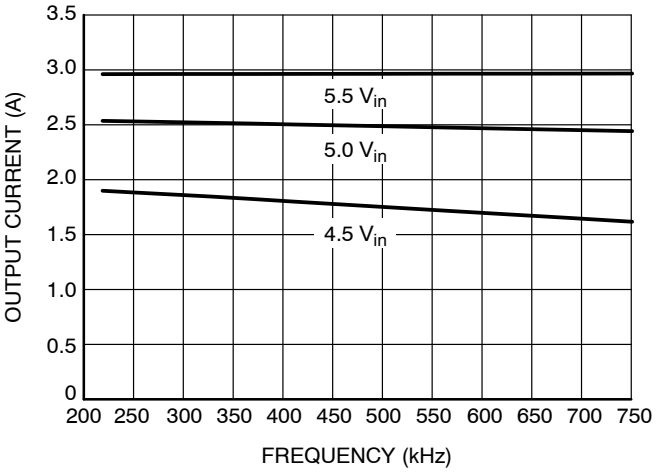

Figure 28. Maximum Currents vs. Operating Frequency due to Toff min limitations 3.3 Vout

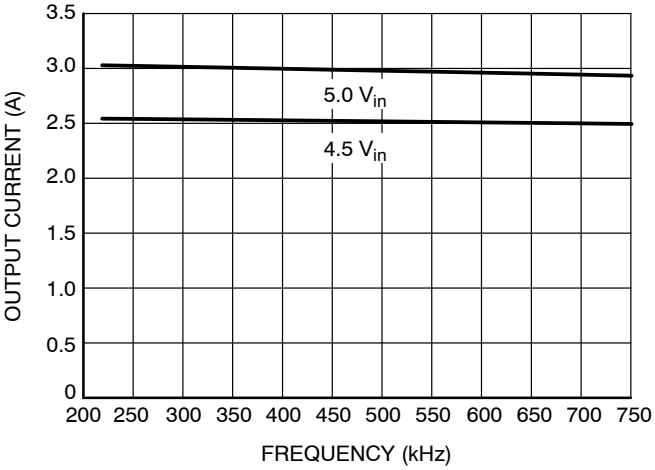

Figure 29. Maximum Currents vs. Operating Frequency due to Toff min limitations 1.8 Vout

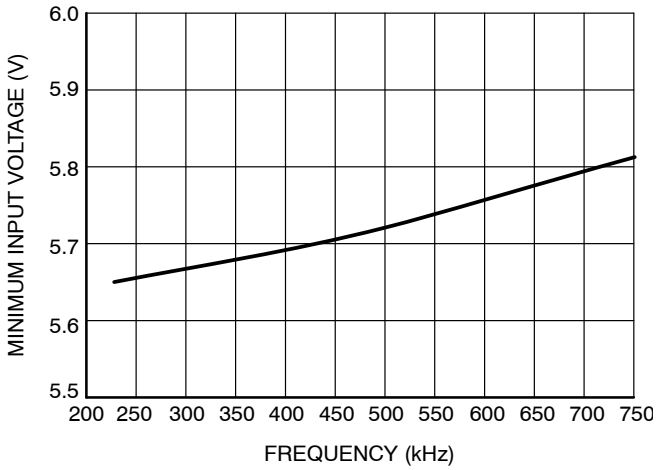

Figure 30. Minimum Input Voltage vs. Operating Frequency, 3A, 3.3 Vout

TYPICAL OPERATING CHARACTERISTICS

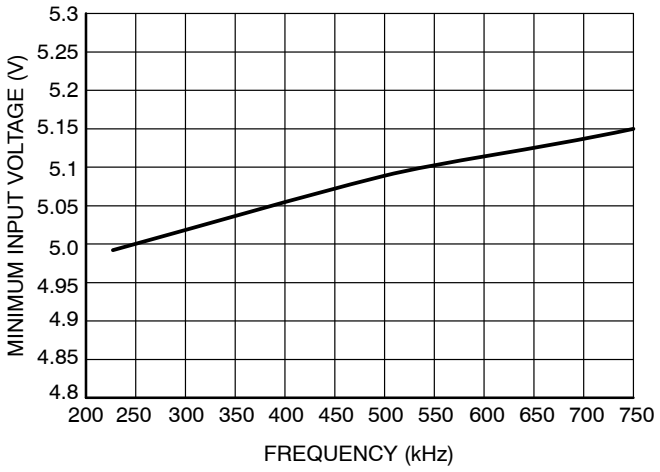

Figure 31. Minimum Input Voltage vs. Operating Frequency, 3A, 1.8 Vout

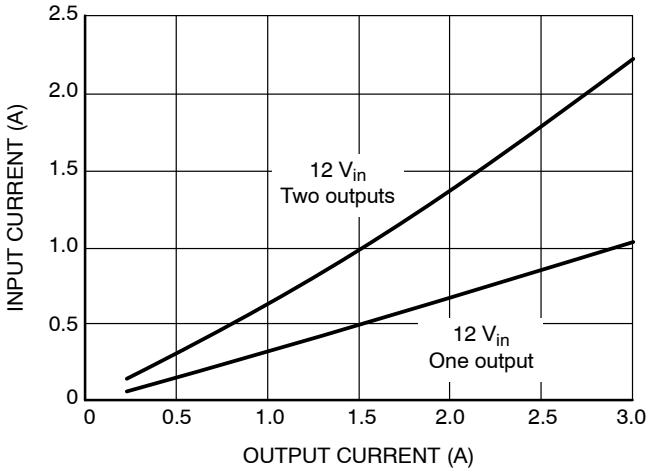

Figure 32. Minimum Input Current 3.3 Vout

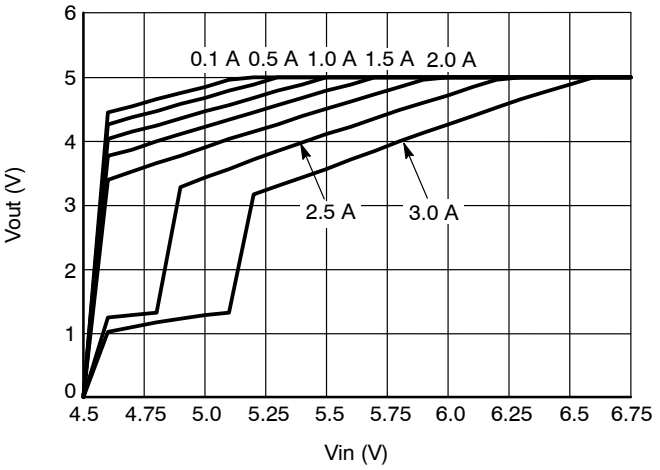

Figure 33. Minimum Input Voltage 5 Vout, 350 kHz

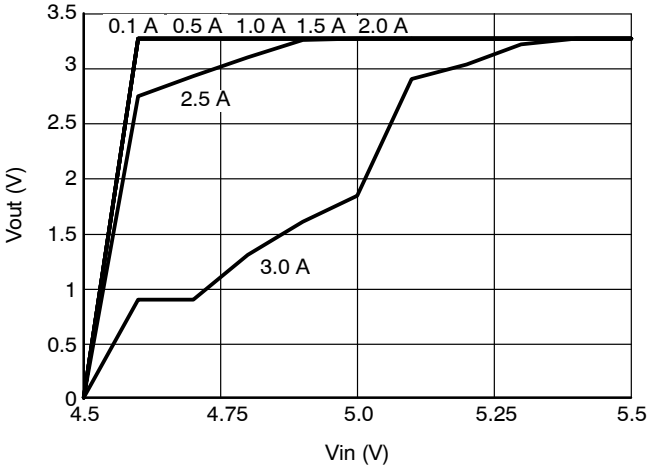

Figure 34. Minimum Input Voltage 3.3 Vout, 350 kHz

DETAILED DESCRIPTION

Introduction

The NCP3121 is a dual channel non−synchronous PWM voltage mode buck regulator. Each channel is identical and has a 3.0 A internal P−FET, compensation, feedback, programmable soft−start, enable and power good pins. These circuits also share the same input voltage, reference voltage, thermal shutdown, undervoltage detect and master oscillator. A simple auto−tracking and sequencing capability can be implemented using the SEQ/TRACK/SS pins.

