Component Design for LV2 Power Electronics (Except Main Battery?)

Fully discharge Li+ cells are 2.5V (not 2V). (2.5, 3.6, 4.2)

Design Review notes

Purpose of late-phase design review

Encourage early design review
Make the designer reconsider the design
Catch bugs / mis-features
Find holes
Fix interface with other systems
Discuss bloat (possibly even eliminate it)
Disseminate the design

■ Probably don't

Check component values (except possibly the few very critical ones, or in case of error)
Eliminate marginal mis-features

Recovery Goals

Safety
Reliable recovery (minimize point failure, provide redundancy, verifiable status)
Low maintenance / Reasonable turn-around times
Maximum independence from FC / main avionics
Low mass
Small size

Specs

one hour independent battery power at 100mA
Main power from a 4-cell Li+ battery or umbilical
integrated 2m DTMF receiver
redundant igniter circuits
CAN/PIC node

■ Vision thing

Battery charge when powered (self initiated?)
Battery monitor
Send shutdown message?
Send wake up message
Implement backup timer?

CAN Node Switch Mode Power Supply (SPS) (200)

SPS Goals

Reliable / Robust power
Low mass
Small size
Good efficiency (>85%)
Low generated RFI
Low cost

SPS Specs

Power from a 4-cell Li+ battery or umbilical

SPS Design notes

The SPS is intended to be implemented on each CAN node. It's design should therefore be reasonably cheap and compact.

Desire a fairly bullet-proof design. This has led to interference filters (common mode choke + capacitors), fuse, transient voltage suppressor, over-voltage crowbar, and an extra current limiting circuit.

The SPS accepts two digital inputs and produces one digital output. It also provides a "power" LED. The input "SPS_off" allows testing of the HAP while main power is up. It also could be used for testing the reset behavior of a CAN node, or for performing a "hard boot" on a misbehaving node. The SPS_off capability can be eliminated from a node by removing a few components without otherwise affecting the node's operation.

The SPS outputs "SPS_Fail" from a CMOS comparator when the output voltage falls below a threshold. This signal is used by the HAP for battery switching. It can also be routed to the PIC.

For small size and high efficiency an approximately 1.5MHz switching supply is used. The LT1767 was chosen because it's the only part that can stand the 18V mains voltage present when on the umbilical. For interference reasons the switching supply has been synchronized to the PIC's crystal oscillator, the second digital input "sync" is used for this function.

■ Issues

The feedback network uses 4 resistors. This is to provide taps for the CMOS comparators that generate the SPS_Fail and over-voltage protection signals. It would be better from a transient response and layout viewpoint to split this into two dividers (thus adding a resistor), however this would increase the divider current drain slightly. It's probably worth it.

Connector(s)

■ CM200 No JST-08PS-JED

JST <www.jst−mfg.com>08PS−JED,right angle male (plug, side entry)

1.25mm	pitch
8	circuits, double row
1A	current rating / pin
50V AC/DC	voltage rating
500VAC	withstanding voltage, 1 minute
(20, 40)mΩ	contact resistance Max (initial, after environmental testing)
300MΩ	insulation resistance

(4.6 x 5.95 x 6.15)mm HxWxD

Glass filled nylon 66, Brass w/copper undercoat, tin plated

The connector should be small, reliable, polarized, and reasonably durable. The current goal is 1A per connection. 4 circuits are required. The board connector should be a right-angle type to allow plugging while using stacked boards. The cable connector should be straight-in. Adequate stress relief is a must.

We decided to use the JST JED series for its small size and decent "feel". The connector is not mechanically robust enough for us, so we’ve decided to use extra mounting hardware to stress relieve the connector, with this change all our goals appear to be met. The 8 circuit connector was chosen allowing us to double up for redundancy.

Integrated Circuit(s)

■ U200 LT1767, 1.25+MHz, 1.5A, DC-DC buck converter (DigiKey LT1767EMS8−ND 6.00$ea)

The selection criterion include size, efficiency, complexity and frequency. This MSOP−8package can reach 87% efficiency supplying 5V @ around 100mA output @ 1.5MHz (synchronized). With 10μA shutdown current, about 1mA quiescent current, and potentially 1A output current while tolerating an 8-20V input range. (I think the 8V is based on 2V/cell, it shouldn’t get that low.)

■ U201 LTC1442, ultra low power dual comparator w/ reference (DigiKey LTC1442IS8−ND 4.88$ea)

Functionally this chip identifies over-voltage and under-voltage conditions. Since it must operate constantly off the Li+ backup battery, micropower is essential. This unit is a single package SO−8or MSOP−8solution which draws only 3.5μA (typ) or 5.7μA (Max).

■ U202 Any single ’HC NOR gate (DigiKey SN74AHC1G02DBVR 0.56$ea) SOT−23−5

This gate blocks the synchronization signal during power fail. This is required because the backup power can keep the synchronization signal active even when U200 is shut down. If the synchronization signal passed directly to U200, U200 would be powered through its input protection diodes during positive excursions of the sync signal. U202 is not required on SPS’ which do not include backup power.

Since a fully discharged Li+ battery is only 2.5V the logic should run in the range of about 2V to 5.2V. Most current CMOS products will do this.

■ U250 4 Bit ’HC binary counter (DigiKey 296−12344−1−ND0.88$ea) TSSOP−16

This is the SN74LV161A (25+MHz)

■ U251, 252 Any single 'HC ExOR gate (DigiKey SN74AHC1G86DBVR 0.56$ea) SOT−23−5

Transistor(s)

■ Q200   IRF7455 SO-8 N-FET (DigiKey IR7455−ND1.53$ea)

International Rectifier IRF7455	
SO−8	package
30V	V_DSS
29mV/C	Break down temp.co. 25C, I_D==1mA
7.5mΩ	R_DS(on)
(, 37, 56)nC	total gate charge @ V_GS==5V
3480pF	typ. input capacitance
15	I_DMax (dissipation limited)
120A	Max pulse current
±12V	V_GSMax
1.2V	Max body diode forward voltage @ 25C, 2.5A, V_GS==0V
96ns	Max body diode reverse recovery time
150nC	Max body diode reverse recovery charge
200mJ	Max pulse avalanche Energy
0.25mJ	Repetitive avalanche Energy
150C	T_JMax
50C/W	Max θ_JA on standard 1"sq. copper clad board, t < 1s.

Crowbar element for CAN node. Q200 shorts the incoming power bus when an over-voltage condition exists at the output of the SPS. Dissipation is not high because of the short duration of activation. Must respond to 5V gate drive and sink minimum of 8A at 5V. Voltage rating 25V minimum.

A very cheap part in the SO−8package is the IRF7463, R_DSON==8mΩ, VDS_Max==30V, 1.55$.
Almost as suitable in the MSOP−8style, IRF7603, R_DSON==35mΩ, VDS_Max==30V, 1.92$.
Perhaps smallest, but pushing it, SOT−223,IRLL014, R_DSON==140mΩ, VDS_Max==55V, 0.68$

The SOT-223 package is spec'd (See irll014n.pdf p.5) as having 10% response to a single 100ms pulse. The effective thermal resistance is therefore at most 12C/W. assuming a16A pulse, the max power is about 36W or 430C temperature rise. Obviously too much. What is the impedance of a 4AHr 4 cell Li+ battery? Ans: about 0.1Ω.

Figuring 0.1Ω per cell, and accounting for the transistor at 0.28Ω and the choke at 0.14Ω the short circuit current of a full battery pack is

```
4.2 * 4 / (4 * 0.1 + 0.28 + 0.14)
```

```
20.4878
```

Safety suggests figuring 30A. In which case the SOT-223 is too small.

Going to the MSOP-8? Ooo the Fairchild NDT455N is much better! (R_DS(on)==20mΩ), but still too small.

Even the MSOP−8is too small … sigh. So it's the SO−8for us.

Ok, just to disprove insanity, here's the justification. Surge current 30A for 50ms Max.

R_DS(on) @ 150C is < 12mΩ, so P_D(Max) is < 11W, transient thermal resistance (equivalent, see irf7455.pdf p.5) is < 7C/W so temperature rise is < 77C. Therefore maximum operating ambient is > 73C, which is pretty damn hot, so there.

■ Q201 ZXMN2A01F SOT-23-3 N-FET (DigiKey ZXMN2A01FCT−ND0.48$ea@10) (Changed to bipolar)

Zetex ZXMN2A01F
SOT-23-3 package
20V V_DSS
120mΩ R_DS(on) V_GS==4.5V, 25C
3.1nC typ. total gate charge @ V_GS==4.5V
299pF typ. input capacitance
1.8 I_DMax (dissipation limited)
10A Max pulse current
±12V V_GSMax
(,.84, .95)V Max body diode forward voltage @ 25C, 0.6A, V_GS==0V
11.2ns typ. body diode reverse recovery time
3.64nC typ. body diode reverse recovery charge
(−55, 150)C T_J range
200C/W Max θ_JA on FR-4, t < 5s.

The problem with the micro-6 & Fairchild part is the cases have mutually incompatible board layout between each other and between the std. SOT-23. DigiKey has a decent SOT-23-3 (4.2A) but only in tape&reel :(.

Oooo, compare Fairchild NDC651 (maybe better)

International Rectifier IRLMS1503
Micro6 package (Similar to SOT-23-6)
30V V_DSS
37mV/C Break down temp.co. 25C, I_D==1mA
200mΩ R_DS(on) V_GS==4.5V, 25C
(, 6.4, 9.6)nC total gate charge @ V_GS==5V
210pF typ. input capacitance
2.6 I_DMax (dissipation limited)
18A Max pulse current
±20V V_GSMax
1.2V Max body diode forward voltage @ 25C, 2.2A, V_GS==0V
54ns Max body diode reverse recovery time
58nC Max body diode reverse recovery charge
(−55, 150)C T_J range
75C/W Max θ_JA on FR-4, t < 5s.

Q201 has been changed to bipolar PNP because it turns out there is an O.C. pin on the PIC, and the bipolar works better with the O.C. (The bipolar is generic 3906-ish.)

Q201 isolates the HAP-powered PIC from the possibly shutdown U200. This can be a generic logic level mosfet. Prefer SOT-23-3 package. Best if Voltage ≥20V.

Couldn’t find a DigiKey part with cool ratings and SOT23-3. IRLMS1503 is in a Micro6. I think a PCB can be laid out for either package.

■ Q202 See Q201

Q202 performs a switching function similar to Q201, see Q201 for details.

Diode(s)

■ CR200 Panasonic Schottky MA2Q705 (MA10705) (DigiKey MA10705CT−ND0.64$ea/10)

Panasonic MA2Q705 (MA10705)
NMiniP2 package, proprietary, but similar to other 2 pin types
30V V_R
1.5 I_DMax (dissipation limited)
30A Max pulse current
0.37V Max forward voltage @ 25C, 1A
50ns Max reverse recovery time
(−40, 125)C T_J range

This is the main diode for the buck topology. The voltage rating requirement is ≥25V.

Average current is given by

```
Id == Io (Vi - Vo)/Vi /. {Io → 1, Vi → 18, Vo → 5} // N
```

$$Id == Io \frac{(Vi - Vo)}{Vi} \text{ /. } \{Io \to 1, Vi \to 18, Vo \to 5\} \text{ // } N$$

```
Id == 0.722222
```

A reasonably conservative choice would be a 1A Schottky diode.

■ CR201 1N4448HWS (DigiKey 1N4448HWSDICT−ND0.30$ea/10)

Diodes Inc. 1N4448HWS
SOD-323 package, 2 pin type possible to shoot a trace
80V V_R
100nA Max reverse current @ V_R @ 25C
50μA Max reverse current @ V_R @ 150C
200mW Max power (assume 25C at leads)
1V Max forward voltage @ 25C, 100mA
4ns Max reverse recovery time
3.5pF Max capacitance
625C/W θ_JA @ 25C
(−65, 150)C T_J range

Considered BAV16WS, but 1N4448HWS has slightly better specs.

Diodes Inc. BAV16WS (1N4148WS)
SOD-323 package, 2 pin type possible to shoot a trace
75V V_R
1μA Max reverse current @ V_R
200mW Max power (assume 25C ???)
1V Max forward voltage @ 25C, 50mA
4ns Max reverse recovery time
2pF Max capacitance
(−40, 125)C T_J range

Also considered the SOD-123 package, which is longer, similar to 1206. Smaller package is nice and according to preliminary analysis, will still give 1206 interchangeability.

Diode provides boost current for U200’s internal switch. Ordinary signal type diodes are adequate. Something we could shoot a lead through is probably good. Voltage rating ≥ 25V.

■ CR202 not used

Formerly this diode prevented reverse current flow from the HAP into the main bus. This is no longer required because of the diode added to the battery backup section of the HAP. Recommend removal of this component. In fact this has been done as of SPS−0.3.

