AND8489/D

A Simple 12 V_{out}, 22 W, Off-line Forward Converter Using ON Semiconductor's NCP1027/1028 Monolithic Switcher

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APPLICATION NOTE

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Introduction

Most power supplies with less than 100 W output utilize a flyback switching topology due to the simplicity and low cost of the circuit implementation. As with any power topology there are trade−offs which usually involve circuit simplicity versus performance and cost. In the case of the forward converter topology, the trade−off usually involves the addition of a freewheeling diode and output choke. It should be noted, however, that this "addition", depending on the source and cost of these extra components, can actually be a "wash" due to the increased output capacitance generally required for a flyback topology. Flybacks usually require multiple low impedance output capacitors and even a small inductor in a pi−filter to minimize output ripple. Since the forward converter utilizes a choke−capacitor output filter, the output capacitor requirements are minimized due to the output choke being the main filtering element. In addition, the peak−to−average switching currents in the forward converter can be almost half of that of an equivalent flyback design which lowers EMI generation and allows the use of a lower current rated MOSFET and output rectifier in the overall converter design.

The low power forward converter in this application note is intended for use in white goods, E−meters, and low power communications equipment where low EMI generation and high efficiency are required. This particular example provides 12 V at up to 2 A peak output current with an efficiency of greater than 80% for typical loads over the entire universal ac input range (90 to 265 Vac).

General Specifications

Topology: Single Switch Forward Converter
Input: 90 to 270 Vac (universal input)
Output: 12 V at 2 A max. (22 W continuous)
Output Ripple: less than 100 mV peak−to−peak at full load
Combined Load/Line Regulation: ±2%
Efficiency: 80% minimum from half to full load
Conducted EMI Compliance: EN55022 Level B (average)
Over−temperature and overcurrent protection

Circuit Operation

The ON Semiconductor NCP1027/1028 series of monolithic switcher is implemented as a single switch forward converter using a resistor−capacitor−diode (RCD) reset scheme which allows for a very simple transformer design (no reset winding) and a maximum duty cycle of up to 80% at minimum input line. An off−the−shelf slug type inductor is utilized for the output choke and minimal output capacitance is required for less than 100 mV of output ripple. The schematic of the forward converter circuit is shown in Figure 1. A conducted EMI input filter is comprised of C1, C2, C3, C10, L1 and L2. L1 and C10 comprise the common mode filter while the remaining components form the differential mode filter. R1 serves as an optional inrush limiter during initial power supply turn−on when C3 and C4 are discharged. The resistor should be a wire wound or a similar construction that can tolerate the high joule surge rating that will be present at initial supply turn−on. Metal film resistors are not recommended due to eventual transient stress fatigue.

The primary control chip can be either the NCP1028 or NCP1027. The 1027 version has a built in OVP sensor on the V_{CC} pin that detects if the chip's operating V_{CC} becomes excessive and will latch the controller off under such a condition. The 1028 does not have this feature and was used in this design because the V_{CC} for the chip is derived by a simple peak detector circuit composed of D7, R8 and C13 which is driven by the auxiliary winding on T1. R8 limits the maximum current to the chip while C9 is the main V_{CC} filter capacitor. Since the peak value of the "raw" V_{CC} voltage on C13 varies with input line, an output OVP function is not usable here, hence, use of the NCP1028. The NCP1027 could be used if indirect output OVP sensing were desired by changing the auxiliary peak detector circuit to a forward converter type rectifier with L/C filter network (integrator) by adding a freewheeling diode after D7 and inserting L4, a 15 mH to 25 mH choke (RF type choke) between the cathode of D7 and C13. The voltage on C13 would then be regulated against line changes and would provide a representative dc analog of the main output voltage. R8

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would then have to be adjusted to provide the proper current into U1's V_{CC} pin such that the OVP trip level could be properly calibrated. The schematic of the V_{CC}/OVP sense configuration using the NCP1027 is shown in Figure 2. For more information on the internal functioning of OVP sensing in the NCP1027 please see the device data sheet at [http://www.onsemi.com/pub_link/Collateral/NCP1027−D.PDF](http://www.onsemi.com/pub_link/Collateral/NCP1027-D.PDF).