The fixed−frequency programmable architecture, driven from a common oscillator, ensures a 180° phase differential between channels. This 180° phase shift between the two channels reduces the common input capacitor requirement and improves the noise immunity. The NCP3121 switching frequency is set by an external resistor and is adjustable between 200−750 kHz. This allows application optimization between efficiency and total solution size.

The output voltage is fed back through an external resistor voltage divider to the FB input pin and compared with the reference voltage, then the voltage difference is amplified through the internal transconductance error amplifier. The output current of the transconductance error amplifier (OTA) is presented at the COMP node where an RC network compensates the regulation control system loop.

The NCP3121 features a programmable soft−start function, which is implemented through the error amplifier and the external compensation capacitor. This feature prevents stress to the power components and limits output voltage overshoot during start−up.

Undervoltage Lockout (UVLO)

Undervoltage lockout (UVLO) is provided to ensure that unexpected behavior does not occur when V_{in} is too low to support the internal rails and power the converter. In case the input voltage is higher than the UVLO threshold (4.3 V standard value, rising voltage), the step down converter operation can be started. This circuit has a 0.2 V hysteresis (typical). If the falling trip is activated, switching ceases and eventually the circuit turns off. When the input circuit is in this state, the currrent consumption is equal 5 mA (typical).

Fixed Frequency Operation

The NCP3121 uses a constant frequency architecture for generating a PWM signal. During normal operation, the oscillator generates an accurate pulse at the beginning of each switching cycle to turn on the main switch. The main switch will be turned off when the ramp signal intersects with the output of the error amplifier (COMP pin voltage). Therefore, the switch duty cycle can be modified to regulate the output voltage to the desired value as line and load conditions change.

The major advantage of fixed frequency operation is that the component selections, especially the magnetic component design, become very easy. The oscillator frequency of the NCP3121 is programmable from 200 kHz to 750 kHz using an external resistor connected from the RT pin to ground. The oscillator works on the double frequency internally. Therefore, both channels have a 180° phase shift of the SW pins.

Out−of−Phase Operation

In out−of−phase operation, the turn−on of the second channel is delayed by half the switching cycle. This delay is supervised by the oscillator, which supplies a clock signal to the second channel which is 180° out of phase with the clock signal of the first channel. The advantages of out−of−phase synchronization are many. Since the input current pulses are interleaved with one another, the overlap time is reduced. The effect of this overlap reduction is to attenuate the input filter requirement, allowing the use of smaller components. Additionally, since peak current occurs during a shorter time period, emitted EMI is also reduced, thereby reducing shielding requirements.

Enable Input

Pull the EN enable input high to enable operation. The EN high signal must occur **after VIN has exceeded 2.7 V** to allow internal Power-on Reset (POR) logic to initialize the IC. Logic low on SEQ forces the NCP3121 into shutdown mode. Connect SEQ to EN for normal operation of a standalone device. In shutdown mode, the NCP3121 is turned off and the supply current is reduced to less than 100 μA. When the enable function is not required, float the EN connection. The NCP3121 will turn itself on once Vin crosses the input UVLO threshold. **Do not pull EN to VIN or a separate supply voltage**. For standalone operation, EN should still be connected to SEQ.

Note: For proper operation of the NCP3121 circuit, no voltage may be pulled high on the output pins. The output capacitors should be discharged. If this condition is not observed when NCP3121 is enabled, the regulator does not start switching. This helps to prevent improper operation of the NCP3121 circuit due to the implemented tracking and sequencing features.

Soft−Start/Stop Control

This capacitor limits the maximum demand on the external power supply by controlling the inrush current peaks to charge the output capacitor and DC load and to attain smoothly increasing output voltage at start−up. A soft start circuit forces the error amplifier output to follow a prescribed voltage ramp when turning on and off. The output capacitor is discharged when V_{in} goes under the UVLO as thermal shutdown or overload detection occurs. The circuit input is presented as a voltage ramp generated by internal current sources tied to an external SS capacitor. The external capacitor on the soft−start node is charged/discharged by the 8.75 μA current from the constant current source, and the voltage on the SS node controls the OTA amplifier output

voltage until the SS capacitor is charged/discharged to a voltage higher than 0.8 V.

Power Good

The power good is an open drain and active high output that indicates when the output voltage has reached 90% (min) of the nominal output voltage. This output can be pulled up to the appropriate level with an external resistor.

The power good comparator senses the voltage at the FB pin, which is a function of V_{out}.

The power good output transistor behavior is shown in the "Typical Operating Characteristics" section. The PG pin is held low during a soft−start. Once a soft−start is completed, the PG goes high if there are no faults and no delays associated with it.

Current Limit

The NCP3121 protects a power system if overcurrent occurs. The NCP3121 contains pulse−by−pulse current limiting to protect the power switch and external components. The current through each channel is continuously monitored. The current limit is set to allow peak switch current in excess of 3.5 A (minimum). Current limiting is implemented by monitoring the high−side P−channel switch current during conduction with a current limit comparator. When the peak of the switching current reaches the current limit, the power switch turns off.

Hiccup Overload Protection (OLM – Over Load Mode)

Hiccup mode is a method of protecting the power supply from damage during overload conditions. Within normal operation, the external soft−start capacitor is pulled up by a current source that delivers 8.75 μA to the SS pin capacitor. The soft−start capacitor continues to charge until it reaches the saturation voltage of the current source, typically V_{ss} = 4 V. When the overload condition is detected, the soft−start capacitor is discharged to 0.1 V and is again charged to 1 V. This is periodically repeated until the overload condition is detected. The transconductance error amplifier output is tied to ground when the soft−start capacitor is discharged.

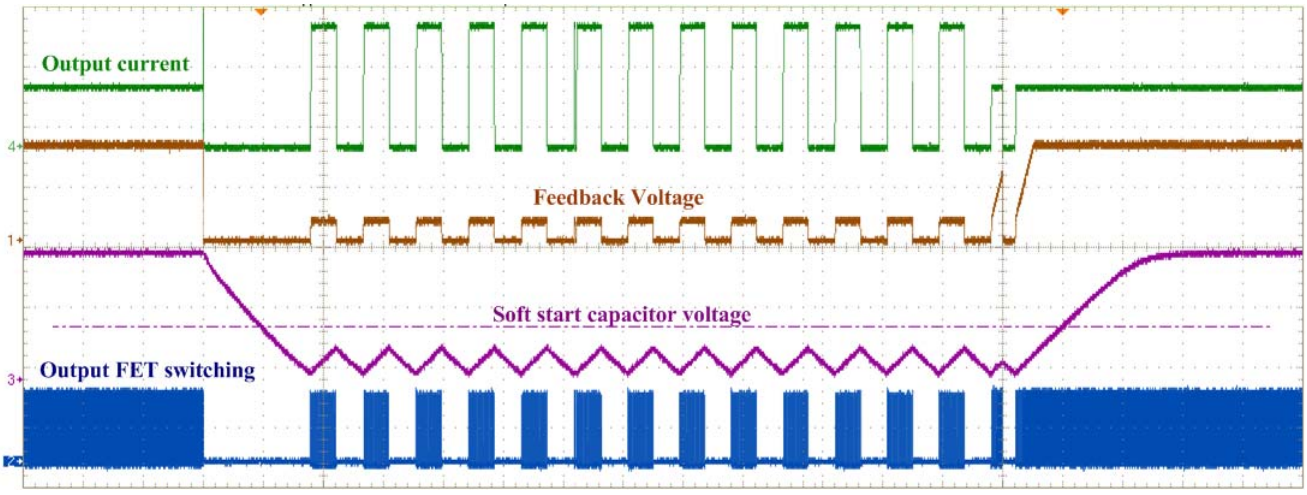

Figure 35. Hiccup Overload Protection

Thermal Shutdown

The NCP3121 has a thermal shutdown feature to protect the device from overheating when the die temperature exceeds 160°C (typically). If the chip temperature exceeds the overtemperature shutdown trip point, the fault signal is activated. This will disable the step down converter operation, and the chip temperature will start to decrease. When the chip temperature drops 15°C below the overtemperature shutdown trip point, the fault signal is deactivated and the step down converter operation starts again with soft−start. The thermal event sends the device immediately into the OFF state. The currrent consumption is equal 5 mA (typical) if the thermal condition is reached.