■ D200 BAT42WS, Schottky diode (DigiKey BAT42WSCT−ND0.52$ea/10)

Diodes Inc. BAT42WS
SOD-323 package, 2 pin type possible to shoot a trace
30V V_R
200mA Max continuous forward current
500mA Peak repetitive forward current
4A Peak non-repetitive forward current
500nA Max reverse current @ V_R @ 25C
100μA Max reverse current @ V_R @ 100C
200mW Max power (assume 25C at leads)
(, 0.26, 0.33)V Max forward voltage @ 25C, 2mA
1V Max forward voltage @ 25C, 200mA
5ns Max reverse recovery time
10pF Max capacitance
625C/W θ_JA @ 25C
(−55, 125)C T_J range

D200 is intended for current limiting. When the SPS output voltage is low during start up, loss of bus power, or due to excessive current draw, the SPS_Fail signal causes D200 to pull low via Q202. The idea here is to limit the maximum current available by preventing Vc from rising above a certain value. The difficulty with this idea is that it's uncertain exactly what current limit value should be chosen or exactly what the Vc switching cutoff voltage is.

The datasheet gives typical values as Gm==2.5A/V, Vc @ 0% duty==0.35V, Vc@1.5A switch current==0.9V.

The maximum current from the error amp over temperature is sink(70, 110, 180)μA, source(80, 120, 160)μA

The original concept was to use a Schottky diode to clamp the switch current at about 0.4V or 1/8A. But this is probably too low, and too easily effected by threshold variations. An ordinary Si diode clamping at 0.5-0.65 V probably gives adequate start up current, and better consistency, but the clamp current is so high forgetting the whole thing seems reasonable. Alternately a resistor might be used to give a more linear and consistent current limit.

The overarching question is whether a current limit is really required. That is can the SPS withstand a continuous short on it's output if the internal current limit (1.5, 2, 3)A is used? The answer at this time is maybe. The ratings seem close. What temperature will U200 reach?

For the time being, a Schottky will be selected here because having a signal-level Schottky on hand seems prudent.

■ D201 Hi−brightYellow LED, 0603 (DigiKey 160−1448−1−ND1.65$/10)

LITEON LTST−C191KGKT
0603 package
clear package color
591nm peak λ (yellow)
2V Vf @ 20mA
60mC Peak Brightness
130° viewing angle
AlInGaP LED

LITEON LTST−C191KGKT(2.25$/10)
0603 package
clear package color
574nm peak λ (green/green-yellow)
2V Vf @ 20mA
35mC Peak Brightness
130° viewing angle
AlInGaP LED

D201 is by request. Its purpose is to aid bench testing. It simply lights up when SPS power is present.
Green was back-ordered, so the color was changed to yellow.

We anticipate disconnecting the LED for flight. A cuttable jumper was to be provided to make this easy, but it was eliminated due to space constraints.

Inductor(s)

■ L200 Coiltronics MP2A−1R5,1.5μH, See PM−4112.pdf

Coiltronics MP2A−1R5
1.5 μH nominal inductance
1.54 μH measured @ 100kHz, 0.25V RMS, 0.0A DC, ±20%
73 mΩ DC Ohms @ 20C typ.
2.02 A RMS current producing +40C temperature rise (excluding core loss)
3.22 A Saturation current defined by 30% loss of inductance. Measured at 20C
2.09 V·μs Volt·time product of 300kHz waveform which when applied across inductor produces core loss equal to 10% of power loss producing +40C temperature rise previously determined.
−40to +125C storage temperature range
−40to +85C operating temperature range
molybdenum permalloy core
rated to 500kHz ??? —I don't see any reason for this.
Body Length exclusive of leads 5.88 mm, Length with leads 7.5 mm. Width 5.2 mm, Height 1.8 mm.

Note, some of this design is based on 2V/cell, which is too low. On the other hand, assuming a slightly wider voltage range isn't too harmful, and may even be wise.

Maximum available current == max switch current −p-p inductor current / 2

Derivation (buck converter)

Equal volt-seconds implies

$$Vo == D\,Vi$$

The current increase through the inductor in one cycle (ΔI) is

$$\frac{Vi - Vo}{L} * Ton,\quad Ton == T * D == \frac{1}{f} * D \Rightarrow \Delta I == \frac{1}{f}\,\frac{Vo}{Vi}\,\frac{Vi - Vo}{L}$$

Average current is

$$Io == Ip - \Delta I / 2 == Ip - \frac{Vo\,(Vi - Vo)}{L\,f\,Vi} \Big/ 2 == Ip - \frac{Vo\,(Vi - Vo)}{2\,L\,f\,Vi}$$

The peak current is limited by the IC to 3A Max. Inductor saturation will not destroy the IC due to this current limit.

Inductor parameters

L	inductance
μ	core permeability
ℓ	magnetic path length
A	cross sectional area
g	air gap thickness (maybe)
B	magnetic field strength (Max BM (saturation))
N	number of turns
I	winding current (Peak Ip, Avg. Iavg)
Ip	peak winding current
ΔI	ripple current (p-p current change, current increase during on time)
E	energy stored
Vc	volume of core

Inductor formula [A and ℓ in cm]

$$L == 0.4\,\pi \times 10^{-8}\,\frac{\mu\,A\,N^2}{\ell + \mu\,g}\,[H]$$

$$B == 0.4\,\pi\,\frac{\mu\,I\,N}{\ell + \mu\,g}\,[Gauss]$$

$$E == \frac{I^2\,L}{2}\,[J]$$

The core should not saturate. That is B<BM when I==Ip.

Use the inductor formula to eliminate N

$$\text{Solve}\left[L == k\,ee\,\frac{\mu\,A\,N2}{\ell + \mu\,g},\,N2\right][[1,\,1]]\ //\ \text{FullSimplify}$$

$$N2 \to \frac{L\,(\ell + g\,\mu)}{A\,ee\,k\,\mu}$$

$$B^2 == \left(k\,\frac{\mu\,I\,N}{\ell + \mu\,g}\right)^2\ /.\ N \to \sqrt{\frac{L\,(\ell + g\,\mu)}{A\,ee\,k\,\mu}}\ //\ \text{PowerExpand}\ //\ \text{FullSimplify}$$

$$B^2 == \frac{k\,L\,I^2\,\mu}{A\,ee\,\ell + A\,ee\,g\,\mu}$$

Note that

$$Vc == A\,\ell$$

```
Solve[B^2 == k L I^2 μ / (Vc ee + A ee g μ), Vc][[1, 1]] /. L I^2 → 2 E
```

$$Vc \to \frac{-A B^2 \, ee \, g \, \mu + 2 k \, E \, \mu}{B^2 \, ee}$$

The minimum core volume for a given maximum flux B and energy storage E is therefore

$$Vc == \mu \left(2 E \frac{0.4 \pi}{B^2 \times 10^{-8}} - A g\right) == \mu \left(I^2 L \frac{0.4 \pi}{B^2 \times 10^{-8}} - A g\right)$$

Note that lengths above are in cm, otherwise normal.

Putting $A\ell$ back in for Vc and solving for E gets

$$\ell + g \mu == \left(2 \mu E \frac{0.4 \pi}{A B^2 \times 10^{-8}}\right)$$

$$E == A (\ell + g \mu) \frac{B^2 \times 10^{-8}}{2 \mu 0.4 \pi}$$

So the energy handling capacity is proportional to the fakey path length (ℓ+gμ). The inductance as stated before is inversely proportional to the same factor, hence trade-off.

Found a cool formula for DC resistance in an19fa.pdf

$$Rdc == \frac{N \bar{\ell}}{12} 10^{\wedge}\left(\frac{AWG}{10} - 4\right) [\Omega]$$

Here N is number of turns, ℓ−baris the average turn length in inches, and AWG is the American Wire Gauge number. Example, 35 turns of 2.4 in average in #14 wire ⇒ 0.0176 Ω

For AC losses, see an19fa.pdf, they also do ferrite core losses.

Our particular case. It would be prudent to peak at about 1 A, 1/2 A is a good minimum goal. The typical peak switch current is 2A, assume a 15V input, limiting the peak current (peak being = 1/2 p-p) to 1A implies a certain inductance

```
fnom = N[10*^6 / 6.5] (* design operating frequency *)
```

```
1.53846 × 10^6
```

```
fmin = 1.1*^6 (* Assumes loss of synchronization *);
```

```
Vo (Vi - Vo) / (2 L f Vi) /. {Vo → 5, Vi → 15, f → fnom} ; (* Iavg *)
```

```
Solve[1 == %, L][[1, 1]]
```

```
L → 1.08333 × 10^-6
```

So even 1μH will satisfy the current requirements.

The peak inductor current is

```
ViMax = 4.2 * 4 ; ViMin = 2 * 4 ; Vshore = 18 ;
(* some input voltage limits *) (* Note, this is assuming 2 V/cell *)
```

$$Ip == Io + \frac{Vo\,(Vi - Vo)}{2\,L\,f\,Vi} \;/.\; \{Vo \to 5,\ Vi \to Vshore,\ f \to fmin,\ L \to 1*\hat{}-6,\ Io \to 1\}$$

$$Ip == 2.64141$$

It’s desirable for the inductor to saturate above this current.

Consider Coiltronics SD18−1R2,5.2 mm square, 1.8 mm tall. Shielded bobbin style.
Coiltronics DR73 series is adequate, but oversized.
Coiltronics MP2 series toroid for SMT. 5.88mm x about 4mm, 1.8 mm tall.
Coiltronics MP2A is same but better. Same size. See PM−4112.pdf

Part	L[μH]	I@+40 C[A]	Isat@70 %[A]	RDC[mΩ]	Vμs typ.
SD18 – 1 R2	1.2 ± 20 %	2.97	2.95	29.4	2.55
MP2 – 1 R0	1.0 ± 20 %	1.67	2.1	103	□
MP2A – 1 R0	1.0 ± 20 %	2.11	3.63	67	2.00
MP2A – 1 R5	1.50 ± 20 %	2.02	3.22	73	2.09

Volt·μs typ. is Vs product required to generate core loss equal 10% of total losses for 40 C temperature rise (@ 100 kHz for SD18, 300kHz for MP2A)

For our case the Vs product equation is

$$(Vi - Vo)\ Ton == Vo\ Toff$$

$$Vi\ Ton == Vo\ T$$

$$(Vi - Vo)\ \frac{Vo}{Vi\ f} == D\ Vi\ Toff$$

$$\frac{Vo}{f}\left(1 - \frac{Vo}{Vi}\right) ==$$

Evidently the product is a maximum when Vi is maximum

$$\frac{Vo}{f}\left(1 - \frac{Vo}{Vi}\right) \;/.\; \{Vi \to Vshore,\ Vo \to 5,\ f \to fnom\} \;//\; N$$

$$2.34722 \times 10^{-6}$$

This seems acceptable.

In some ways i dislike this procedure, there is no technical data on core loss, frequency characteristics, etc. Even the core material is unspecified, though it's probably that MnxZNx stuff. On the other hand the manufacturer claims to have optimized the design, so yay if they're not full of it. Alas tempus fugit.

This inductor value will cause discontinuous operation when the output current is below (worst case)

```
Vo (1 - Vo / Vi)
---------------- /. {Vi → ViMin, Vo → 4.5, f → fnom, L → 1.5*^-6}
     2 f L

0.426562
```

This is the usual situation. However this is not a big problem unless the regulator is cycle-skipping. Even skipping isn't necessarily a big problem, but it could be a source of RFI. Since i can't find a minimum on-time spec for this part, assume 70ns, which seems pretty reasonable. The minimum peak current is

```
ViMin - 5
--------- * 70*^-9 /. L → 1.5*^-6
    L

0.14
```

The decay time for this current is

```
Ip L
---- /. {Ip → %, L → 1.5*^-6} // EngineeringForm
 5

42. × 10^-9
```

So the average output current is

```
f Ip Toffmin / 2 /. {f → fnom, Toffmin → %, Ip → %%}

0.00452308
```

The minimum load will be more than this so everything is fine.

■ L201, L202 Coiltronics CMS1−11,120μH, See PM−4313.pdf

Coiltronics CMS1−11
120 μH nominal inductance (100kHz, 0 DC A, 0.1 Vrms)
1.35 A Irms Max (160 C max temp)
70 mΩ DC Ohms @ 20C typ. (per leg)
1.10 μH Leakage inductance
2.9 pF Interwinding capacitance
−40to +125C storage temperature range
−40to +85C operating temperature range
300 VDC maximum voltage
internal "hot spot" temperature limit 130C
molybdenum permalloy core
rated to 500kHz ??? —I don't see any reason for this.
Maximum dimensions: Body Length 9.4 mm, Width 7.2 mm, Height 2.6 mm.