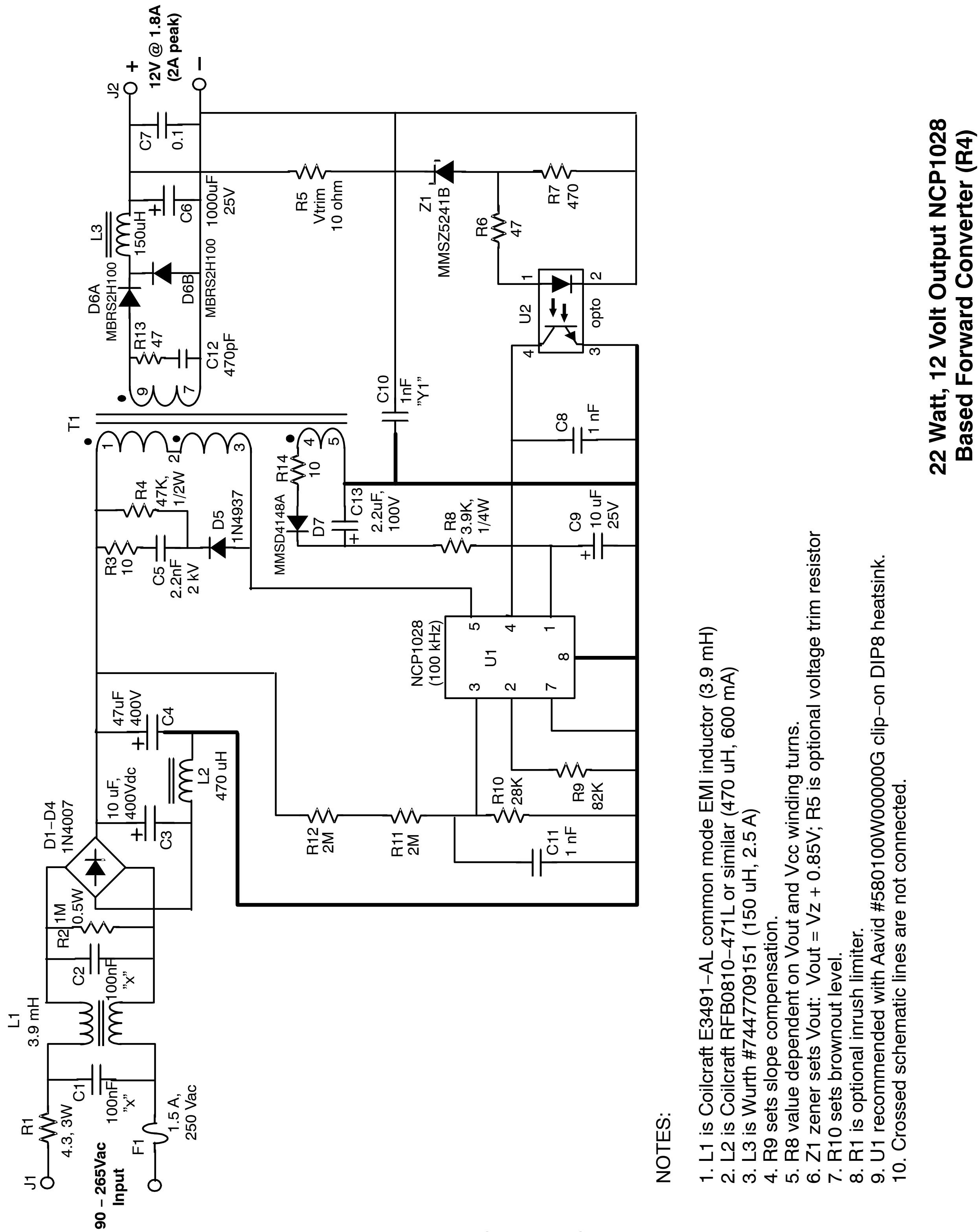

Figure 1. Forward Converter Schematic

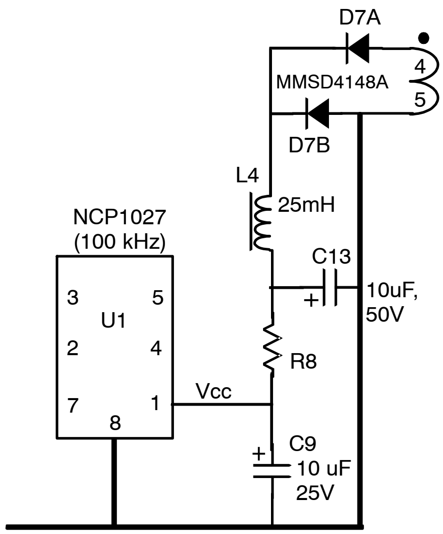

Figure 2. NCP1027 Implementation with Primary V_{CC} OVP Sensing

The network of R10, R11, R12, and C11 provides for brown−out sensing of the rectified mains by monitoring the level of the dc bulk voltage. The level is set so that the chip shuts down when the mains is at approximately 75 Vac. The trip level threshold can be easily adjusted with R10.