APPLICATION & DESIGN INFORMATION

Inductor

The output inductor may be the most critical component in the converter because it will directly affect the choice of other components and dictate both the steady state and transient performance of the converter. When choosing inductors, one might have to consider maximum load current, core and copper losses, component height, output ripple, EMI, saturation and cost. Lower inductor values are chosen to reduce the physical size of the inductor. A higher value cuts down the ripple current and core losses and allows more output current. In general, the output inductance value should be as low and the output inductor physically as small as possible to provide the best transient response and minimum cost. If a large inductance value is used, the converter will not respond quickly to rapid changes in the load current. On the other hand, an inductance value that is too low will result in very large ripple currents in the power components, resulting in increased dissipation and lower converter efficiency.

A good standard for determining the inductance to use is to select the inductor peak−to−peak ripple current to be approximately 25% of the maximum switch current. Also, make sure that the inductor peak current is below the maximum switch current limit and the selected inductor type saturation current specification is higher than the peak current through the switch.

The maximum current in the inductor while operating in the continuous current mode is defined as the load current plus one half of the ΔI_L currrent:

$$I_{LP} = I_{LOAD} + \frac{1}{2}\Delta I_L$$

The inductance value can be calculated by:

$$L = \frac{V_{OUT}(V_{IN} - V_{OUT})}{V_{IN} \cdot \Delta I_L \cdot f_{OSC}}$$

Therefore, the inductor peak current, I_{LP}, can be calculated by:

$$I_{LP} = I_{LOAD} + \frac{V_{OUT}(V_{IN} - V_{OUT})}{2 \cdot V_{IN} \cdot L \cdot f_{OSC}}$$

where;

I_{LOAD} is the output load current

V_{OUT} is the output voltage

V_{IN} is the input voltage

ΔI_L is the peak−to−peak inductor ripple current

f_{OSC} is the switching frequency of the oscillator

The choice of the appropriate inductor type depends not only on the calculated inductance value, saturation current rating and parasitic serial resistance, but also on the required physical dimensions, EMI requirements (shielded or open inductor) and the price. Examples of suitable inductors from various manufacturers are shown in the table below.

Table 1. Calculated Inductor Values

Calculated coils, I ripple peak−peak 20%					
	f [kHz]	**200**	**350**	**500**	**750**
	I_{out} [A]				
12 V_{in} to 7.5 V_{out}	2 A	36 μH	20 μH	14 μH	10 μH
	3 A	24 μH	14 μH	10 μH	6.5 μH
12 V_{in} to 5 V_{out}	2 A	36 μH	20 μH	15 μH	10 μH
	3 A	24 μH	14 μH	10 μH	6.5 μH
12 V_{in} to 3.3 V_{out}	2 A	30 μH	17 μH	12 μH	8 μH
	3 A	20 μH	12 μH	8 μH	5.4 μH
5 V_{in} to 3.3 V_{out}	2 A	14 μH	8 μH	5.6 μH	3.7 μH
5 V_{in} to 2.5 V_{out}	2 A	16 μH	9 μH	6.3 μH	4 μH
5 V_{in} to 1.8 V_{out}	2 A	15 μH	8.2 μH	5.8 μH	3.8 μH
	3 A	10 μH	5.5 μH	3.8 μH	2.6 μH

Table 2. Inductor Examples

L [μH]	Part Number	Shielded/ Non−shielded	I_{rms} [A]	DCR max [mΩ}	Manufacturer	Web
33	DO5010H−333	N	3.0	66	Coilcraft	www.coilcraft.com
	PF0382.333NL	N	3.1	65	PULSE	www.pulseeng.com
	MSS1278−333	S	3.1	80	Coilcraft	www.coilcraft.com
	74458133	N	3.0	66	WE	www.we−online.com
	PF0552.333NL	S	3.7	54.1	PULSE	www.pulseeng.com
22	DS5022P−223	S	3.1	59	Coilcraft	www.coilcraft.com
	P0648.223	N	3.3	61	PULSE	www.pulseeng.com
	74458122	N	3.5	47	WE	www.we−online.com
	MSS1246T−223	S	3.14	70	Coilcraft	www.coilcraft.com
	PF0382.223NL	N	3.5	47	PULSE	www.pulseeng.com
15	DO3316P−153	N	3.1	46	Coilcraft	www.coilcraft.com
	P0751.153NL	N	3.0	46	PULSE	www.pulseeng.com
	MSS1260T−153	S	3.5	40	Coilcraft	www.coilcraft.com
	74459115	S	3.5	48	WE	www.we−online.com
	74458115	N	4.0	36	WE	www.we−online.com
10	DO3340P−103	N	3.5	40	Coilcraft	www.coilcraft.com
	DS5022P−103	S	3.9	42	‘Coilcraft	www.coilcraft.com
	7445610	N	3.3	45	WE	www.we−online.com
	74459010	S	3.9	40	WE	www.we−online.com
	DO3316P−103	N	3.5	34	Coilcraft	www.coilcraft.com
	P0751.103NL	N	3.8	38	PULSE	www.pulseeng.com
9	P1169.123NL	S	3.5	37	PULSE	www.pulseeng.com
8.2	DS3316T−822	N	4.15	32	Coilcraft	www.coilcraft.com
	MSS1246−822	S	4.67	35	Coilcraft	www.coilcraft.com
6.8	74456068	N	3.8	34	WE	www.we−online.com
5.6	DO33165−562	N	4.65	21	Coilcraft	www.coilcraft.com
	74456056	N	4.0	32	WE	www.we−online.com
	DO5022P−562	S	4.1	3	Coilcraft	www.coilcraft.com
5.0	MSS7341−502	S	4.7	24	Coilcraft	www.coilcraft.com
3.3	DO3316P−332	S	4.7	26	Coilcraft	www.coilcraft.com
	DS5022P−332	S	3.3	39	Coilcraft	www.coilcraft.com

Output Rectifier Diode

When the high−side switch is on, energy is stored in the magnetic field in the inductor. During off time, the internal MOSFET switch is off. Since the current in the inductor has to discharge, the current flows through the rectifying diode to the output. A Schottky diode is recommended due to low diode forward voltage and very short recovery times, which positively impacts the step down voltage converter’s overall efficiency. Another choice could be fast recovery or ultra−fast recovery diodes. It should be noted that some types of these diodes with an abrupt turn−off characteristic may cause instability or EMI troubles.

The peak reverse voltage is equal to the maximum input voltage. The peak conducting current is clamped by the current limit of the NCP3121. Use of Schottky barrier diodes reduces diode reverse recovery input current spikes. For switching regulators operating at low duty cycles, it is beneficial to use rectifying diodes with somewhat higher RMS current ratings (thus lower forward voltages). This is because the diode conduction interval is much longer than that of the transistor. Converter efficiency will be improved if the voltage drop across the diode is lower. The average current can be calculated from:

$$I_{D(AVG)} = \frac{I_{LOAD}(V_{IN} - V_{OUT})}{V_{IN}}$$

Table 3. Schottky Diode Example

Part Number	Description	V_{RRM} min [V]	V_F max [V]	$I_{O(rec)}$ max [A]	Package	Web
MBRA340T3G	3 A, 40 V Schottky Rectifier	40	0.45	3	SMA	www.onsemi.com
MBRS340T3G	3 A, 40 V Schottky Rectifier	40	0.5	3	SMC	www.onsemi.com
MBRS330T3G	3 A, 30 V Schottky Rectifier	30	0.5	3	SMC	www.onsemi.com

The worst case of the diode average current occurs during maximum load current and maximum input voltage. The rectifying diodes should be placed close to the SW pin to avoid the possibility of ringing due to trace inductance.

Input Capacitor

The input current to the step down converter is discontinuous. The input capacitor has to maintain the DC input voltage and to sustain the ripple current produced by internal MOSFET switching. For stable operation of the switch mode converter, a low ESR capacitor is needed to prevent large voltage transients from appearing at the input. Therefore, ceramic capacitors are preferred, but the circuit works in a stable manner also with electrolytic capacitors. It must be located near the regulator and use short leads. Also, paralleling ceramic capacitors will increase the regulator stability.