This part is a common mode choke. It was chosen heuristically on the basis of reasonable package size and low DC losses. Based on a quick search, to get greater attenuation requires either a larger package or 30+% more DC resistance. The current rating is adequate.

Capacitors(s)

■ C200 22μF, 1206, X5R, Panasonic ECJ−3YB0J226M(DigiKey PCC2242CT−ND12.87$/10)

Panasonic ECJ−3YB0J226M
22 μF nominal capacitance
6.3V WV DC
X5R dielectric
1206 case size
±20% tolerance
−35to +85C operating temperature range
Height 1.6 mm.

The output capacitor for a high frequency buck switching power supply. The dynamic properties of this capacitor affect the loop stability so a moderately stable X7R or X5R type is typical. A tantalum cap. will work fine here, but is probably bigger. The ceramics seem so much cooler. Voltage rating ≥ 6.3V.

RMS ripple formula (from 1767f.pdf, check this???(Correct up to factor 0.58)).

$$\Delta\text{Irms} \to 0.29\,\frac{\text{Vo}\,(\text{Vi}-\text{Vo})}{\text{f L Vi}}\ \text{/. \{f} \to \text{fnom, L} \to \text{1.5*^-6, Vo} \to \text{5, Vi} \to \text{ViMin\} // N}$$

$$\triangle\text{Irms} \to 0.235625$$

Output ripple current is maximized at high input voltage.

$$\Delta\text{Irms} \to 0.29\,\frac{\text{Vo}\,(\text{Vi}-\text{Vo})}{\text{f L Vi}}\ \text{/. \{f} \to \text{fnom, L} \to \text{1.5*^-6, Vo} \to \text{5, Vi} \to \text{Vshore\} // N}$$

$$\triangle\text{Irms} \to 0.453796$$

5mV would be a good target for the output ripple.

$$\text{5*^-3} == \frac{\Delta\text{Irms}}{\text{fnom C}}\ \text{/. \%}$$

$$\frac{1}{200} == \frac{2.94968\times 10^{-7}}{\text{C}}$$

```
Solve[%, C][[1, 1]] // EngineeringForm
```

$$\text{C} \to 58.9935\times 10^{-6}$$

an−19p.23 contradicts this

$$C \geq \frac{1/(8\,L\,f^2)}{Vpp/(Vo\,(1-Vo/Vi))}$$

```
/. {Vpp → 5*^-3, f → fnom,
   L → 1.5*^-6, Vo → 5, Vi → Vshore} // N // EngineeringForm
```

$C \geq 25.4282 \times 10^{-6}$

The biggest in the DigiKey catalog is only 22μF, sad.

C201 10μF, 25V, 1210, X5R, Panasonic ECJ−4YB1E106M(DigiKey PCC2243CT−ND 12.63$/10)

Panasonic ECJ−4YB1E106M (used to be ECJ−4YF1E106Z)
10 μF nominal capacitance
25V WV DC
X5R dielectric
1210 case size
+/−20% tolerance
−35to +85C operating temperature range
Height 2.5 mm.

This is an input capacitor for a high frequency buck switching power supply. The standard choice is a 1−5 μF ceramic capacitor. The Y5V or similar dielectrics are suitable. Voltage rating should be at least 20V.

Input ripple current (buck) is (This formula is from 1767f.pdf, is this really correct???)

$$\Delta Irms \to Io\sqrt{Vo\,(Vi-Vo)/Vi^2}\ /.\ \{Io \to 1,\ Vo \to 5,\ Vi \to ViMin\}\ //\ N$$

$\Delta Irms \to 0.484123$

The ripple in the capacitor is

$$V == Q/C$$

Try 2.2μF

$$\frac{\Delta Irms}{fnom\ C}\ /.\ \{\%,\ C \to 2.2{*}\hat{}{-6}\}$$

0.143036

This seems ok, but check the web for a better value in 1206.

I can get 10μF in 1210 size, X5R, this should be excellent.

■ C202 Any 25+V 0.1μF ceramic cap in 1206 package (DigiKey PCC104BCTCT−ND 0.35$ea)

Panasonic ECJ−4VB1H104K
0.1 μF nominal capacitance
50V WV DC
X7R dielectric
1206 case size
±10% tolerance
−55to +125C operating temperature range
Height 1.0 mm.

Boost cap for U200. Voltage at least 20V. Linear recommends 0.1μF, who are we to argue?

■ C203 22μF, Tantalum, 25V, KEMET T491 series T491D226K025AS (DigiKey 399−1639−1−ND 0.74$ea)

KEMET T491 series T491D226K025AS
22 μF nominal capacitance
25V WV DC
tantalum dielectric
7343−D case size
±10% tolerance
0.06 Max dissipation factor
>8.3μA? Max leakage spec. meets EIA 535BAAC (also stated leakage<CV/100 but not less than 1/2μA?)
0.2Ω ESR @ 100kHz
<866mA? Max ripple current
−55to +125C operating temperature range
Height 2.8 mm.

This is not a critical component. Its primary function is as a noise filter. The secondary function is local energy storage. Voltage rating should be 20+V. High value and low ESR are desirable. Tantalum was chosen for compactness and low ESR. It’s desirable to limit package size, which becomes the primary constraint.

Originally selected a Panasonic TEL EEJ−L1ED336R(33μF). This is fine, but it’s 33$/10. This is too expensive. There is a KEMET version available in one-ses, but the 22μF version is only 0.74$. Simply cannot not do that.

■ C204 0.33μF, 25V, 1206, X7R, Panasonic ECJ−3VB1E334K (DigiKey PCC1889CT−ND 0.25$ea)

Panasonic ECJ−3VB1E334K
0.33 μF nominal capacitance
25V WV DC
X7R dielectric
1206 case size
±10% tolerance
−55to +125C operating temperature range
Height 0.85 mm.

This is a non-critical component. It’s function is as a very high frequency noise filter. To be effective in this application requires low ESR which suggests a ceramic type capacitor. It’s also desirable to select a fairly low cost component. Voltage rating ≥20V.

DigiKey has 0.33μF caps for 25¢ while the 1μF version is 45¢, given the current mood the 0.33μF version should suffice.

■ C205/C206

Frequency compensation component. See R206 for explanation.

■ C207 0.68μF Panasonic ECJ−3VB1C684K (DigiKey PCC1880CT−ND0.44$ea/10)

Panasonic ECJ−3VB1C684K
0.68 μF nominal capacitance
16V WV DC
X7R dielectric
1206 case size
±10% tolerance
−55to +125C operating temperature range
Height 0.85 mm.

Partly determines comparator response. See R211 for explanation.

■ C208/C209

Frequency compensation component. See R206 for explanation.

■ C210 Any 25+V 0.1μF ceramic cap in 1206 package (DigiKey PCC104BCT−ND0.35$ea)

Check out C202

Effects dynamic characteristics of power-down shutdown. See R209 for explanation.

This is a de-glitch capacitor. It also adds some time-constant delay. The value may as well be 0.1μF.

```
τ → R C /. {R → r1 r2 / (r1 + r2)} /. {r1 → 60*^3, r2 → 10*^3, C → 0.1*^−6} // N //
 EngineeringForm

τ → 857.143 × 10^-6
```

Basically 1ms, that’s fine.

■ C211 470nF, 16V, ceramic cap in 1206 package (DigiKey PCC1878CT−ND$2.55/10)

Panasonic ECJ−3VB1C474K
470 nF nominal capacitance
16V WV DC
X7R dielectric
1206 case size
±10% tolerance
−55to +125C operating temperature range
Height 0.85 mm.

See R202

■ C250 0.1μF, ceramic cap in 1206 package

Bypass cap. See C202, possibly cheaper would do.

■ C251 ???pF, ceramic cap in 1206 package

The purpose of this cap is to de-glitch a race condition in the counter logic. This race condition is unlikely to crop up, but this capacitor will eliminate it if required.

Resistors(s)

R200 10.0kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−10.0KFCT−ND 1.17$/10)

N.B.
As of SPS v0.7 the divider arrangement has changed. The voltage feedback divider is separated from the output window comparator. As a result the calculations below need to be re-done.
This needs to be re-calculated for the HAP, which uses a slightly different output voltage (See CR402).

Panasonic P−60.4KFCT−ND
8.25kΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

R200, R201 form a voltage divider that provides voltage feedback for the SPS buck regulator.

The SPS output voltage for the HAP is modified to 4.69V (see CR402). (Normally the output is 5V.)

The LT1767 data sheet provides a formula for the divider values based on the nominal reference voltage and the expected FB-pin output current of 250nA (500nA Max). The suggested value for R200 based on the datasheet is 10k . Following this line

$$R201 \to \frac{Vo - Vfb}{Vfb / R200 - Ib} \text{ /. } \{Vo \to 4.69\,V,\ Vfb \to 1.2\,V,\ Ib \to 250*\text{^}-9\,V/\Omega\} \text{ /. } R200 \to 10*\text{^}3\,\Omega \text{ // N // EngineeringForm}$$

$$R201 \to (29.1441 \times 10^{3})\ \Omega$$

For 5V

$$R201 \to \frac{Vo - Vfb}{Vfb / R200 - Ib} \text{ /. } \{Vo \to 5\,V,\ Vfb \to 1.2\,V,\ Ib \to 250*\text{^}-9\,V/\Omega\} \text{ /. } R200 \to 10*\text{^}3\,\Omega \text{ // N // EngineeringForm}$$

$$R201 \to (31.7328 \times 10^{3})\ \Omega$$

The current in R200 is around

```
1.2 / 10*^3 (* A *) // EngineeringForm
```

$$120. \times 10^{-6}$$

```
% / 250*^-9
```

```
480.
```

This is 480 times the bias current, which should be fine.

Using standard values

```
Solve[R201 == (Vo - Vfb)/(Vfb/R200 - Ib), Vo][[1, 1]]
```

$$Vo \to -\frac{Ib\,R200\,R201 - R200\,Vfb - R201\,Vfb}{R200}$$

```
% /. {R200 → 10*^3 Ω, R201 → 29.4*^3 Ω, Vfb → 1.2 V, Ib → 250*^-9 V / Ω}
```

Vo → 4.72065 V

This should be fine.

Old calculations (<v0.6) :

R200−203 form a voltage divider which serves 3 purposes. As the voltage measuring element of the regulator feedback loop, as the measuring element for the over voltage protection, and as the measuring element for the undervoltage detection/current limiting mechanism.

The LT1767 data sheet provides a formula for the divider values based on the nominal reference voltage and the expected FB-pin output current of 25μA (50μA Max)

The reference voltage is fairly tight. ±2% over temperature.

```
RA → (V - Vfb)/(Vfb / RB - Ib) /. {V → 5, Vfb → 1.2, Ib → 25*^-6} /. RB → 10*^3 // N //
  EngineeringForm
```

$RA \to 40. \times 10^3$

To keep bias current errors in the 1/4-1/2 % range, keep RB below 10-20k

```
RA → (V - Vfb)/(Vfb / RB - Ib) /. {V → 5, Vfb → 1.2, Ib → 25*^-6} /. RB → 20*^3 // N //
  EngineeringForm
```

$RA \to 108.571 \times 10^3$

Current could be saved by going to 20kΩ, but the increased sensitivity isn’t a good idea since we’re hooking other opamps to this divider, the total divider current with a 10k base is only ~100μA, which is ok here.

Overvoltage definition: (6.0, 6.1, 6,2)V. Undervoltage definition: (4.6, 4.7, 4.8)V.

With a 50k total resistance, the correct tap points are easily found. The nominal reference voltage of U201 is 1.182V.