Since the NCP1028/1027 controllers utilizes current mode control, and the duty ratio of the converter can exceed 50%, it is necessary to provide slope compensation to the internal current sensing to avoid sub−harmonic instabilities if and when the duty ratio exceeds 50%. This compensation is provided by R9 and is adjustable depending on the level of reflected inductor magnetizing current seen by the controller's current sense circuit.

The output rectifier/filter stage is a conventional forward converter rectifier/freewheel diode implementation consisting of dual diode D6, output choke L3, and main output capacitor C6. Snubber network C12 and R13 attenuates noise and switching spikes on D6 while capacitor C7 provides additional high frequency output noise filtering.

Output voltage sensing and feedback is accomplished via zener diode Z1 and optocoupler U2 and the associated circuitry. When the output voltage exceeds the zener voltage (plus Vf of the optocoupler photo diode) the opto turns on and establishes control of the feedback pin (pin 4) of U1. R7 is necessary to provide a minimum zener current to avoid poor regulation effects that could occur if the zener were operated near the "knee" of its associated transfer curve. Despite the simplicity, this sensing scheme allows for adequate line and load regulation and a reasonable unity gain bandwidth for sufficient 120 Hz output ripple attenuation. The output voltage can be adjusted by changing the value of Z1. For 12 V output, a 11 V zener (MMSZ5241B) is used for Z1, and if necessary, resistor R5 can be adjusted until the desired output voltage is reached. Temperature and set−point variation due to zener and optocoupler tolerances are typically less than ±2%. Capacitor C8 helps to filter noise on the feedback pin and stabilize the overall feedback loop.

Overcurrent and over−temperature protection is inherent in the NCP1028/1027 monolithic controllers. The current limit level is set by the MOSFET's peak current sensing level of the current mode control circuitry. This will occur at approximately 2.3 A of output current for this design and will result in a "hiccup" type of start−stop overcurrent limiting.

Magnetics Design

As with most switching power supplies, the key to effective performance is the design of the power transformer. In this application it was desirable to have a small, yet efficient transformer similar to what a similar flyback topology would require, but without the reset winding that single−switch forward converters usually employ. For this low of power level, a resistor−capacitor−diode (RCD) type of snubber reset scheme was chosen for its simplicity and for the fact that it will allow the converter duty ratio to exceed 50% which helps with transformer core utilization. Of course there is a price to pay and that is allowing U1's internal MOSFET drain voltage to swing up high enough for full core reset, but not sufficiently high as to damage the device under high line transient conditions. A small E25/13/7 type ferrite core (also known as an EF25) was chosen for minimal space. Using the standard transformer design equation to calculate the required primary turns at minimum dc bulk voltage yields:

$$
\begin{aligned}
Np &= \frac{\left(Vpx10^{8}\right)}{(f \times B_{max} \times Ae)} \\
&= \frac{\left(120\ V \times 10^{8}\right)}{\left(100\ kHz \times 2300\ gauss \times 0.52\ cm^{2}\right)} \qquad \text{(eq. 1)} \\
&= 100
\end{aligned}
$$

Where Vp is minimum bulk voltage, f is the switching frequency, B_{max} is a reasonable maximum flux density in the ferrite core, and Ae is the cross sectional area of the core. Checking the available core bobbin width (~0.5 inches or 12.7 mm) indicates that approximately 25 turns of #28 AWG magnet wire will occupy one full winding layer comfortably, so four layers will be required for the full primary. To minimize leakage inductance, we can wind half of the primary (50 turns over two layers) first, then the secondary and auxiliary V_{CC} windings, and finally the second half of the primary. By "sandwiching" the secondary between the primary halves, the primary−to−secondary leakage will be minimized. The 12 V secondary turns are given by:

$$Ns = Np \times \frac{(V_s + Dvf)}{(Vp \times D_{max})}$$

$$= 100 \times \frac{(12\ V + 1\ V)}{(100 \times 0.8)} \qquad \text{(eq. 2)}$$

$$= 16$$

Where Vs is the main secondary voltage; Dvf is the forward drop of diode D6, and Dmax is the maximum duty ratio of the converter. From wire tables it appears that up to 22 turns of #26 AWG magnet wire (adequate of the average secondary current) will fit across one layer of the bobbin. In order to allow for tolerances in the chip's duty ratio and efficiency issues, 20 turns were selected for a margin allowance and to allow end−cuffing of the secondary winding with Mylar tape for safety insulation issues. 18 turns would have probably also been an acceptable compromise if more margin were desired.