The RMS value of the input capacitor current ripple is:

$$I_{RMS} = I_{LOAD}\sqrt{D(1 - D)}$$

The duty cycle is:

$$D = \frac{V_{OUT} + V_D}{V_{IN} + V_D - V_{DSAT}}$$

where:

V_D is the voltage drop across the rectifying diode and

V_{DSAT} is the switch saturation voltage on the power MOSFET.

The equation reaches its maximum value with duty cycle = 0.5, where:

$$I_{RMS} = \frac{I_{LOAD}}{2}$$

Losses in the input capacitor can be calculated using the following equation:

$$P_{CIN} = I_{RMS}^2 \cdot ESR_{CIN}$$

where:

ESR_{CIN} is the effective series resistance of the input capacitance.

The input capacitor voltage ripple depends on the C_{IN} capacitor value. Therefore, the input capacitor can be estimated by:

$$C_{IN} = \frac{I_{LOAD}}{f_{SW} \cdot \Delta V_{IN}} \cdot \frac{V_{OUT}}{V_{IN}} \cdot \left(1 - \frac{V_{OUT}}{V_{IN}}\right)$$

Output Capacitor

The output capacitor filters output inductor ripple current and provides low impedance for load current changes. The principle consideration for the output capacitor is the ripple current induced by the switches through the inductor. It supplies the current to the load in DCM or during load transient and filters the output voltage ripple. For low output ripple voltage and good stability, low ESR output capacitors are recommended. The inductor ripple current acting against the ESR of the output capacitor is the major contributor to the output ripple voltage.

An output capacitor has two main functions: it filters the output and provides regulator loop stability.

The ESR of the output capacitor and the peak−to−peak value of the inductor ripple current are the main factors contributing to the output ripple voltage value.

The output voltage ripple is given by the following equation:

$$\Delta V_{OUT} = \frac{V_{OUT}}{f_{SW} \cdot L} \cdot \left(1 - \frac{V_{OUT}}{V_{IN}}\right) \cdot \left(ESR + \frac{1}{8 \cdot f_{SW} \cdot C_{OUT}}\right)$$

where:

ESR is the equivalent series resistance of the output capacitor.

The output capacitor value can by expressed by:

$$C_{OUT} = \frac{\Delta I_L}{8 \cdot f_{SW} \cdot (\Delta V_{OUT} - \Delta I_L \cdot ESR)}$$

These components must be selected and placed carefully to yield optimal results. Key specifications for output capacitors are their ESR (equivalent series resistance) and ESL (equivalent series inductance) values. For best transient response, a combination of low value/high frequency and bulk capacitors placed close to the load will be required.

For most applications, a 22 μF ceramic capacitor should be sufficient. X5R or X7R dielectrics ceramic capacitors are recommended.

Soft−Start Capacitor Selection

The soft−start time is programmed by an external capacitor connected from the SS pin to AGND, which can be calculated by:

$$C_{SS} \approx \frac{t_{SS} \cdot 8.75\ \mu A}{0.8\ V}$$

where:

− t_{SS} is the soft−start/stop interval.

Note: See the “Sequencing and Tracking” section on how to use this capacitor.

Output Voltage Programming

The controller will maintain 0.8 V at the feedback pin. Thus, if a resistor divider circuit is placed across the feedback pin to V_{OUT}, the controller will regulate the output voltage in proportion to the resistor divider network in order to maintain 0.8 V at the FB pin.

Table 4. Output Voltage Setting

V_{OUT} [V]	8	7.5	6	5	4	3.3	2.5	1.8	1.2
R_1 [kΩ]	180	360	130	68	300	47	51	20	10
R_2 [kΩ]	20	43	20	13	75	15	24	16	20

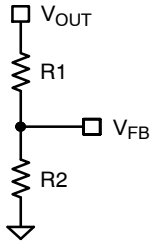

Figure 36. Feedback divider

The relationship between the resistor divider network and the output voltage is shown in the following equation:

$$R_2 = R_1\left(\frac{V_{REF}}{V_{OUT} - V_{REF}}\right)$$

where:

V_{REF} is the circuit's internal voltage reference, which equals 0.8 V.

Resistor R1 is selected based on a design trade−off between efficiency and output voltage accuracy. For high values of R1, there is less current consumption in the feedback network. However, the trade−off is output voltage accuracy due to the bias current in the error amplifier. Once R1 has been determined, R2 can be calculated.

Selecting the Switching Frequency

Selecting the switching frequency is a trade−off between component size and power losses. Operation at higher switching frequencies allows the use of smaller inductor and capacitor values. Nevertheless, it is common to select lower frequency operation because a higher frequency results in lower efficiency due to MOSFET gate charge losses. Additionally, the use of smaller inductors at higher frequencies results in higher ripple current, higher output voltage ripple, and lower efficiency at light load currents. The value of the oscillator resistor is designed to be linearly related to the switching period. There are two ways to determine the RT resistor value: by using the standard curve shown in Figure 37 or by using Table 5. The frequency on the RT pin will set the master oscillator. The actual operating frequency on each channel will be one−half the master oscillator.

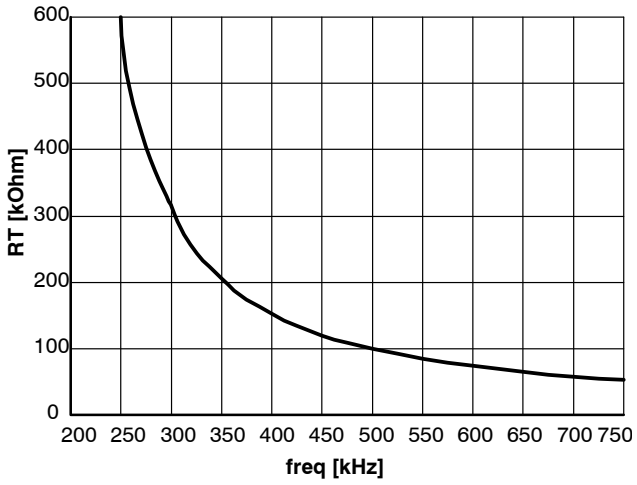

Figure 37. Switching Frequency Selection

Table 5. Switching Frequency Selection

Freq. [kHz]	200	250	300	350	400	450	500	550	600	650	700	750
RT [kΩ]	open	649	316	205	154	121	100	84.5	73.2	64.9	57.6	52.3

Sequencing of Output Voltages

Some microprocessors and DSP chips need two power supplies with different voltage levels. These systems often require voltage sequencing between the core power supply and the I/O power supply. Without proper sequencing, latch−up failure or excessive current draw may occur that could result in damage to the processor's I/O ports or the I/O ports of a supporting system device such as memory, an FPGA or a data converter. To ensure that the I/O loads are not driven until the core voltage is properly biased, tracking of the core supply and the I/O supply voltage is necessary.

Designing a system without proper power supply sequencing for signal processing devices like DSPs, FPGAs, and PLDs may create risks as to reliability or proper functionality. The risk comes when there are active and inactive power supply rails on the device for a long time. During this time, the ESD structures, internal circuits and components are stressed from interference between different voltages (from the two separate power supply rails). When these conditions persist on multi−supply devices for long time periods (this is a cumulative phenomenon), the life of the products (DSP, FPGA, and

PLD devices) is drastically reduced. The failure is often a result of high currents flowing to the pins or the high voltage difference between pins.

In that case, the signal processors require multiple power supplies generating different voltage levels for core and I/O peripherals over time. NCP3121 offers ratiometric sequencing, sequential sequencing and tracking sections to manage the output voltages behavior during start−up and power−down. Basically, the DSP, FPGA, and PLD manufacturers do not specify the method of power sequencing, but they do specify restrictions on the time or voltage differences during power−up and power−down. The power−up sequence for microprocessors should be finished approximately within a few seconds to prevent the risks mentioned above. For more information, see the microprocessor manufacturers' datasheets.

Ratiometric Sequencing

In the ratiometric sequencing mode, multiple outputs start ramping at the same time and also reach the regulation level at the same time. When common EN is pulled down, the output voltages are going down at the same time. See Figure 38. This functionality is created by using the same capacitor values as the soft−start capacitors for all outputs and by connecting all EN + SEQ pins together. To ensure this behavior, the soft start capacitors should have values greater than the time constant of the output inductor and output capacitor.