First solve for the over voltage tap (Vref is the reference voltage of U201)

```
Vref == Vfb r0/(r0 + r1) /. Vfb → (Ib RA + V) RB/(RA + RB) // FullSimplify
```

$$Vref == \frac{r0\,RB\,(Ib\,RA + V)}{(r0 + r1)\,(RA + RB)}$$

```
Solve[{%, RB == r0 + r1}, {r0, r1}][[1]] // FullSimplify
```

$$\left\{r0 \to \frac{(RA + RB)\ Vref}{Ib\ RA + V},\ r1 \to RB - \frac{(RA + RB)\ Vref}{Ib\ RA + V}\right\}$$

```
% /. {RA → 40.0*^3, RB → 10.0*^3, Ib → 25*^-6, Vref → 1.182, V → 6.1}
```

```
{r0 → 8323.94, r1 → 1676.06}
```

Now solve for the under voltage tap

$$\text{Vref} == \text{Vfb} + \frac{\text{V} - \text{Vfb}}{\text{RA}}\ \text{r2}\ /.\ \text{Vfb} \to (\text{Ib RA} + \text{V})\ \frac{\text{RB}}{\text{RA} + \text{RB}}\ //\ \text{FullSimplify}$$

$$\text{Vref} == \frac{\text{Ib}\ (-\text{r2} + \text{RA})\ \text{RB} + (\text{r2} + \text{RB})\ \text{V}}{\text{RA} + \text{RB}}$$

```
Solve[{%, RA == r2 + r3}, {r2, r3}][[1]] // FullSimplify
```

$$\left\{r3 \to -\frac{(RA + RB)\ (V - Vref)}{Ib\ RB - V},\ r2 \to \frac{Ib\ RA\ RB + RB\ V - (RA + RB)\ Vref}{Ib\ RB - V}\right\}$$

```
% /. {RA → 40.0*^3, RB → 10.0*^3, Ib → 25*^-6, Vref → 1.182, V → 4.7}
```

```
{r3 → 39528.1, r2 → 471.91}
```

These have to hit standard values, try 8.25k , 1.69k , 475Ω , 39.2k

$$\text{Solve}\left[\text{RA} == \frac{\text{V} - \text{Vfb}}{\text{Vfb} / \text{RB} - \text{Ib}},\ \text{V}\right][[1, 1]]\ //\ \text{FullSimplify}$$

$$V \to -Ib\ RA + Vfb + \frac{RA\ Vfb}{RB}$$

$$\text{Solve}\left[\text{Vref} == \frac{\text{r0 RB (Ib RA + V)}}{(\text{r0} + \text{r1})\ (\text{RA} + \text{RB})},\ \text{V}\right][[1, 1]]\ //\ \text{FullSimplify (*overvolt*)}$$

$$V \to -Ib\ RA + \frac{(r0 + r1)\ (RA + RB)\ Vref}{r0\ RB}$$

```
Solve[Vref == (Ib (-r2 + RA) RB + (r2 + RB) V)/(RA + RB), V][[1, 1]] //
 FullSimplify (*undervolt*)
```

$$V \to \frac{Ib\,(r2 - RA)\,RB + (RA + RB)\,Vref}{r2 + RB}$$

```
sub = {RB → r0 + r1, RA → r2 + r3, Ib → 25.0*^-6, Vfb → 1.2, Vref → 1.182};
```

```
val = {r0 → 8.25*^3, r1 → 1.69*^3, r2 → 475., r3 → 39.2*^3};
```

```
Vo → Vfb (1 + RA/RB) - RA Ib /. sub /. val (* output voltage *)
```

Vo → 4.99786

```
Vov → Vref ((r0 + r1) (RA + RB))/(r0 RB) - Ib RA /. sub /. val (* over voltage *)
```

Vov → 6.1166

```
Vuv → (Ib (r2 - RA) RB + (RA + RB) Vref)/(r2 + RB) /. sub /. val (* under voltage *)
```

Vuv → 4.69551

Definitely close enough.

■ R201 29.4kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−29.4KFCT−ND 1.17$/10)

See R200 for details

■ R202 19.6kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−19.6KFCT−ND 1.17$/10)

As of SPS v0.7 the window comparator has its own divider consisting of R202-204. The target impedance is about 100kΩ .

As extra fun a feedback capacitor (C211) is employed to assure 10ms of Over Voltage Protect (OVP) signal which should be sufficient to blow the fuse.

Using the previous definitions : Overvoltage (6.0, 6.1, 6,2)V, Undervoltage (4.6, 4.7, 4.8)V. The reference voltage is 1.182V.

```
windowparm = {Vr → 1.182 V, Vlow → 4.7 V, Vhigh → 6.1 V, Rz → 100*^3 Ω};
```

```
windoweq = {Vr == Vlow * (R202 + R203) / Rz,
   Vr == Vhigh * (R202) / Rz, Rz == R202 + R203 + R204};
```

```
Solve[windoweq, {R202, R203, R204}][[1]]
```

$$\left\{R204 \to -\frac{-Rz\,Vlow + Rz\,Vr}{Vlow},\ R203 \to -\frac{-Rz\,Vhigh\,Vr + Rz\,Vlow\,Vr}{Vhigh\,Vlow},\ R202 \to \frac{Rz\,Vr}{Vhigh}\right\}$$

```
% /. spsparm
```

{R204 → 74851.1 Ω, R203 → 5771.89 Ω, R202 → 19377. Ω}

Stick to std values

```
windowsoln = {Vr → 1.182 V, R202 → 19.6*^3 Ω, R203 → 5.9*^3 Ω, R204 → 75.0*^3 Ω};
```

```
Solve[windoweq, {Vlow, Vhigh}, Rz][[1]]
```

$$\left\{Vhigh \to \frac{(R202 + R203 + R204)\ Vr}{R202},\ Vlow \to \frac{(R202 + R203 + R204)\ Vr}{R202 + R203}\right\}$$

```
% /. windowsoln
```

{Vhigh → 6.06077 V, Vlow → 4.65847 V}

This is a fine result. The impedance at the positive voltage reference is

$$Rzplus = \left(\left(\frac{R1\,R2}{R1 + R2}\ /.\ \{R1 \to R202,\ R2 \to (R203 + R204)\}\right)\ /.\ windowsoln\right)$$

15777.5 Ω

Seek C211 so that the on time of the OVP output is 10ms even if V_SPS goes to zero. Assume initially OVP steps from 0V to 5V, V_SPS is 0V but Vinplus is 1.182V. How long does it take to discharge C211 to 1.182V? Neglect R205.

The initial voltage is

```
1.182 V + 5 V
```

6.182 V

The overvoltage should be neglected for several reasons, mostly because it’s discharged via the input diodes of U201 through R203, which is fairly small (5k). Assume the initial voltage is 5V.

```
Solve[Vx == Vi e^(-t/(R C)), t][[1, 1]]
```

Solve::ifun : Inverse functions are being used by Solve, so some solutions may not be found.

$t \to -C\,R\,\mathrm{Log}\left[\frac{Vx}{Vi}\right]$

```
t == 10*^-3 s /. % /. {Vx → Vr, Vi → 5 V, R → Rzplus} /. windowsoln
```

$22754.8\,C\,\Omega == \frac{s}{100}$

```
Solve[%, C][[1, 1]] /. Ω → s / F // EngineeringForm
```

$C \to \left(439.468 \times 10^{-9}\right) F$

Not too scary, the time constant is

```
R C /. % /. {R → Rzplus, F → s / Ω} // EngineeringForm
```

$\left(6.93371 \times 10^{-3}\right) s$

Moving to a 0.1μF cap raised the time constant to

```
R C /. C → 0.1*^-6 F /. {R → Rzplus, F → s / Ω} // EngineeringForm
```

$\left(1.57775 \times 10^{-3}\right) s$

This might be a little slow, maybe 470nF.

■ R203 5.90kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−5.90KFCT−ND 1.17$/10)

See R202 for details

■ R204 75.0kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−75.0KFCT−ND 1.17$/10)

See R202 for details

■ R205 47.5kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−47.5KFCT−ND 1.17$/10)

R205 functions as a isolation and current limiting device. I think it may as well be the standard pull-up value (47k). Higher risks slow response or noise sensitivity, and it won’t help hold up the voltage because of R203. Lower is not necssesary if the pull-up value is chosen correctly. (At one point this was marked 100k, IDKW.)

■ R206 ???kΩ , Any ???% or better film resistor in 1206 package

R206, along with C205,206, and possibly C208,209, forms the frequency compensation component of the SPS switching supply.

As a basic reference take an-19 p.48-50. A more detailed reference is an76.

Adequate testing includes using the actual power source (to account for source impedance) and testing load transients over input/output range and temperature. Check transients on start up and short circuit.

The goal is to have the highest possible cross over frequency while maintaining a phase margin of 45+° and a gain margin better than −10dB.

1st estimate DC loop gain (gm is error-amp gain, R_ea is error-amp load resistance to ground, Gm is output stage gain, and R_L is the output load resistance == Vo/Io.) Load resistance ranges from about 10Ω to 200+Ω (from 500mA to ≈20mA). Not sure which is worse for stability.

$$\frac{Vfb}{Vo}\,(gm * Rea)\,(Gm * RL)\ /.\ \{gm \to 850*^{\wedge}-6,\ Gm \to 2.5,\ Vo \to 5.0,\ Vfb \to 1.2,\ Rea \to 412*^{\wedge}3,\ RL \to 10\}$$

```
2101.2
```

Gain (66dB) increases at lower output currents, up to factor >20.

The highest corner frequency is at highest current. Target here is 500mA⇒ R_L == 10Ω

$$f0 \to 1/(2\,\pi\,RL\,Co)\ /.\ \{RL \to 10,\ Co \to 22*^{\wedge}-6\}\ //\ N$$

$$f0 \to 723.432$$

Conventional buck design proceeds from the assumption that voltage-mode control is bad because the inductor and output capacitor form a tank circuit (double pole) at a resonant frequency within the gain-range of the supply. Hence current-mode control whose gain roll-off is dominated by the R_L*Co output pole. Conventionally the phase lag is limited by the zero created from Co and Co’s ESR. For ceramic output caps this doesn’t work (ESR too low). The classic solution is to add a zero to the error amplifier’s response by inserting a series resistor (Rc) into the compensation capacitance (Cc) (See data sheet for LT1375/1376 13756fc.pdf, p 21). Increasing Rc raises loop BW and therefore improves transient response. However, as Rc increases, gain at higher frequencies makes Vc susceptible to noise injected from the switching node. This can be filtered with a parallel capacitor (Cf) from Vc to ground. The pole of this capacitor is typically at 1/5th the switching frequency to provide significant switching noise attenuation, but still be high enough not to disturb loop dynamics. (13756fc.pdf p.22 is excellent on loop testing). It’s desirable that the gain of the loop be <1 at the switching frequency.

Additionally two capacitors CA and CB (C208, 209 in this case) have been added. These are not usually required but can be beneficial. CA is a speed-up capacitor. Its zero produces a bump in the phase plot "Ideally, the peak of this bump is centered over the loop’s cross-over frequency." (See an76 p.5). CB usually functions as a noise filter. Typical pole frequencies are 1/2 to 1/3 of the switching frequency. It may interact with CA, and can be used to swamp fb-pin input capacitance if that is important (it isn’t).

An76 p.9 fig.8 has an alternate load pulser for SPS testing which looks good.

Here are the most basic equations for buck SPS

$Av_{(fb)} == Vref / Vo$	Voltage gain due to feedback divider
$Av_{(ea)} == gm * R_{ea}$	Voltage gain of error amp
$Av_{(mod)}$	Voltage gain of modulator section
$DC\ gain == Av_{(fb)} * Av_{(ea)} * Av_{(mod)}$	Total DC voltage gain
$fp_o == (2 \pi R_L\ Co)^{-1}$	Output pole frequency
$fz_o == (2 \pi\ ESRo * Co)^{-1}$	Output zero frequency (unusably high for ceramic caps)
$fp_{ea} == (2 \pi R_{ea}\ Cc)^{-1}$	error amp pole frequency
$fz_{ea} == (2 \pi\ Rc\ Cc)^{-1}$	error amp zero frequency
$fp_{fb} == (2 \pi\ Rc\ Cf)^{-1}$	compensation feedback pole frequency
$fz_A == (2 \pi\ RA\ CA)^{-1}$	speed-up zero frequency
$fp_B == (2 \pi\ RB\ CB)^{-1}$	slow-down pole frequency

The basic idea is to begin with "reasonable values" and apply transient loads similar to expected loads at several values of DC output current. This is done for all interesting voltage input, input resistance, and temperature conditions.

Initially Cc is chosen large and Rc small. Loop response is checked (overdamped expected otherwise raise Rc, if required raise Cc). Then Cc is reduced in 2:1 steps until ringing begins, then Rc is increased by 2:1 steps until ringing ceases. Then Cc is again reduced, etc. until no further improvement is possible.

The other components can be guessed at and tried. 10:1 experiments will verify the ballpark ranges.

Variational testing is done by replacing compensation components over the 2:1 range at room temperature and average conditions. No significant transient response deviations should be noted.

Final check out is done with the selected compensation component values over the full voltage/load/temperature matrix. Using the final input supply and output load is helpful at this stage.

A good starting point for U200 should be Cc == 2.2nF and Rc == 0Ω .

Note that "small signal" ≡ "closed loop" response is indicated by 1) symmetric response on both edges of transient, 2) linear range at Vc pin 3) dV/dt @ Vc pin less than slew rate limit (calculate from Vc pin impedance, ea drive).

■ R207 100kΩ , Any 5% or better film resistor in 0805 package (DigiKey P−100KCCT−ND 0.91$/10)

See R200 for typical specs.

Changed this to the pull-up for a bipolar because SPS_Off was changed to an O.C. output, same value should work.

R207 discharges the gate on Q201. While not strictly required, this provides some noise immunity, and assures that Q201 stays off if SPS_Off is tri-stated.

100kΩ is reasonable. A harsh RF environment would be required to overcome a 100k shunt, yet the increase in current when on is still smallish.