Since the V_{CC} will be derived from peak charging of C13, and the V_{CC} in U1 is clamped at approximately 9 V with an internal zener, the V_{CC} winding turns were ratioed from the primary to yield approximately 15 V when the line bulk is at the minimum level (120 Vdc).

$$Ns(V_{CC}) = \left(V_{aux} + \frac{V_{D7}}{V_{bulk\,min}}\right) \times 100$$

$$= \left[\frac{(15 + 1)}{120}\right] \times 100 \qquad \text{(eq. 3)}$$

$$= 13.3{>}{>}14 \text{ turns}$$

Assuming a max chip V_{CC} current drain of 1.5 mA, the approximate value of R8, the V_{CC} current limiting resistor, can be calculated: (15 V – 9 V)/0.0015 A = 4.0 kΩ >> use 3.9k.

One other consideration that should be checked is the peak inverter MOSFET current to make sure it is within the proper current limit ratings of the NCP1028. Since the circuit is operating in continuous conduction mode with respect to the output choke, the peak primary current will be the sum of three components; the peak load current, the choke magnetizing current (both reflected through the transformer turns ratio) and the primary magnetizing current of the transformer. The magnetizing current represents stored energy in both magnetic elements and should ideally be minimized. Unfortunately this would mean a high as possible inductance for both which is not practical. This leads us to the output choke design. Assuming a maximum peak output current of 2 A, the minimum dc current, or critical current in the choke is typically chosen to be 10% to 20% of the max load current. For this case we choose 15% which is 0.3 A. This current is the point at light output load where the choke current just becomes discontinuous. Using the V = L x dI/dt relationship we can rearrange to find L: L = V x (dt/dI) where V is the voltage across the choke at low line (calculated using the transformer turns ratio of 5:1 and V_{out}); dt is the maximum on−time of the MOSFET at 100 kHz (8 μs); and dI is the choke ripple current which is <u>twice</u> the chosen critical choke dc current level (0.6 A).

$$L = \left[\left(\frac{120\ V}{5}\right) - 12\ V\right] \times \left(\frac{8\ \mu s}{0.6\ A}\right) = 160\ \mu H \qquad \text{(eq. 4)}$$

>> choose 180 μH, a common value.

A 180 μH, 3 A rated off−the−shelf ferrite inductor from Coilcraft was chosen for the output inductor.

Now we can calculate the various MOSFET/T1 primary current components at minimum line:

- Reflected primary dc load current: 2 A/5 = 400 mA (I_{out}/turns ratio)
- Reflected choke magnetizing current: 0.6 A/5 = 120 mA
- Transformer magnetizing current: This involves knowing the transformer's primary L:

$$Lp = \text{core AL value} \times Np^2 \times 10^{-3} = 1500 \times (100)^2 \times 10^{-3} = 15{,}000\ \mu H \text{ or } 15\ mH \qquad \text{(eq. 5)}$$

This will typically be slightly less due to micro gaps between the core halves, so choose 12 mH.

Where AL for E25/13/7 core of 3C90 material is 1500 per manufacturer core specs.

So dI = (Vp x dt)/L yields: (120 V x 8 μS)/12,000 μH = 0.080 A or 80 mA.

The peak MOSFET/T1 primary current is then:

$$400\ mA + 120\ mA + 80\ mA = 600\ mA \qquad \text{(eq. 6)}$$

This value is well within the rated 720 mA minimum peak current limit level of the NCP1028. Assuming a nominal duty ratio (D) of 0.55 at 120 Vac input, then the rms primary current will be the square root of D times the peak current = 0.74 x 600 = 0.44 A. The original selection of #28 AWG primary wire is sufficient to handle this current with acceptable temperature rise.

Likewise the T1 secondary rms current for 120 Vac input and 1.8 A output can also be calculated: 0.74 x 1.8 A = 1.33 A which can also be handled with an acceptable temperature rise using the #26 AWG wire selected for the 12 V secondary.