For proper operation in this mode, using a common soft−start capacitor for both channels is not recommended.

Note: If enable control is not required, float the EN/SEQ connections rather than pulling them to VIN or a separate supply voltage. The NCP3121 will enable itself once VIN crosses the input UVLO threshold.

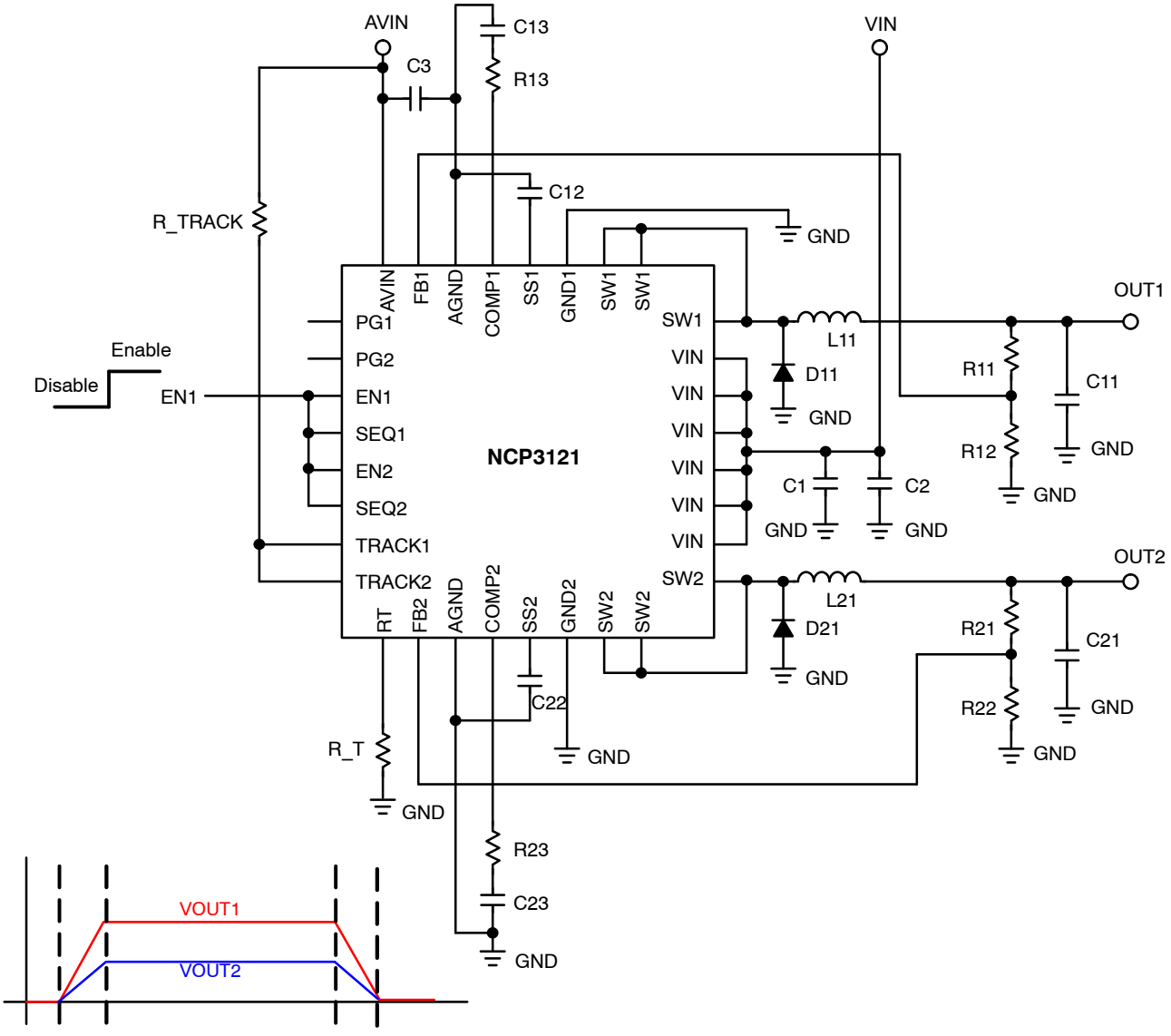

Figure 38. Ratiometric Sequencing Configuration

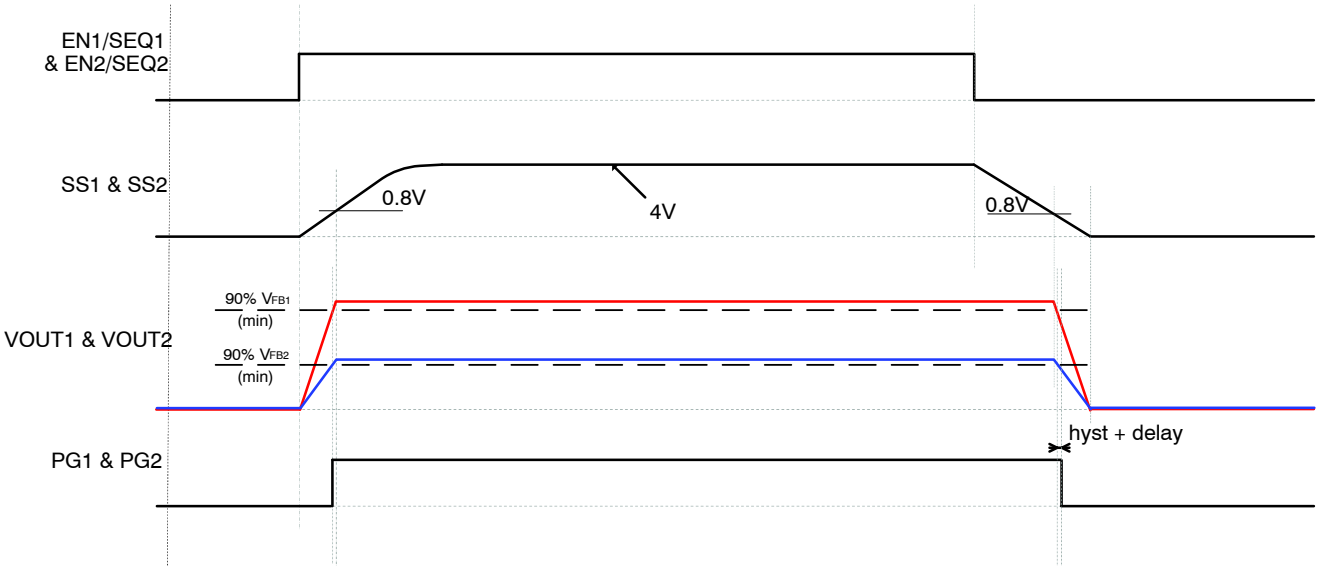

Figure 39. Typical Behavior of Ratiometric Sequencing Mode

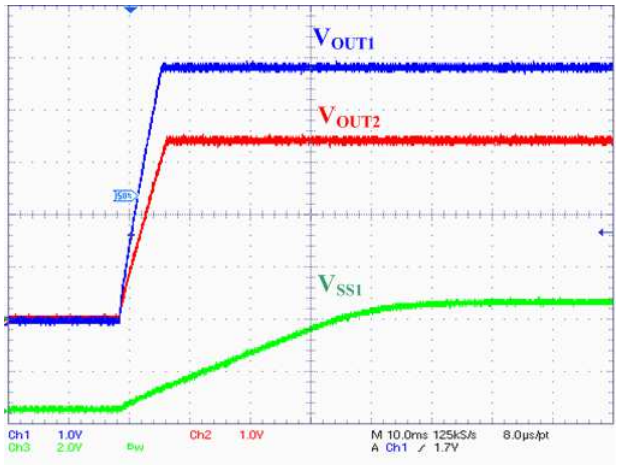

Figure 40. Ratiometric Mode − Power−up

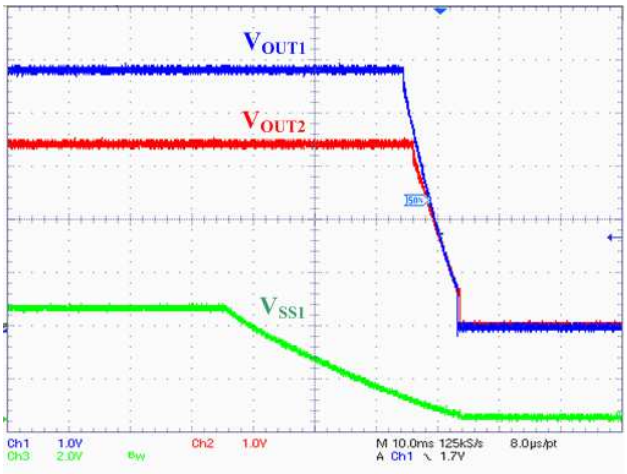

Figure 41. Ratiometric Mode − Power−down

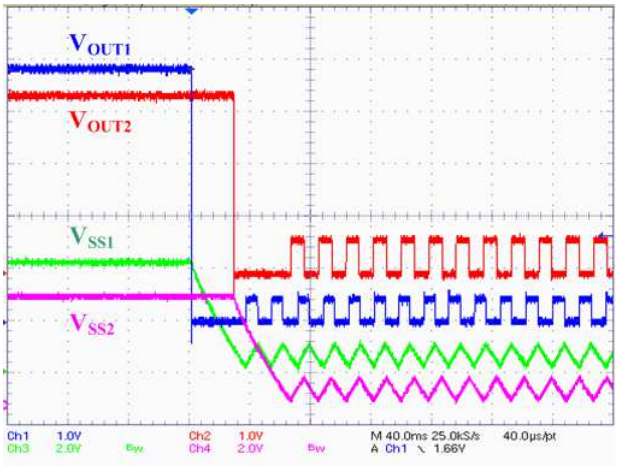

Figure 42. Ratiometric Mode − Start of OLM

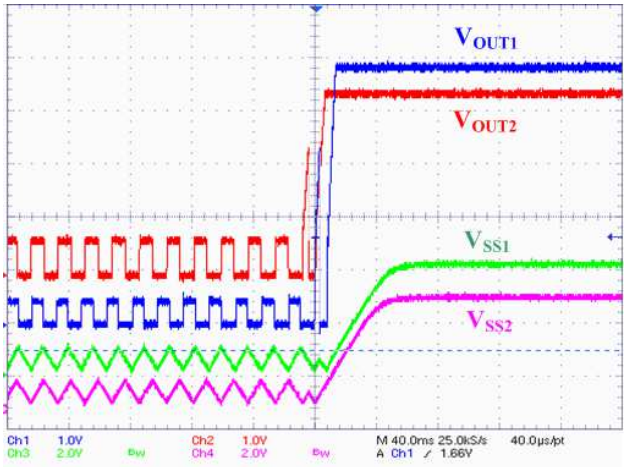

Figure 43. Ratiometric Mode − End of OLM

Sequential Sequencing (First−Up/Last−Down Sequence Configuration)

In sequential sequencing mode, the second output voltage starts ramping when the first output voltage is already settled and its power good signal is set. Figure 44 shows the NCP3121's configuration and standard waveforms. The rising slope of both voltages can be selected independently