■ R208 10.0kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−10.0KFCT−ND 1.17$/10)

See R200 for typical specs.

Changed Q201 to bipolar, still same deal here.

R208 separates the gate of Q201 from direct connection to the PIC. A direct connection would certainly work, but the resistance isolates the gate from any funny business in the PIC power supplies. What is more reliable depends on the spectrum of failure modes. Intuition and some experience sez that direct connection to a FET gate could be problematic.

Even at 4.5V, the gate voltage will reach 4V

```
4.5 * 100 / 110 // N
```

```
4.09091
```

Since the threshold voltage is under 3V, the scheme should work.

■ R209 60.4kΩ , Any 2% or better film resistor in 1206 package (DigiKey P−60.4KFCT−ND 1.17$/10)

See R200 for typical specs.

R209,210 form a resistive voltage divider which shuts down U200 when the input bus voltage falls below a certain value. It also serves to activate U200 when bus voltage is present or restored.

Shutting down U200 when the bus voltage is low does two things. It decreases the chance of over-discharge of the main battery, and it makes the transition to backup power crisper. A crisp transition is less noisy, and seems less risky.

The Anticipated range of the bus voltage is:

```
20 V     Should never get this high
18 V     Shore Power
16.8 V   Maximum Battery Voltage (4.2 V / cell)
14.4 V   Nominal Battery Voltage  (3.6 V / cell)
10 V     Minimum Battery Voltage  (2.5 V / cell)
9 V      Should never get this low
```

The LT1767 datasheet helpfully specifies:

Shutdown voltage range (over temperature) (1.27, 1.33, 1.40)V (generally higher when hotter)
Shutdown current (threshold + 60mV) (over temperature) (7, 10, 13)μA
Shutdown current (threshold −100mV)(over temperature) (4, 7, 10)μA (This appears wrong???)

The two shutdown current entries represent a hysteresis mechanism. When running U200 sources about 10μA into the shutdown pin. In the shutdown state U200 sources only about 3μA. (The data sheet appears wrong about this spec.???)

The datasheet gives design formula. Using 9.75±0.2V as the target range gives:

```
RA → (VH - VL)/(7*^-6) /. {VH → 9.95, VL → 9.55} // N // EngineeringForm
```

```
RA → 57.1429 × 10^3
```

```
RB → Vcenter/((VH - Vcenter) / RA + 3*^-6) /. {VH → 9.95, VL → 9.55, Vcenter → 1.33, %} //
  N // EngineeringForm
```

```
RB → 8.64478 × 10^3
```

Using standard resistors, this would be 57.6K, 8.66K.

```
ra = 60.4*^3; rb = 10.0*^3;
```

```
Vth = 1.33;
```

Check:

Nominal

```
Solve[(V - Vth) / ra - Vth / rb == 10*^-6, V]
```

```
{{V → 9.9672}}
```

```
Solve[(V - Vth) / ra - Vth / rb == 3*^-6, V]
```

```
{{V → 9.5444}}
```

Worst case

```
Vth = 1.40; Solve[(V - Vth) / ra - Vth / rb == 13*^-6, V]
```

```
{{V → 10.6412}}
```

```
Vth = 1.27; Solve[(V - Vth) / ra - Vth / rb == 2*^-6, V]
```

```
{{V → 9.0616}}
```

Worst case includes manufacturing spread, so probably this is ok. Shut down hysteresis current is fairly flat w/ temperature, so if we test and trim away from 10V, SPS should work over the whole battery range. An trim could be added to fine-tune this action if needed.

■ R210 10.0kΩ , Any 2% or better film resistor in 1206 package (DigiKey P−10.0KFCT−ND 1.17$/10)

See R200 for typical specs and calculations.

■ R211 590Ω , Any 5% or better film resistor in 1206 package (DigiKey P−590FCT−ND 1.17$/10)

See R200 for typical specs and calculations.

R211 forms part of a low pass filter on the reference output of U201. The low pass filter is intended to prevent false triggering due to noise on the reference output, particularly due to switching noise or other interference on the power input lines. The associated capacitor, C207, should be as large as convenient. The 16V Y5V style is available as a 2.2μF unit only 0.85mm high. The LTC1442 datasheet specifies 300Ω minimum resistor for stability. 390Ω gives a quite adequate safety margin.

While filling out the spec, the mind changed. The problem is Y5V looses a lot of capacitance at elevated temperature. Lower capacitance values require more resistance for stability. So at the minimum the resistance should be figured based on minimum capacity over temperature. This would be 0.44μF (−80%)implying a resistance of about 700Ω. An X5R or similar dielectric holds <±10% of its initial value, so it is a better choice here.

A 0.68μF X7R type can be had for 44¢ and 0.85mm tall, this should be fine. The required resistance is about 550Ω. Choose 590Ω as a safe bet.

■ R212 44.2kΩ , Any 2% or better film resistor in 1206 package (DigiKey P−44.2KFCT−ND 1.17$/10)

See R200 for typical specs.

See R213 for calculations.

■ R213 4.70MΩ , Any 2% or better film resistor in 1206 package (DigiKey 311−4.70MFCT−ND 1.35$/10)

Phicomp 311−4.70MFCT−ND
4.70MΩ nominal resistance
200V WV DC or RMS
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

R212, 213 form a hysteresis network for U201 (LTC1442).

Here are the specs
Vref 1.182±1.5% over temperature
Current source for ΔVref==1mV, (100, 200,)μA
Current sink for ΔVref==2.5mV, (10, 20,)μA
Voltage noise [100, 100k]Hz == 100μVrms typ.
Vhyst range [Vref−50mV,Vref]

If a 100mV hysteresis is assumed on over voltage and under voltage trip, then 1st the divider ratio is needed.

```
{1.182 / 6.1, 1.182 / 4.7}
```

```
{0.19377, 0.251489}
```

```
√(Times @@ %)
```

```
0.220751
```

```
50 mV * %
```

```
11.0376 mv
```

So Vref−11mVshould be applied to the Vhyst pin.

As a 1st try, set the hysteresis divider current to 0.25μA

```
RA → Vh / Ih /. {Vh → 11*^-3, Ih → 250*^-9} // N // EngineeringForm
```

```
RA → 44. × 10^3
```

```
RB → (Vref - Vh) / Ih /. {Vref → 1.182, Vh → 11*^-3, Ih → 250*^-9} // N //
 EngineeringForm
```

```
RB → 4.684 × 10^6
```

Possible standard values are 4.70M & 44.2k

Check

```
Vh → Vref RA/(RA + RB) /. {Vref → 1.182, RA → 44.2*^3, RB → 4.70*^6} // N //
 EngineeringForm
```

$$Vh \to Vref \frac{RA}{RA + RB}$$

```
Vh → 11.0123 × 10^-3
```

Seems fine.

■ R214 ~1.8kΩ , Any 5% or better film resistor in 1206 package

See R304

Current limiting resistor for SPS power indicator LED, see D201.

Assuming a high efficiency LED, we can probably use 2mA, a typical forward voltage is 1.5V.

```
r → (5 - 1.5) / (2*^-3)
```

```
r → 1750.
```

■ R215 100kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−100KFCT−ND 1.17$/10)

See R205

R215 provides a DC path from the SPS-node ground to the chassis ground. There’s no known reason for a lesser value. 100K is high enough for decent isolation.

Miscellaneous

■ F200 Littelfuse "1206" (3216) 429-series Fast Acting fuse (DigiKey F1234CT−ND 0.70$ea)

Littlefuse 429−FH
1A nominal current
63V Max interrupting voltage
3216 (1206) case size
−55to +125C operating temperature range
Height 1.2 mm.
Blow times @ 25C
 100% > 4hr
 200% < 5s
 300% < 0.2s
 800% < 50ms, > 2ms

Fast acting fuse. Protects each CAN node in the event of catastrophic fault and likewise protects the power bus.

The exact value is dependent on complex details, and really should be measured. However is we assume the maximum SPS output is 1A @ 5V the maximum input should be about

```
Io    Vo
-- ------- /. {Io → 1, η → 0.75, Vo → 5.2, Vimin → 2.5 * 4} // N
η   Vimin
```

```
0.693333
```

We seek to interrupt fast, which requires ~800% overload. The 1A model should not present a problem.

A question has been asked concerning the blowing of this fuse in an over-voltage protection scenario. When the overvoltage protection triggers, the R_DS_ON of Q200 reaches 7.5 mΩ . The input cap is 33μF, giving a time constant of

```
7.5*^-3 * 33*^-6 (* s *) // EngineeringForm
```

```
247.5 × 10^-9
```

The discharge energy is

```
1 / 2 * 15^2 * 33*^-6 (* J *) // N // EngineeringForm
```

```
3.7125 × 10^-3
```

Assuming an internal resistance of 0.1Ω , the fusing energy is

```
0.1 {3^2 * 200*^-3, 8^2 * 25*^-3} (* J *)
```

```
{0.18, 0.16}
```

So the input cap has nothing like the energy needed to blow the fuse.

The peak current in OVP mode is a function of the wiring but the choke is about 70mΩ, the fuse is 0.1Ω, and the wiring is at least 50mΩ . The 4-cell Li+ battery has an internal resistance of around 0.1Ω , so the peak current is

```
15 / (0.07 + 0.1 + 0.05 + 0.075 + 0.1) (* A *)
```

```
37.9747
```

The time to blow at this current is

```
0.18 / (%^2 0.1) (* s *) // EngineeringForm
```

```
1.2482 × 10^-3
```

The voltage across the common mode choke assuming 38A ~1ms is

```
240*^-6 * 38 / 1*^-3 (* V *) // N
```

```
9.12
```

So the choke will be a middlin limit on the rate of current increase.

The droop on a typical 10μF output cap assuming a 50mA load over 1ms is

```
(50*^-3 * 1*^-3) / 10*^-6 (* V *) // EngineeringForm
```

```
5
```

The voltage droop on the bus should be

```
38 * (0.1 + 0.05) (* V *)
```

```
5.7
```

So the conclusion is that the nodes don’t have enough energy storage to either ride out a bus brown-out, or blow their own fuse. However the bus should be stiff enough to hold up while blowing a fuse, so things may be ok.

TVS200 Diodes Inc. SMBJ18A Transient Voltage Suppressor (DigiKey SMBJ18ADICT−ND 0.71$ea/10)

Diodes Inc. SMBJ18A
18V nominal voltage
SMB package
20V minimum breakdown voltage (1mA)
23.3V Maximum breakdown voltage (1mA)
20.5A Max peak pulse current
29.2V Max clamp voltage at 20.5A
100A Max peak current using single 8.3ms 1/2 sine wave
5μA Max reverse leakage

−55to +150C operating temperature range
Height 2.6 mm.

Protects the node in the event of over-voltage condition. Must be capable of blowing the fuse before U200 is destroyed. U200’s maximum input voltage is 25V.

The SMBJ18A is possibly just a little under-rated, but it’s quite close. I think for any plausible faults it’s fine.

W200 (probably eliminate)

This wire jumper exists to power U201 when the HAP is not present. Since the UPS and SPS power lines both run to the application area, probably eliminate this line and substitute a wire jumper in the application area on non-HAP versions.

W201 Power LED cuttable jumper

Cut this to disconnect the power-indicator LED. The jumper can be easily re-connected, either with a SMD zero-ohm jumper in 1206-ish package, or just a hunk of fine gage wire.

W202

This wire jumper bypasses U202, which is only present on HAP versions. W202 is cut prior to installation of U202.

High-Availability Power Supply (HAP) (400)

HAP Goals

Reliable / Robust power
Low mass
Small size
Good efficiency (>85%)
Low generated RFI

HAP Specs

One hour independent battery power at 100mA
Provide main power status to the PIC

HAP Design notes

The HAP provides 5V power to the CAN node in the event of main power failure. The HAP is not present on typical CAN nodes only those that must operate independently of main power. The only node currently incorporating the HAP is the recovery node.

The design of the HAP allows the node to be powered down under software control. Power is automatically restored whenever main power is available.

The HAP operates off a single Li+ cell. There are other ways to achieve the specifications but a single Li+ cell was chosen for several reasons.

- Multi cell batteries are less reliable. Besides mechanical connections, over time multi cell batteries often have problems with equalization and failure of individual cells. The life of a cell is harder in a multi cell pack compared to the relatively pampered conditions in a single cell battery. Also the single cell battery is easier to monitor and interpret, its behavior and its state are less complex.
- A single cell is probably lighter than any multi cell pack of equal capacity.
- Multi cell Li batteries are not utilized as effectively as single cell batteries. Some small Li+ cells are not rated for use in multi cell packs.
- Li cells were chosen because of their superior energy density and availability in convenient sizes at moderate cost.
- A rechargeable battery was chosen because of maintenance, reliability, and cost issues. A primary cell usually has to be un-soldered to be changed. It can be difficult to ascertain the state of charge of a primary battery. In comparison, a rechargeable battery can be tested and over time its state of charge can be reliably estimated. The only certain way to get high reliability from a primary battery is to install a new one well before it's estimated capacity is consumed, which entails expense and effort.