Figure 3 shows the final design of the forward converter transformer T1.

The components for the RCD reset snubber (R4, C5, D5) were selected based on U1's internal MOSFET drain voltage waveforms during worst case line and load conditions and the primary inductance of T1. Since C5 will resonate with T1's primary inductance after MOSFET turn−off, the value of C5 should be selected such that the resonant frequency of this L/C network is less than half of the inverter switching frequency. In this case T1 has a primary inductance of about 12 mH, so with C5 at 2.2 nF, this would result in a resonant frequency of about 31 kHz. Keeping this resonant frequency low will minimize resonant voltage peaking of the drain waveform at turn−off. R4 discharges C5 during each switching cycle so the capacitor can control the voltage level across T1's primary so as to adequately allow the proper volt−second reset conditions for the transformer primary. Waveforms of the drain voltage and associated primary current are shown in Figures 4 and 5 for 120 Vac and 230 Vac inputs, respectively, with an output current of 1.75 A. The peak voltages are well within the voltage rating of the NCP1028 MOSFET drain, and are, in fact, lower than what would exist if a conventional reset winding were used on the transformer. The voltage margins were also completely adequate at a high line of 265 Vac input. In addition, the quasi−resonant effect of the snubber capacitor C5 interacting with T1's primary inductance helps to shape the drain waveform so as to minimize leakage inductance ringing and spikes, thus reducing EMI issues. This is clearly demonstrated in the conducted EMI profile (Level B) shown in Figure 8. This was taken with a load of 1.8 A (22 W).

The output voltage ripple for a 1.75 A load is shown in Figure 6, with a scale of 100 mV per division vertical.

The efficiency versus load curves are shown in Figure 7 for both 120 and 230 Vac input. Note that the supply efficiency is at or above 80% from half to full load.

MAGNETICS DESIGN DATA SHEET

Project / Customer: ON Semiconductor – 24 watt, 12 vout NCP1028 Fwd Conv

Part Description: 24 watt NCP1027 resonant reset forward conv. xfmr (Rev 3)

Schematic ID: T1

Core Type: EF25 (E25/13/7); 3C90 material or similar

Core Gap: No gap

Inductance: (Primary) 12 mH minimum

Bobbin Type: 10 pin horizontal mount for EF25

Windings (in order):

Winding # / type	Turns / Material / Gauge / Insulation Data
Primary A (1 – 2)	50T of #28HN over 2 layers (25 TPL). Insulate for 1 kV to next winding. Self leads to pins.
Vcc (4 – 5)	14 turns of #28 HN over 1 layer, close wound and centered in window. Self leads to pins. Insulate to 3 kV to next winding
12V Secondary (9 – 7)	20 turns of #26 triple insulated wire over one layer. Self leads to pins.
Primary B (2 – 3)	Same as Primary A. Insulate with tape and self–leads to pins.

Hipot: 3 kV from primaries & Vcc to secondary for 1 minute.