by the soft−start capacitors' values (C12, C22). When the enable pin is deactivated, the second output voltage decreases first, followed by the first output voltage. The control logic is based on the internal power good signal; no delay is added. The signal has the same threshold values as the power good signal shown in the electrical table. The sequential sequencing mode is also called first−up/last−down and is ideal for DSPs with separate power supplies for the core and the I/O ports.

Note: If enable control is not required, float the EN(first)/SEQ(last) connection rather than pulling it to VIN or a separate supply voltage. For Figure 44 this is EN1/SEQ1. For Figure 45, this is EN1/SEQ4. The NCP3121 will enable itself once VIN crosses the input UVLO threshold.

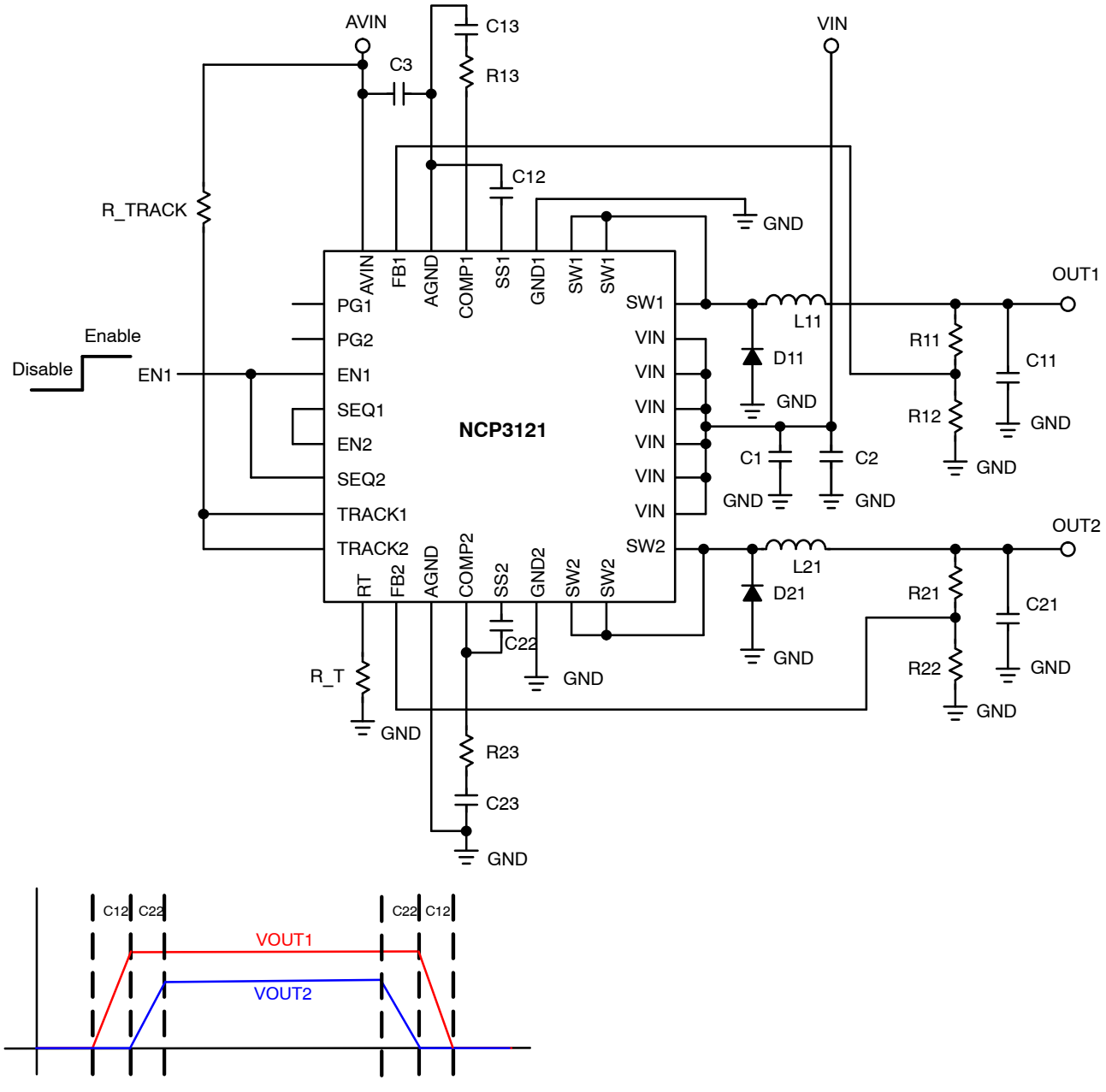

Figure 44. Sequential Configuration

Daisy Chain Operation

The last−up/first−down power output has its SEQ pin tied to the EN of the first−up/last−down power output. Each output in the chain has its power−up delay set by the soft−start ramp−up of the supply. This feeds its EN and its power−down delay set by the soft−start ramp−down of the supply that feeds its SEQ pin.