The most obvious way to incorporate a battery backup into the SPS supply is to diode switch a battery into the SPS input. The SPS minimum input voltage is 8V, which would require 4 Li+ cells (Possibly could be tweaked to 3 cells.) This is not a very attractive option, and has been rejected for aesthetic reasons.

If a mistake is being made with the design of the HAP, it would have to be using a rechargeable battery instead of a primary battery. There is a certain loss of capacity in going to the rechargeable battery, but the difference in complexity is not that great, assuming that the boost supply is used. The only way to get away from the boost supply is either to use the 4 cell backup, or go to a buck/boost on the input supply. 4 cells is intuitively trouble. Maybe the buck/boost option should have been explored more, maybe revision 2?

The HAP accepts two digital inputs and produces two digital outputs. There are two input power sources to the HAP. "V_BK" is battery backup power from the battery backup (BB see elsewhere in this section). The other power source is regulated power from the SPS. SPS power is not used directly by the HAP. The HAP uses SPS as a power up command. The HAP can be powered down so some mechanism must be provided to power the HAP back up. This happens automatically when SPS is restored.

Using power from V_BK two output power supplies are produced. The "HAP" high availability power, can be provided as long as battery or SPS power is available. However the HAP can be switched off by the PIC. HAP power is automatically restored when SPS is available. The other output power source "UPS" is the interruptible power supply, it remains within a diode drop of the battery voltage even when the HAP is powered down. When the HAP is up the UPS voltage is essentially equal to the HAP voltage.

The HAP provides digital output "HAP_Good" and its complement "HAP_Good\" from a comparator internal to the boost converter IC. These signals are used to disconnect certain voltages from the rest of the system when the HAP is off. This prevents phantom current drains. The HAP_Good\ signal is used internally by the HAP system to power down the HAP supply.

As was the case for the SPS, small size and high efficiency dictates that the HAP uses a high frequency switching supply. The HAP is a boost converter which like the SPS is synchronized to the PIC’s crystal oscillator using the "sync" digital input.

The final input to the HAP is the "Power_Down" command. This is generated by the PIC. The purpose of Power_Down is to completely shut off the power to the node. During power down only devices connected to the UPS receive any power. UPS connected devices are micro power in order to maximize battery life. Power down is necessary because otherwise a HAP could go no more than a few hours disconnected from the main supply without destroying the Li+ battery. A software power down is regarded as more reliable and less trouble than a mechanical switch.

■ HAP Issues

The SPS output voltage for the HAP has to be raised slightly from the non-HAP version to meet the battery charger minimum input voltage specs, see CR402

During power down certain key data could be recorded in EEPROM. This would include the current date and the state of the battery. If the battery encounters any unusual conditions (such as overdischarge) this should be recorded into EEPROM.

Since the HAP uses software powerdown, there is no way to turn it off without a CAN system or a 2m transmitter. This inconvenience has been construed as a feature. It may however require a lamp to be sure that the thing is really off. (This has been partially addressed. A jumper (J400) has been added which can shut the HAP down manually.)

It might be cleaner to power up the HAP using the SPS_Fail signal rather than the SPS supply directly. The same idea might be applied to the saturable rectifiers to guarantee their sequencing. — Done.

BB Design notes

The HAP incorporates a Li+ battery. For purposes of conceptualizing the design the elements dealing with the battery have been split off from the rest of the HAP as the battery backup (BB). The BB section contains the battery, its charge and monitoring circuitry, a transistor which acts as a saturable rectifier to switch in the battery during main power failure, a Schottky diode which prevents reverse current flow to the SPS, and an analog multiplexer that allows the PIC to monitor either the battery charge current, or the state of the SPS_Fail signal.

Complete charging of a Li+ battery requires a minimum two level current source with a strict voltage limit. This is supported in hardware by the BB section. The hardware could possibly be minimized by using a more elaborate charging algorithm in the PIC. This approach was rejected as too much work and too prone to error.

Several battery parameters are measured namely voltage, temperature, and charging current. Discharge current is not measured.

The BB has five digital inputs and three analog outputs. It uses three input power supplies and provides two output power supplies.

Two of the inputs, Fast_Charge and Trickle_Charge, control the charging current into the battery when the SPS power is applied. The charging voltage is limited to a safe value by circuitry internal to the charge controller, therefore the battery is not expected to explode even in the event of a software failure. The Trickle_Charge input also controls the analog multiplexer. If trickle charge is asserted the PIC sees the battery current. If deasserted the SPS_Fail signal is monitored. The active high SPS_Fail was chosen because a logic high voltage of > 4V cannot be generated by the battery current monitor during charging. The circuitry of the charger is such that Trickle_Charge is overridden by Fast_Charge, so Trickle_Charge is don’t-care during fast charge.

The other three digital inputs to BB are SPS_Fail, HAP_Good and its complement HAP_Good\ . SPS_Fail is an input to the analog mux and switches the saturable rectifier (Q404). HAP_Good controls the analog switch which separates the battery from one of the PIC ADC pins. This switch is required because otherwise an un-powered PIC would draw battery current through its input protection diodes. HAP_Good\ is used inhibit the analog mux. This function is reminiscent of the case involving the analog switch and the battery voltage, but since all the inputs should be at zero volts during power down, inhibiting the mux shouldn’t be required, but it’s done anyway because the mux has a pin for it. (Maybe it’s slightly safer to do it.)

BB’s three input power supplies are SPS, the main power input. UPS, which is used for the analog switches that must function even during power down. UPS is used over V_BB or B_BK because UPS runs at a higher voltage when the HAP is up, and at almost the same voltage — wrong! Can use V_BK and avoid a diode drop on U406. UPS is still correct for U405.

The last input supply is V_ADC. This is really a reference voltage for the battery thermistor. This supply should be the same as the PIC ADC reference voltage. This makes the thermistor voltage measurement ratiometric and eliminates a possible error due to ADC voltage reference drift.

BB provides two output power supplies. V_BB is simply a direct connection to the battery. No application currently uses this supply. The other output is V_BK. V_BK is connected to V_BB by a saturable rectifier. The rectifier part allows connecting V_BK to a higher voltage supply while the saturable part allows high efficiency. There is a little magic involving the saturable rectifier. When the SPS comes up, current will possibly dump into the battery. The SPS current limit will prevent very high currents, and the duration is very brief, limited by the SPS_Fail signal from the SPS comparator. Still it was partly this magic, and a desire for reliability that made CR402 a convention diode.

■ BB Issues

The use of UPS power for the battery voltage analog switch (U406) maybe wrong. UPS can be less than the battery voltage!
No, this looks ok. The voltage drop for the UPS is across a lightly loaded Schottky diode, so the difference is almost negligible, and anyway less than the threshold for U406’s input protection diodes. Also during power down the switch is always off, so performance demands are low. This should be checked in operation, but is probably ok.

Maybe CR402 should be a saturable rectifier? (This is explored a little in CR402’s write up.)

Integrated Circuit(s)

■ U400 LTC3401, 3MHz, 1A, DC-DC boost converter (DigiKey LTC3401EMS−ND5.50$ea)

The selection criterion include size, efficiency, complexity and frequency. This MSOP−10package can reach 90+% efficiency supplying 5V @ around 100mA output @ 1.5MHz (synchronized). With <1μA shutdown current, <1mA quiescent current, and potentially 1A output current while tolerating up to 6V input or output.

■ U401, 402 NC7SZ14, Single Schmitt inverter (DigiKey NC7SZ14M5CT−ND0.96$ea)

What is really better is a dual inverter in a SOT-6-style package (e.g. NC7WZ14), but digikey doesn’t have them.

These inverters snap-up the lowpass filtered PwrGood signal from U400 making it crisp for the power switch-over circuitry and providing some hysteresis. Two inverters are used to properly drive Q400 and provide a positive logic HAP_Good signal. Other single gate styles could be substituted but the 1.8V rating on the selected part is nice because of the potential low battery voltage operation.

■ U403 Single D−FFin SSOP8−P−0.50APkg. (DigiKey TC7W74FKCT−ND0.56$ ea)

Changed to flip−flop,was TC54VC4502ECB713, voltage monitor (DigiKey 158−2066−1−ND0.83$ea)

This threshold detector is chosen for low quiescent power, suitable output drive (active pull up) and compatible voltage range.

U403 does two essential things. It wakes the HAP up when SPS power is available and is shuts the HAP down when SPS power is down and it is commanded by the PowerDown signal.

Additionally it exhibits the parasitic effect of shutting down the HAP in the event of a HAP brown-out. This behavior could be construed as a feature.

Every time this circuit is looked at, a desire arises to modify it in some way. It seems that most of the functions of required are available elsewhere in the circuit. The essential properties of the circuit are
1) HAP is always powered up when SPS power is good
2) HAP is powered down cleanly when SPS power is absent and PowerDown is asserted.

It was the 2nd point that drove the choice of a low voltage reset. The reset is fed from two power sources, as long as neither one is good the HAP is powered down. To replicate this using the available logic the two obvious signals are SPS_Fail from the SPS and the HAP’s own HAP_Good / HAP_Good\ signals. By cases:

SPS_Fail ≡ HAP_wake\ == 0, HAP stays up
SPS_Fail == 1
HAP_off == 0 , HAP stays up

HAP_off == 1, HAP goes down
HAP down (HAP_Good\ == 1), HAP_off == X

If the pullup for LT3401’s PwrGood output pin was derived from the Shutdown\ pin while Shutdown\ was controlled weakly from HAP_off, and through a diode by ~SPS_Fail we would have:

While ~SPS_Fail, Shutdown\ == 1 by virtue of the diode. Assuming UPS voltage was in range, PwrGood would be high, pulled up via Shutdown\ . If HAP_off\ is low, Shutdown\ remains high due to the superiority of the diode over the resistor.

When SPS_Fail == 1, Shutdown\ remains high while, HAP_off\ (from the PIC) is high because of the resistor. If HAP_off\ goes low Shutdown\ will also go low, possibly delayed by a capacitor.

Once the LT3401 shuts down, the PwrGood line will go low, thus pulling the Shutdown\ input low.

This scheme seems to work. There is a question as to the relative values of the PwrGood pull-up and the resistor on the PIC’s Powerdown signal.

It seems cleaner to try a logic element.

Using a FF :
To cause power down, HAP_Good == 1 while a transition occurs. Reset by SPS_Fail low.
While SPS_Fail low, FF reset, so Shutdown\ == 1.
When SPS_Fail goes high, FF remains reset.
If D input, HAP_Good == 0 no amount of clocking can set the FF
If HAP_Good == 1, and the FF is clocked Low → High, then the HAP powers down, and will not come back up until SPS_Fail goes low.

Seems great.

■ U404 LTC1734, Lithium-Ion (Li+) Battery Charge controller (DigiKey LTC1734ES6−4.1−ND 2.75$ea)

Selection criterion include small package, small total parts count. The ability to properly charge the battery.

The ’1734 uses an external pass transistor, but is otherwise a reasonably complete system. The 4.1V version might be selected because it is adaptable to any Li chemistry, however the 4.2V version may be more appropriate???.

If the 4.1V version was selected, a resistor could be added in place of W400.

■ U405 TC7W53FK, 2 Channel Analog Multiplexer (DigiKey TC7W53FKCT−ND0.56$ea)

Toshiba TC7W53FK
SSOP8−P−0.5Apackage
(2,, 7)V Vcc Range
(0,, −6)V Vee Range
(, 55, 120)Ω R_DS(on) 25C @ 4.5V
>1GΩ typ. off resistance (but consider feedthrough capacitance)
150Ω Max @ −40C, 4.5V
150Ω Max @ 85C , 4.5V
±(, , 60)nA Leakage 25C
±600nA Max leakage @ 85C
(, 11, 20)pF input capacitance
(, 0.75, 2)pF feedthrough capacitance
200mW P_DMax (dissipation limited)
−40C- 85C operating temperature range

U405 is a single analog multiplexer. Powered from the UPS (battery) supply.

The multiplexer isolates the PIC from the battery voltage when the PIC is powered off. In fact this function maybe unnecessary, but it is very cheap because the mux also selects between the SPS_Fail and the Batt_Curr (Battery Current) signals. When the battery is under charge, the battery current is selected, otherwise the SPS_Fail signal is selected. If the HAP_Good\ signal is asserted, both signals are disconnected.

The Toshiba part is adequate in terms of resistance and operating voltage (it’s spec’d to run down to 2V).