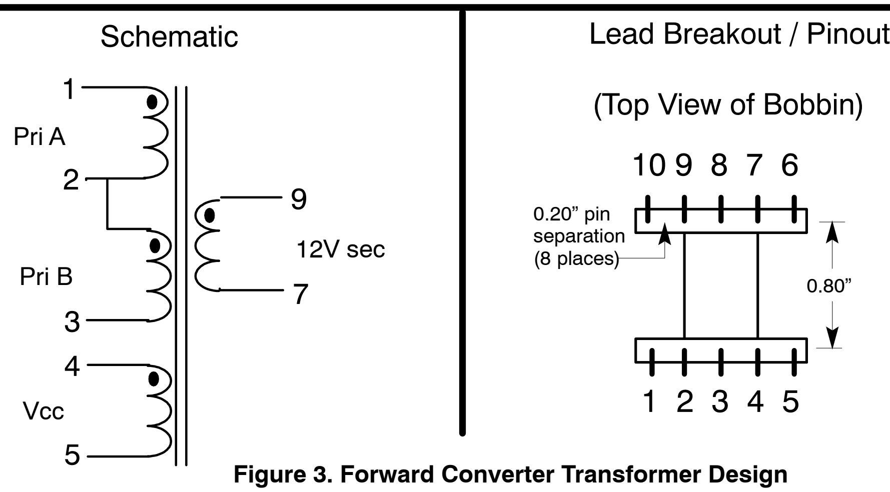

Figure 3. Forward Converter Transformer Design

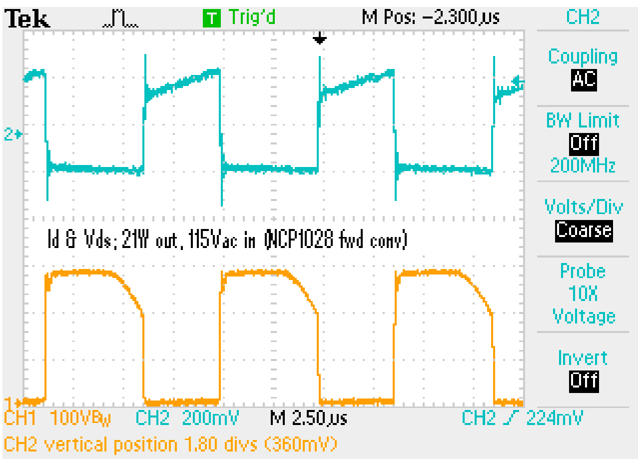

Figure 4. MOSFET Drain Voltage and Current at 120 Vac (1.75 A Load)

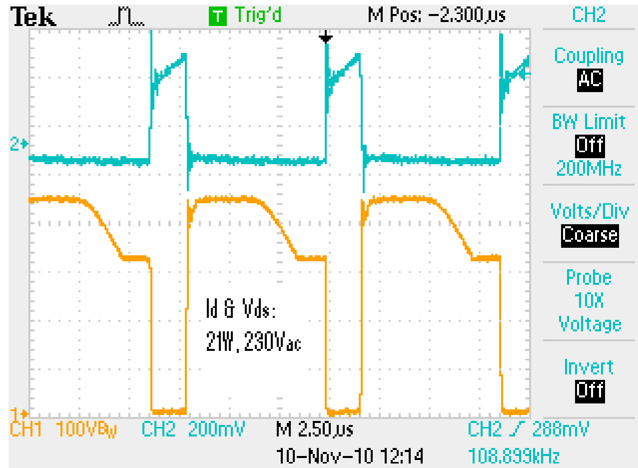

Figure 5. MOSFET Drain Voltage and Current at 230 Vac (1.75 A Load)