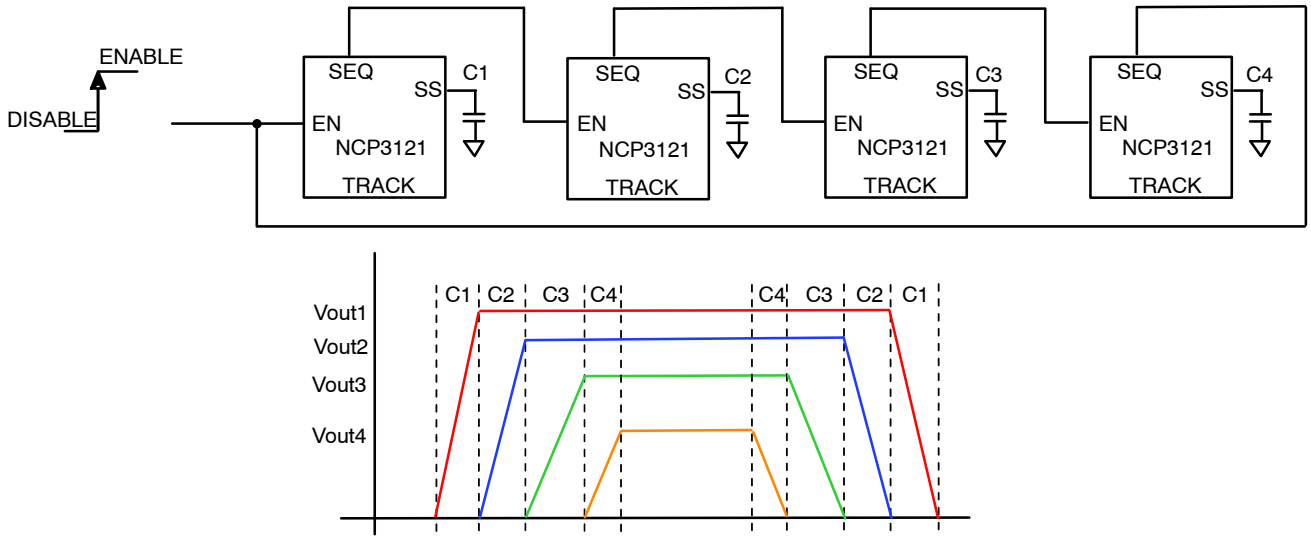

Figure 45. Simplified Drawing of Daisy−chained NCP3121's

When the first voltage rail has reached a specific voltage level, the next voltage rail is enabled and its rise is monitored until it has reached the power good trip point. At this point, the next voltage rail is enabled. This continues until all voltage rails have been enabled (see Figure 46). Power−down sequencing is just the opposite of the power−up sequence.