■ U406 TC4S66F, single analog switch (DigiKey TC4S66FCT−ND0.68$ea)

Toshiba TC4S66F
SMV-5 package
18V V_DSS Max
(, 290, 950)Ω R_DS(on) 25C
>1GΩ typ. off resistance (but consider feedthrough capacitance)
800Ω Max @ −40C
1200Ω Max @ 85C
(, 0.1, 100)nA Leakage 25C
1μA Max leakage @ 85C
10pF typ. input capacitance
0.5pF typ. feedthrough capacitance
200mW P_DMax (dissipation limited)
−40C- 85C operating temperature range

U406 is a single analog switch. It is powered from the UPS (battery) supply. (In fact typical specs require 3V to operate. The minimum on UPS is approximately 2.2V. This should still work. All that is required is that U404, 405 remain off at low supply voltages, this should be checked???.

This switch isolate the PIC from the battery voltage when the PIC is powered off. Fairly high on resistance can be tolerated because the input impedance of the PIC's ADC is reasonably high. 1kΩ is definitely good enough.

■ U407 NC7SZ14, Single Schmitt inverter (DigiKey NC7SZ14M5CT−ND0.96$ea)

This could be a plain inverter, but a Schmitt was specified to reduce the number of different parts. Either way is ok.

Transistors(s)

■ Q400 See Q404

Turns HAP supply off.

■ Q401 See Q201

Q401 disconnects the feedback voltage divider on the boost supply when the HAP is powered down. If this was not done then the divider current would drain the battery. On the other hand, a very large resistance divider could be used, but that would result in stability problems.

The drain at the nominal battery voltage is about

```
3.6 / 150*^3 (* A *) // EngineeringForm
```

```
24. × 10^-6
```

Assuming 100mA as the intolerable drain

```
(100*^-3 / %) / 24 (* days *)
```

```
173.611
```

Admittedly, this is a long time. However the 150k divider may prove a problem and divider current might need to be increased. Also there's no apriori reason to assume the HAP would not be in storage for a year or more, and on that time scale the divider current could be a real problem.

■ Q402 Eliminated in HAP version 0.3

■ Q403 Zetex FCX717 SOT89 PNP (DigiKeyFCX717CT−ND0.87$ea)

Zetex FCX717
SOT89 package
12V V_DSS
(, 0.11, 0.15)V V_CE(Sat) Ic==1A, Ib==10mA
100nA Max collector cut off current
10A Max pulse current
300 round value for Hfe
500mA Ib (Abs. Max)
(−55, 150)C T_J range
1W Max Power on FR-4 (15×15×0.6)mm

This is the PNP series pass transistor required by U404.

The input voltage hovers near 4.75V. The cell voltage varies down to 2.5V. Assuming 420mA maximum charge current the dissipation is

```
(4.75 - 2.5) 0.420 (* Watts *)
```

```
0.945
```

To handle this power level a thermal resistance of 50°C/W would be desirable. 70C/W is about the max.

The SOT89 package probably can dump about 1W. This should be ok because the very low voltages during recharge are of short duration, and the ratings to handle the power are there.

The beta is fairly high on this unit, so be careful of overall loop stability.

■ Q404 IRF(2nd source) P-Channel MOSFET Si4435DY (DigiKey SI4435DY−ND1.50$ea)

International Rectifier (2nd source) Si4435DY, also Zetex
SO−8 package
30V V_DSS
19mV/C Break down temp.co. 25C, I_D==1mA
35mΩ R_DS(on) @ V_GS=4.5V
(, 40, 60)nC total gate charge @ V_GS==10V
2320pF typ. input capacitance
6.4A I_DMax (dissipation limited) V_GS==10V
50A Max pulse current
±20V V_GSMax
1.2V Max body diode forward voltage @ 25C, 2.5A, V_GS==0V
51ns Max body diode reverse recovery time
50nC Max body diode reverse recovery charge
150C T_JMax
50C/W Max θ_JA on standard FR-4 board, t < 5s.

Q404 functions as a saturable rectifier. When the SPS_Fail signal is false (i.e. SPS is good) the gate-source voltage on Q404 is zero or negative and Q404 acts as a normal diode, in this case blocking current from V_BK into the battery. If the V_BK voltage falls below V_BB for any reason the body diode of Q404 will conduct providing power to the HAP.

Theoretically the only way V_BK can fall below V_BB is if SPS fails, in this case SPS_Fail will be asserted and U407 will in turn provide gate voltage to Q404 decreasing the voltage drop to near zero.

If the SPS voltage suddenly rises the SPS_Fail signal will rapidly (μ-seconds) drop. During the switch over fairly large currents may flow into the battery (though limited by the SPS switch limit) but the duration is so short that no harm will be done.

For conservative design Q404 should be capable of dissipating the load current in the un-saturated state (~1/2A×0.7V≃400mW), though this could be pushed.

■ Q405,406 See Q201

Diode(s)

■ CR400 See CR200

CR400 is the "optional diode" from the LT3401 data sheet.

■ CR401 Renumbered to D401

■ CR402 Diodes Inc. Fast Si-Diode MURS120 (DigiKey MURS120DICT−ND0.60$ea)

Diodes Inc. MURS120, 1A Super-Fast rectifier

SMB	package
200V	V_R
1.5A	I_DMax (dissipation limited)
40A	Max pulse current
0.875V	Max forward voltage @ 25C, 1A
0.710V	Max forward voltage @ 150C, 1A
25ns	Max reverse recovery time
2.62mm tall	
(−65, 175)C T_J range	

Originally this diode was configured as a N-Fet saturable rectifier similar to Q404. This was not successful because the LTC3401 is spec'd at 120ns minimum on time. This implies a maximum input voltage of 4.07V to maintain constant frequency switching. In fact the slight discontinuous mode switching that results from excessive input voltage *might* not be a big problem. It's the uncertainty over the consequences of the added noise that drives us to eliminated it.

The 4.07V limit can be eased by raising the HAP output voltage to 5.2V (should work). The maximum input voltage rises to 4.24V. This works fine for the Li+ cell, which is never over 4.2V anyway, but the battery charger wants more like a 4.55V input (4.68V to be really happy). Of course limiting the input voltage to the battery charger also decreases power dissipation in the pass transistor, which is good.

As a 1st cut, try to make every one happy. Let the input to the battery charger be 4.78V (to allow for some slop). CR402 should drop a minimum of 0.54V at around 10mA, but be capable of passing the maximum switch current of U400 (1.6A, but only around 1/2A average) with an as low as possible additional voltage drop.

If the actual minimum drop is 0.64V there would be a little slack in meeting the 4.24V constraint. This should make everything happy.

The selection search, which will be brief, consists of finding a Si-diode (for the high initial voltage drop) that maintains a fairly low voltage drop into the 1A range.

The minimum operating voltage for the battery charger is 4.55V, assume 4.6V is more realistic. Let the nominal target for SPS voltage be 4.69V±2% == [4.60, 4.78]V

■ D400 Eliminated

1N4448HWS, see CR201.

Eliminated.

D400 allows the μP to activate the reset function, but not to countermand it. This de-glitches the PowerDown cycle.

■ D401 Eliminated

Same type as D200

The original purpose of D401 was to provide a signal from the SPS to the HAP allowing the HAP to wake up when the SPS was powered up. This function has been shifted to the SPS_Fail signal and D401 has been eliminated.

Inductor(s)

■ L400 Coiltronics MP2A−1R5(Same as L200), 1.5μH, See PM−4112.pdf

Coiltronics MP2A−1R5
1.5 μH nominal inductance
1.54 μH measured @ 100kHz, 0.25V RMS, 0.0A DC, ±20%
73 mΩ DC Ohms @ 20C typ.
2.02 A RMS current producing +40C temperature rise (excluding core loss)
3.22 A Saturation current defined by 30% loss of inductance. Measured at 20C
2.09 V·μs Volt·time product of 300kHz waveform which when applied across inductor produces core loss equal to 10% of power loss producing +40C temperature rise previously determined.
−40to +125C storage temperature range
−40to +85C operating temperature range
molybdenum permalloy core
rated to 500kHz ??? —I don’t see any reason for this.
Body Length exclusive of leads 5.88 mm, Length with leads 7.5 mm. Width 5.2 mm, Height 1.8 mm.

Manufacturer (Linear) recommends L>3/f [H]. Ripple constraints suggest:

$$L > \frac{Vimin\,(VoMAX - Vimim)}{f \times \Delta I \times VoMAX}\,[H]$$

Maximum available current == (max switch current −inductor ripple current/2) × efficiency / boost-ratio

Derivation (boost converter)

Equal volt-seconds implies

$$(1 - D)\,(Vo - Vi) == D\,Vi$$

$$Vo == \frac{Vi}{1 - D},\ D == 1 - \frac{Vi}{Vo}$$

The current increase through the inductor in one cycle is

$$\frac{Vi}{L} * Ton,\ \ Ton == T * D == \frac{1}{f} * D \Rightarrow \Delta I == \frac{D}{f}\,\frac{Vi}{L}$$

Average current is

$$Iavg == Ip - \Delta I / 2 == Ip - \frac{D\,Vi}{f\,L} \Big/ 2 == Ip - \frac{D\,Vi}{2\,L\,f}$$

This current is decreased by the same factor the voltage is increased, (1−D), less (by definition) the efficiency. We therefore get the maximum boost-current available as

```
Io == η (1 - D) (Ip - D Vi/(2 L f))
```

As a rule of thumb, the ripple current is typically 20−40%of the peak current

```
Io == η (1 - D) (Ip - 0.3 Ip / 2) (* Rule of Thumb *) // FullSimplify
```

```
Io == -0.85 (-1 + D) Ip η
```

Using the 3401 numbers, assuming 5.2V out, 4.25V in gets

```
% /. {η → 0.92, Ip → 1.6, D → 1 - 4.25/5.2}
```

```
Io == 1.02262
```

Of course we’ll actually get something less than this, but this is ball-park. Now find L by using the rule of thumb.

```
Ip3401 = 1.6 (* peak switch current limit for LTC3401 *);
```

```
Solve[ΔI == D Vi/(L f), L][[1, 1]] // FullSimplify
```

$$L \to \frac{D\,Vi}{f\,\Delta I}$$

```
% /. {D → 1 - Vi/Vo} /. {f → fnom, ΔI → 0.3 * Ip3401, Vi → 4.25, Vo → 5.2} //
 EngineeringForm
```

$$L \to 1.05143 \times 10^{-6}$$

So basically one micro-Henry. I_sat should be >1.6A, ferrite core, and for ~1/8W DC loss, we want ≲50mΩ .

Referring to the calculations for L200, it seems either MP2A-1R0 or MP2A-1R5 would be ok. To minimize inventory, the 1R5 should be selected but for fun run both.

The maximum inductor ripple current for 1μH is

```
VbkH = 4.25 ; VbkL = 3 ; VHAP = 5.2 ; (* HAP voltage limits *)
```

```
ΔI → D Vi/(L f) /. {D → 1 - Vi/Vo} /. {Vi → VbkL, Vo → VHAP, f → fnom, L → 1*^-6} // N
```

```
ΔI → 0.825
```

For 1.5μH

```
ΔI / 1.5 /. %
```

```
0.55
```

This is closer to the rule of thumb's 30%. The slightly increased DC resistance of the 1.5μH unit probably lowers the efficiency v. the 1μH inductor (probably the reduced AC core loss doesn't make up for the upped DC loss). Considering the advantage of reduced inventory, the 1.5μH unit seems the way to go.

Going back to the originally stated constraints

$$L > \frac{\text{VbkL}\,(\text{VHAP} - \text{VbkL})}{\text{fnom} \times 0.3\,\text{Ip3401} \times \text{VHAP}}$$

$$L > 1.71875 \times 10^{-6}$$

Close enough.

Note this inductor does not meet the IC manufacture's recommended value of ≥ 3/f[H] which would be 2μH in this case. Obviously it is close. I don't see why this should be a problem.

The volt × second product for this case is

$$\frac{D\,\text{Vi}}{f} == \left(1 - \frac{\text{Vi}}{\text{Vo}}\right) \frac{\text{Vi}}{f}$$

Maximize this

$$\text{Solve}\left[D\left[\left(1 - \frac{\text{Vi}}{\text{Vo}}\right) \text{Vi}, \text{Vi}\right] == 0, \text{Vi}\right][[1, 1]]$$

$$\text{Vi} \to \frac{\text{Vo}}{2}$$

In our case this means Vi is minimum

$$\left(1 - \frac{\text{Vi}}{\text{Vo}}\right) \frac{\text{Vi}}{f} \text{ /. \{Vi} \to \text{VbkL, Vo} \to \text{VHAP, f} \to \text{fnom\} // N // EngineeringForm}$$

$$825. \times 10^{-9}$$

This is quite small, which should mean AC losses are small.