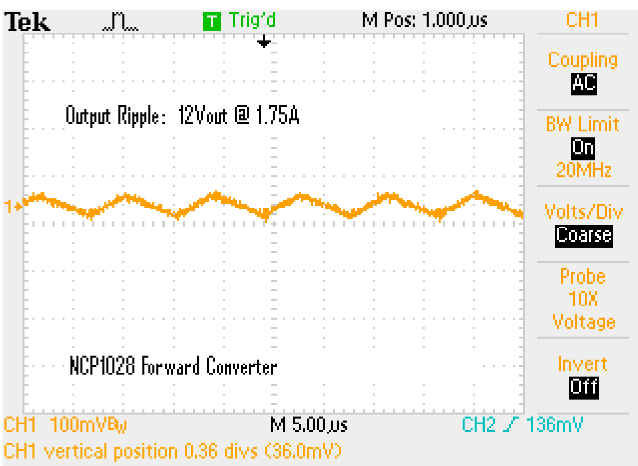

Figure 6. Output Ripple

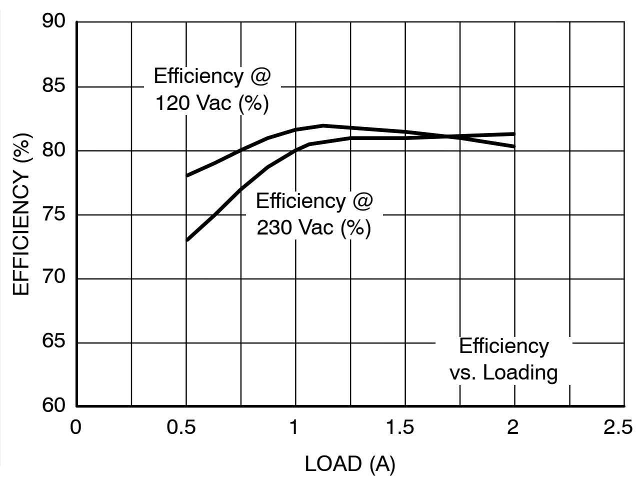

Figure 7. Efficiency versus Load Curves

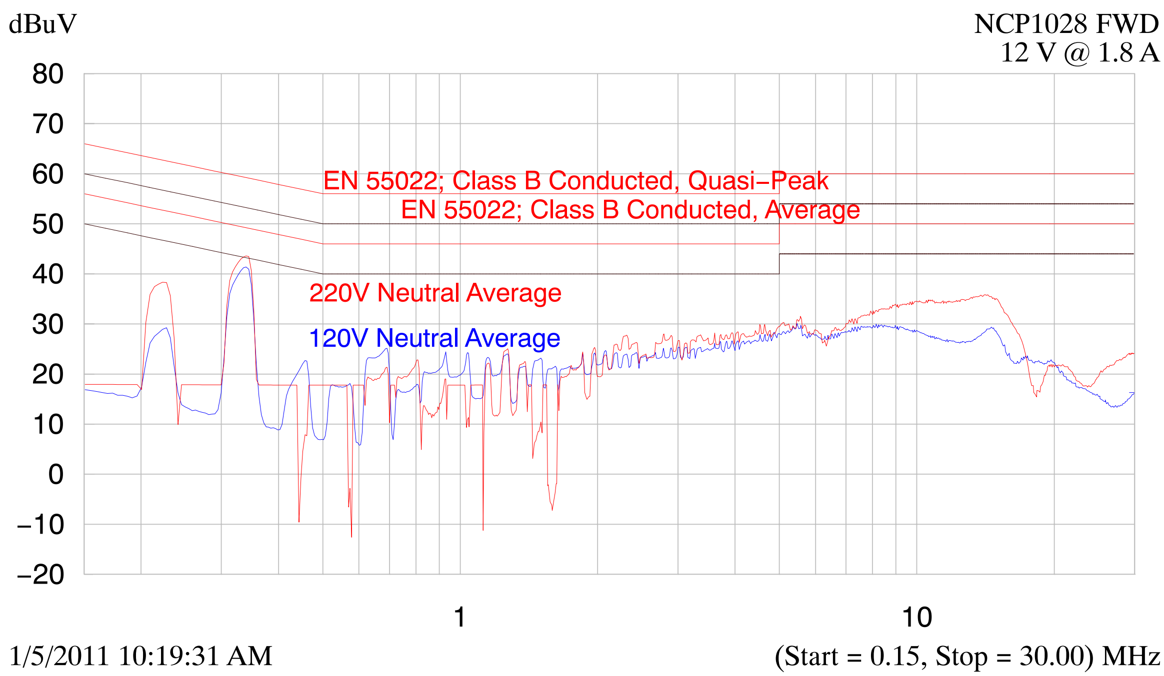

Figure 8. Average Conducted EMI Profile at 120 Vac (blue) and 220 Vac (red) 22 W Output (Note: lower dashed line is 6 dB margin level for Class B)

BILL OF MATERIALS FOR 12 VOUT, 22 W, NCP1027/1028 FORWARD CONVERTER

Designator	Qty	Description	Value	Tolerance	Footprint	Manufacturer	Manufacturer Part Number	Substitution Allowed
D6A, D6B	2	Schottky diode	2 A, 100 V		SMB	ON Semiconductor	MBRS2H100T3G	No
D1, 2, 3, 4	4	Diode − 60 Hz,	1 A, 800 V		SMA	ON Semiconductor	MRA4007	No
D5	1	Diode − fast recov	1 A, 600 V		axial lead	ON Semiconductor	1N4937	No
D7	1	Signal diode	100 mA, 100 V		SOD−123	ON Semiconductor	MMSD4148A	No
Z1	1	Zener diode	12 V, 500 mA		SOD−123	ON Semiconductor	MMSZ5241B	No
U2	1	Optocoupler	CTR >/ = 0.5		DIP4 SMD	Vishay or NEC	SFH6156A−4 or PS2561L−1	Yes
U1	1	Monolithic Controller	100 kHz		DIP8	ON Semiconductor	NCP1027 or NCP1028	No
C1, C2	2	”X” cap, box type	100 nF, X2		LS = 15 mm	Rifa, Wima	TBD	
C10	1	”Y1” cap, disc type	1 nF, Y1		LS = 7.5 mm	Rifa, Wima	TBD	
C5	1	Ceramic cap, disc	2.2 nF, 2 kV	5%	LS = 7.5 mm	Rifa, Wima	TBD	
C8, C11	2	Ceramic cap, monolythic	1 nF, 50 V	10%	1206	AVX, Murata	TBD	
C7	1	Ceramic cap, monolythic	100 nF, 50 V	10%	1206	AVX, Murata	TBD	
C12	1	Ceramic cap, monolythic	470 pF, 200 V	5%	1206	AVX, Murata	TBD	
C3	1	Electrolytic cap	10 μF, 400 Vdc	10%	LS = 5 mm, D = 12.5 mm	UCC, Panasonic	TBD	
C4	1	Electrolytic cap	47 μF, 400 V	10%	LS = 7.5 mm, D = 16 mm	UCC, Panasonic	TBD	
C9	1	Electrolytic cap	10 μF, 25 Vdc	10%	LS = 2.5 mm, D = 6.3 mm	UCC, Panasonic	TBD	