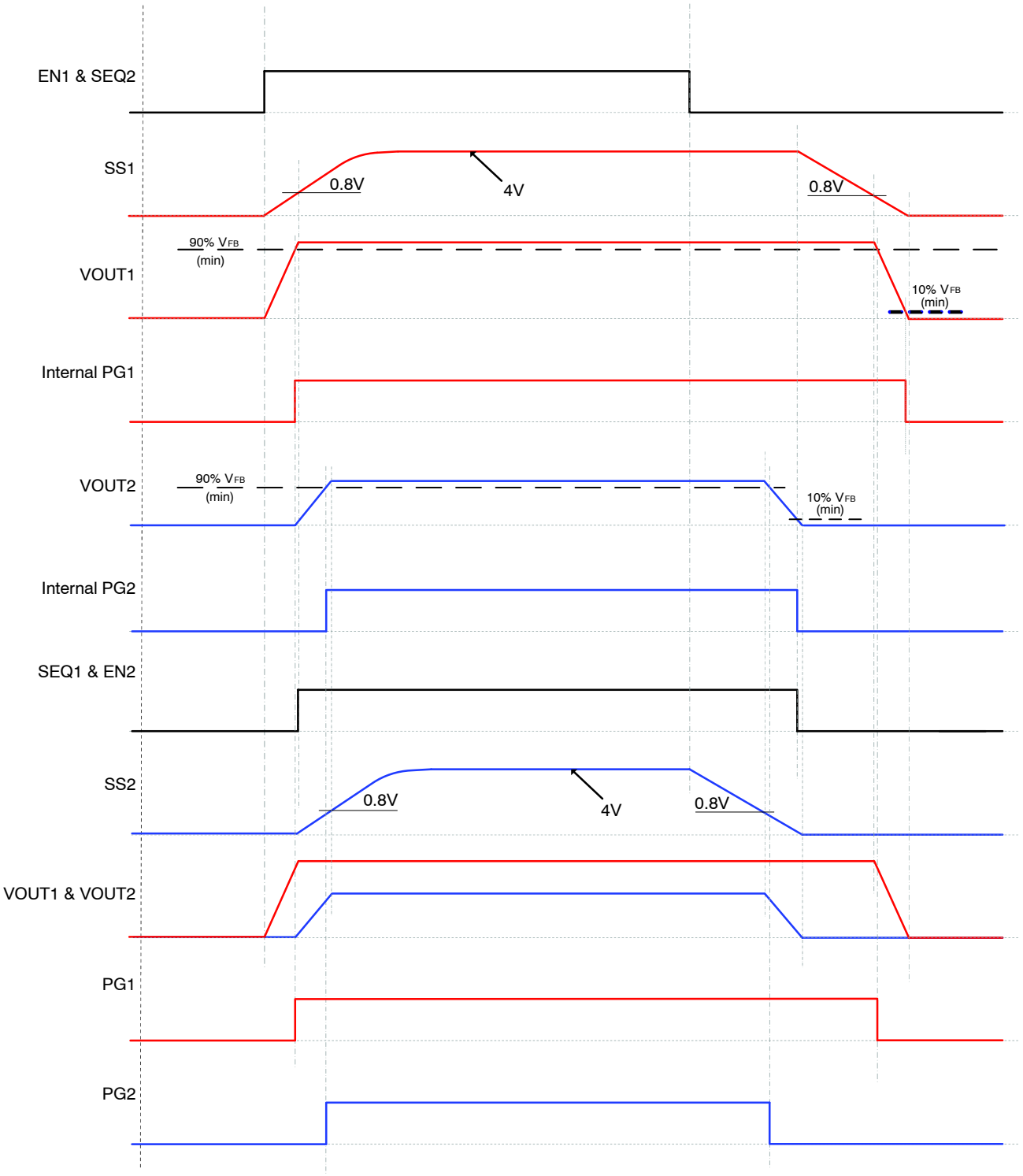

Figure 46. Typical Behavior of Sequential Mode

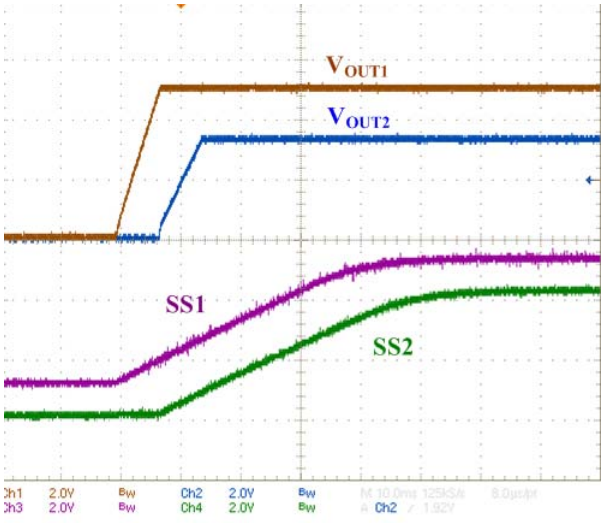

Figure 47. Sequential Mode − Power−up

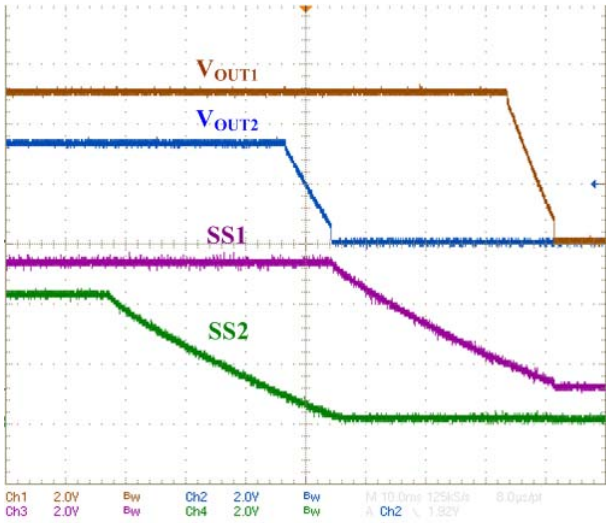

Figure 48. Sequential Mode − Power−down

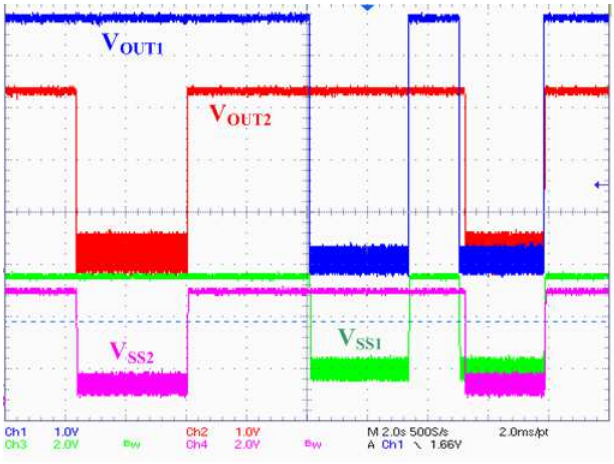

Figure 49. Sequential Mode − Power−down

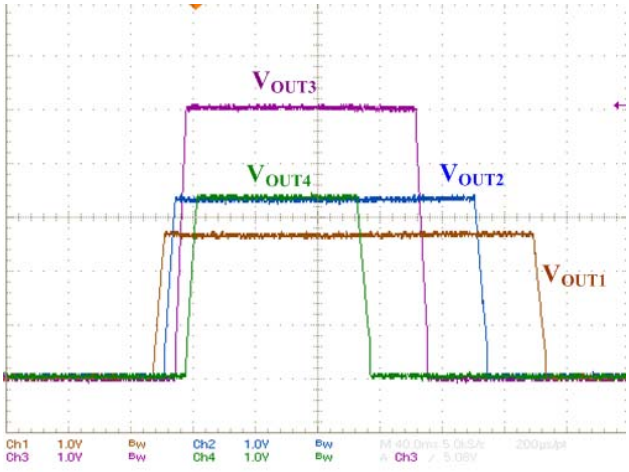

Figure 50. Daisy Chain of Four Outputs

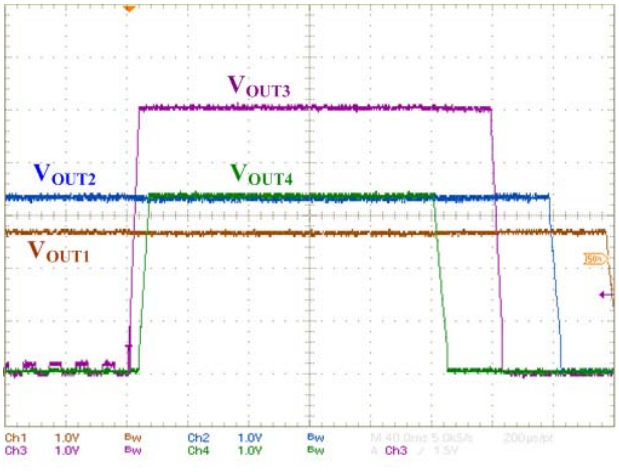

Figure 51. OLM of the 3rd Output in Daisy Chain

Tracking

Voltage tracking is enabled by applying a ramp voltage to the TRACK pin. When the voltage on the TRACK pin is below 0.8 V, the feedback voltage will regulate to this tracking voltage. When the tracking voltage exceeds 0.8 V, tracking is disabled and the feedback voltage will regulate to the internal reference voltage.

In this start−up sequence, the tracking pin is used to match the output voltage ramps exactly. Higher output voltage will continue to rise past the lower regulated point. This is achieved by dividing the higher output voltage by the same ratio as the lower voltage feedback components and connecting the divided voltage into the TRACK pin of the lower voltage. Track pins must be tied high in the normal operation (except in the tracking mode).

The output voltage during tracking can be calculated with the following equation:

$$V_{OUT} = V_{TRACK}\left(1 + \frac{R5}{R6}\right) \qquad V_{TRACK} < 0.8\ V$$

Note: If enable control is not required, float the EN/SEQ connections rather than pulling them to VIN or a separate supply voltage. The NCP3121 will enable itself once VIN crosses the input UVLO threshold.

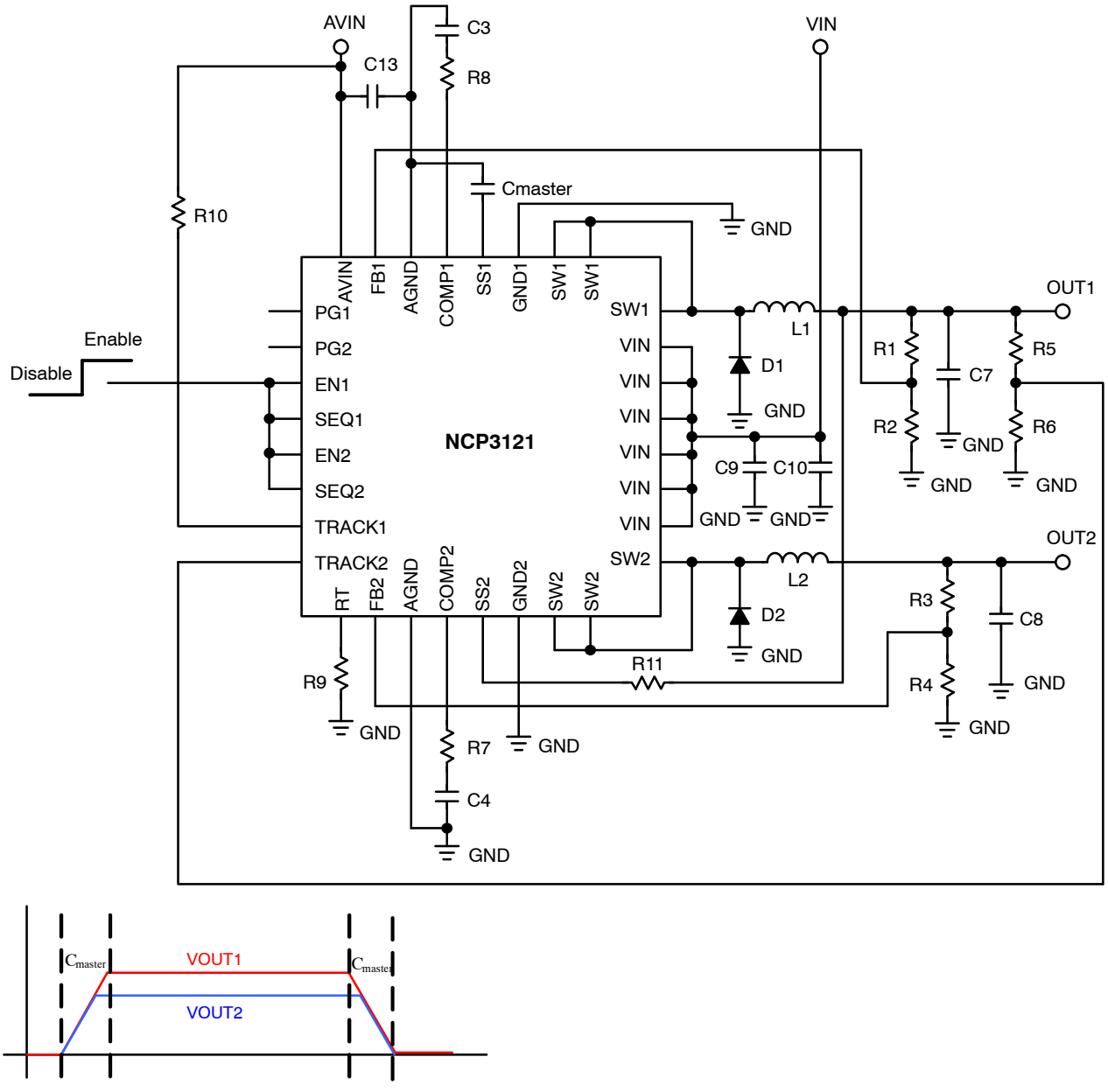

Figure 52. Tracking Configuration