BILL OF MATERIALS FOR 12 VOUT, 22 W, NCP1027/1028 FORWARD CONVERTER

Desig-nator	Qty	Description	Value	Tolerance	Footprint	Manufacturer	Manufacturer Part Number	Substitution Allowed
C13	1	Electrolytic cap	2.2 µF, 100 V	10%	LS = 2.5 mm, D = 6.3 mm	UCC, Panasonic	TBD	
C6	1	Electrolytic cap	1000 µF, 25 V	10%	LS = 5 mm, D = 12.5 mm	UCC, Panasonic	TBD	
R1	1	Resistor, 3W, Wire wound	4.4 Ω, 3 W	10%	LS = 7.5 mm, D = 7 mm	Ohmite, Dale	TBD	
R2	1	Resistor, 1/2W, metal film	1 Meg, 1/2W	10%	Axial lead; LS=12.5mm	Ohmite, Dale	TBD	
R4	1	Resistor, 1/2W metal film	47k, 1/2W	10%	Axial lead; LS=12.5mm	Ohmite, Dale	TBD	
R3, R14	2	Resistor, 1/4W SMD	10 Ω	5%	SMD 1206	AVX, Vishay, Dale	TBD	
R6, R13	2	Resistor, 1/4W SMD	47 Ω	5%	SMD 1206	AVX, Vishay, Dale	TBD	
R5	1	Resistor, 1/4W SMD	TBD (10 Ω)	5%	SMD 1206	AVX, Vishay, Dale	TBD	
R7	1	Resistor, 1/4W SMD	470 Ω	5%	SMD 1206	AVX, Vishay, Dale	TBD	
R11, R12	2	Resistor, 1/4W SMD	2 MΩ	5%	SMD 1206	AVX, Vishay, Dale	TBD	
R10	1	Resistor, 1/4W SMD	28k	5%	SMD 1206	AVX, Vishay, Dale	TBD	
R9	1	Resistor, 1/4W SMD	82k	5%	SMD 1206	AVX, Vishay, Dale	TBD	
R8	1	Resistor, 1/4W SMD	3.9k	5%	SMD 1206	AVX, Vishay, Dale	TBD	
F1	1	Fuse, TR−5 style	1.5 A		TR−5, LS = 5 mm	Minifuse		
	1	Heatsink for U1			DIP8 clip−on	Aavid	Aavid 580100W00000G	
L3	1	Inductor (output choke)	150 µH, 2.5 A	5%	1210 SMD (12 x 12 mm)	Wurth	7447709151	
		(alternate)	180 µH, 2.2 A	5%	Axial lead choke (1.1”)	Coilcraft	PCH−45X−184LT	
L2	1	Inductor (EMI choke)	470 µH, 600 mA		See Wurth Drawing	Wurth Magnetics	744772471	
L1	1	EMI Inductor	3.9 mH,		See Coilcraft Drawing	Coilcraft	E3491−AL	
T1	1	Transformer	E25/13/7 core		See Mag Drawing	Wurth Magnetics	750312228	
J1, J2	2	Screw Terminal			LS = 0.2”	DigiKey	# 281−1435−ND	

References:

NCP1028 Data Sheet: http://www.onsemi.com/pub_link/Collateral/NCP1028−D.PDF

NCP1027/NCP1028 Application Notes, Design Notes and Reference Designs:

1. http://www.onsemi.com/PowerSolutions/supportDoc.do?type=AppNotes&rpn=NCP1028
2. http://www.onsemi.com/PowerSolutions/supportDoc.do?type=AppNotes&rpn=NCP1027
3. http://www.onsemi.com/PowerSolutions/supportDoc.do?type=Reference Designs&rpn=NCP1027
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