Capacitors(s)

■ C400 22μF, 1206, X5R, Panasonic ECJ−3YB0J226M(DigiKey PCC2242CT−ND12.87$/10)

Panasonic ECJ−3YB0J226M(Same as C200)
22 μF nominal capacitance
6.3V WV DC
X5R dielectric
1206 case size
±20% tolerance
−35to +85C operating temperature range
Height 1.6 mm.

Output capacitor for a high frequency boost switching power supply. The dynamic properties of this capacitor affect the loop stability so a moderately stable X7R or X5R type is typical. A tantalum cap. will work fine here, but is probably bigger. The ceramics seem so much cooler. Voltage rating ≥ 6.3V.

Basic ripple formula

$$\text{Vpp} == \frac{\Delta \text{I Vi}}{\text{C Vo f}} \text{ /. } \{\Delta\text{I} \to 0.55,\ \text{Vpp} \to 5{*}{\wedge}{-3},\ \text{Vi} \to 4.25,\ \text{Vo} \to 5.2,\ \text{f} \to \text{fnom}\} \text{ // N}$$

$$0.005 == \frac{2.92188 \times 10^{-7}}{\text{C}}$$

```
Solve[%, C][[1, 1]] // EngineeringForm
```

$$\text{C} \to 58.4375 \times 10^{-6}$$

I think this is p-p, so i get a free factor of two, and anyway what i got is a 22μF cap, so that’s what to use.

■ C401 10μF, 1206, Y5V, Panasonic ECJ−3YF1A106Z(DigiKey PCC1894CT−ND4.91$/10)

Panasonic ECJ−3YF1A106Z
10 μF nominal capacitance
10V WV DC
Y5V dielectric
1206 case size
+80/−20% tolerance
−25to +85C operating temperature range
Height 1.6 mm.

This is an input capacitor for a high frequency boost switching power supply. The standard choice is a 1−5 μF ceramic capacitor. The Y5V or similar dielectrics are suitable. Voltage rating should be at least 6.3V.

The ripple in the capacitor is

$$\text{V} == \text{Q} / \text{C}$$

Try 2.2μF

```
ΔIrms
------- /. {ΔIrms → 0.55, C → 2.2*^-6}
fnom C

0.1625
```

I like to stick w/the 10μF, can't use the SPS units since they are 1210 case size.

■ C402 Eliminated

Old value was 0.1μF, eliminated in HAP v0.5 .

See C202.

C402 introduces delay into the \shutdown circuit of U400. The largest convenient value is 0.1μF. With on order 1kΩ source resistance, the time constant is

```
0.1*^-6 * 1*^3 (* seconds *) // EngineeringForm

100. × 10^-6
```

The time required to drop the voltage on C402 1V assuming the maximum input current of U403 (4.2μA) is

```
1 * 0.1*^-6
----------- (* seconds *) // EngineeringForm
  4.2*^-6

23.8095 × 10^-3
```

This seems about right.

■ C403 10μF, 10V, Y5V, Panasonic ECJ–3VF1C225Z(DigiKey PCC1898CT–ND4.17$/10)

Changed to 10μF, see also C201

Panasonic ECJ−3VF1C225Z
2.2 μF nominal capacitance
16V WV DC
Y5V dielectric
1206 case size
+80/−20% tolerance
−25to +85C operating temperature range
Height 0.85 mm.

Output filter for HAP supply. C400 is the main filter. C403 is positioned after switch Q400. Assuming on order 1nH of stray inductance, the resonant value of C403 is about.

```
Solve[Fdrive == 1/Sqrt[L C], C][[1, 1]] // FullSimplify
```

$$C \to \frac{1}{\mathrm{Fdrive}^2\, L}$$

```
% /. {Fdrive → 1.53846*^6, L → 1*^-9} // N // EngineeringForm
```

```
C → 422.501 × 10^-6
```

So resonance won't be a problem.

The largest practical value in 1206 size is 2.2μF, which is the value chosen. — This is dumb. Use a 10μF, we have them, and they're better anyway. (The voltage drops to 10V, but so what.)

■ C404 Any 25+V 0.01μF ceramic cap in 1206 package (DigiKey PCC104BCTCT−ND 0.35$ea)

Panasonic ECJ−4VB1H104K
0.01 μF nominal capacitance
50V WV DC
X7R dielectric
1206 case size
±10% tolerance
−55to +125C operating temperature range
Height 0.6 mm.

C404 deglitches the PwrGood signal from U400. The value is non-critical, and really to be determined.

■ C405 See C412

Forget this, use the same as C412.

Panasonic ECS−T1AY106R
10 μF nominal capacitance
10V WV DC
Tantalum dielectric
EIA−A case size
±20% @20C tolerance
CV/100 A DC leakage @ 20C, (or 0.5μA, whichever is greater)
<0.06 Tan-δ
−55to +125C operating temperature range
Height 1.6 mm.

This capacitor provides high frequency bypassing for U404's feedback loop. The manufacturer's recommended value is 10μF, so that's what we've used.

■ C406 Eliminated

No need to de-glitch in v 0.5, eliminated.

See C404

C406 is solely to deglitch and delay the shutdown signal to Q401. It is redundant in the graph-sense with C410. C406 may be unnecessary, whether this is true is best determined by observation of the hardware. A preliminary specification of 0.01μF has been given anyway.

■ C407-C409

Frequency compensation component. See R403 for explanation.

■ C410 Eliminated

Was concerned with power-down timing on old circuit, eliminated in v0.5 .

See C404

See C406

■ C411

Beware, C411 was reassigned in HAP version 0.3. Frequency compensation component. See R403 for explanation.

■ C412 10μF, 10V, Y5V

Use the HAP unit, see C401

Panasonic ECJ−3YF1A106Z
10 μF nominal capacitance
10V WV DC
Y5V dielectric
1206 case size
+80/−20% tolerance
−25to +85C operating temperature range
Height 1.6 mm.

This could be a ceramic (Y5V) unit or a Tantalum, which is best? See C401.

This is the input bypass cap for the battery charger. The manufacturer recommends 1-10μF, we have no compelling reason not to use a 10μF unit as chosen here.

Resistors(s)

■ R400 eliminated

Panasonic P−1.00KFCT−ND
1.00kΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

R400 has been eliminated in a re-design

R400 serves as a wire-OR’ing element for input voltage from the SPS and UPS node. If SPS voltage is zero, the input to U403 will be equal to UPS, less the voltage drop across R400. The supply current for U403 (TC54VC45) is 4.2μA Max. The minimum anticipated value for the SPS voltage is 4.60V. The maximum voltage threshold for U403 is 4.59V. 10mV drop at 4.2μA gives

```
10*^−3 / 4.2*^−6
```

```
2380.95
```

Choosing a value of 1kΩ will assure a sufficiently low voltage drop while maintaining acceptably low current during override operations. R408 must be able to pull down U403 to the low voltage threshold. The minimum threshold voltage is 4.5*0.98 == 4.4V. Assume an input voltage of 5.2V, the maximum resistance value for R408 is

```
Solve[r408/(r400 + r408) == Vo/Vi, r408][[1, 1]] // FullSimplify
```

$$r408 \to \frac{r400\ Vo}{Vi - Vo}$$

```
% /. {r400 → 1000, Vo → 4.4, Vi → 5.2} // N
```

r408 → 5500.

For simplicity, choose R408 == 1kΩ.

■ R401 36.5kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−36.5KFCT−ND 1.17$/10)

Panasonic P−36.5KFCT−ND
36.5kΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

R401, 402 form the feedback divider for the HAP boost converter. The reference voltage is 1.25±0.03V. The desired output voltage is 5.2V. The error amp input current is under 50nA, so a very high divider impedance (~2MΩ, value limited by capacitive parasitics) could be used, but noise considerations suggest working closer to the 100kΩ level. Somewhat arbitrarily set the target divider impedance to 150kΩ then the values are

```
Solve[{Vfb == Vo * R401 / Rt, Rt == R401 + R402}, {R401, R402}][[1]] // FullSimplify
```

$$\left\{R402 \to Rt - \frac{Rt\ Vfb}{Vo},\ R401 \to \frac{Rt\ Vfb}{Vo}\right\}$$

```
% /. {Rt → 150*^3, Vfb → 1.25, Vo → 5.2} // N // EngineeringForm
```

$$\{R402 \to 113.942 \times 10^3,\ R401 \to 36.0577 \times 10^3\}$$

Try 115k and 36.5k

```
Vfb → Vo * R401 / (R401 + R402) /. {Vo → 5.2, R401 → 36.5*^3, R402 → 115*^3} // N
```

Vfb → 1.25281

```
Vfb / 1.25 /. % // N
```

```
1.00224
```

Seems OK.

■ R402 115kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−115KFCT−ND 1.17$/10)

Panasonic P−115KFCT−ND
115kΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

See R401

■ R403 ???kΩ , Any ???% or better film resistor in 1206 package

R403, along with C408,409 form the frequency compensation component of the HAP switching supply. C407 also affects the mix by providing a feed-forward component to the output sensing voltage divider.

The notes for R206 also cover frequency compensation. The basic references an-19 p.48-50 and reference an76 still apply.

The DC loop gain is the product of modulator gain and the error amp gain. The error amp gain is around 2000, while the modulator gain is ≈2 Vin/Iout.

The Output filter pole (current mode control) is basically 1/(2π Rout Cout)Hz. Rout is just 1/2*Vout/Iout (??? explain this). Thus the output pole break frequency is

$$\text{f0pole} \to \frac{1}{2\,\pi\,\text{Ro}\,\text{Co}} \text{ /. Ro} \to \frac{1}{2}\,\frac{\text{Vo}}{\text{Io}} \text{ (* Hz *) // FullSimplify}$$

$$\text{f0pole} \to \frac{\text{Io}}{\text{Co}\,\pi\,\text{Vo}}$$

The output zero is

$$\text{f0zero} \to \frac{1}{2\,\pi\,\text{Resr}\,\text{Co}}$$

The boost regulator also has a right-half-plane-zero (RHP0), which can lead to instability (no response through the feedback path). The frequency is given by

$$\text{fRHP0} \to \frac{\text{Vi}^2\,\text{Ro}}{2\,\pi\,\text{L}\,\text{Vo}^2} \text{ (* Hz *)}$$

This frequency can be low if Ro (the load resistance) has a low value. The typical strategy is to roll off the loop gain prior to the RHP0.

Following the nomenclature from the note under R206, the compensation capacitor and series resistor (Cc, Rc) and the filter compensation capacitor (Cf), the compensation network introduces two poles and a zero as before.

The leading pole (assuming Cc >> Cf, which is the usual case) is formed by Cc and the modulator input resistance, which is around 20MΩ

```
fp1 → 1/(2 π Ri Cc) (* Hz *) /. {Ri → 20*^6, Cc → 100*^-12} // N
```

```
fp1 → 79.5775
```

The 1st pole is quite close to DC, even with a fairly low value of Cc (here assumed to be 100pF).

Here is the remaining pole and zero

$$fp2 \to \frac{1}{2\,\pi\,Rc\,Cf}$$

$$fz1 \to \frac{1}{2\,\pi\,Rc\,Cc}$$

As before, the input voltage divider is bypassed by a speedup capacitor (CA) and a noise filtering, and counter balancing capacitor (CB).

The break frequencies produced by CA and CB are at

$$fA \to \frac{1}{2\,\pi\,RA\,CA},\ fB \to \frac{1}{2\,\pi\,RB\,CB}$$

See also R206

■ R404 47.5kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−47.5KFCT−ND 1.17$/10)

Panasonic P−47.5KFCT−ND
47.5kΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

This is the pullup resistor for the PwrGood output signal. The value should be >>R400. As a compromise try 47k.

■ R405 27.4kΩ , Any 2% or better film resistor in 1206 package (DigiKey P−27.4KFCT−ND 1.17$/10)

Panasonic P−27.4KFCT−ND
27.4kΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

R404 sets the center frequency of U400’s oscillator. For synchronization, the center frequency should be about 30% less than the desired synchronization frequency, according to the manufacturer.

```
Solve[Fosc == 3*^10 / Rosc, Rosc][[1, 1]] // FullSimplify
```

$$Rosc \to \frac{30000000000}{Fosc}$$

```
% /. Fosc → (10*^6 / 6.5) (0.70) // N // EngineeringForm
```

$$Rosc \to 27.8571 \times 10^3$$

27k would be fine, 28k is good too. 5% is fine, 2% perhaps a bit better.

■ R406 470Ω , Any 5% or better film resistor in 0805 package (DigiKey P−470CCT−ND 0.91$/10)

R406 limits the current available at J400. 470Ω is low enough to override R407+R408 and cause manual shutdown. This is convenient, but dangerous. In flight J400 cannot be accidentally shorted. A different value of R406 could be selected to allow manual shutdown unless OK'd by the PIC.

■ R407 10kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−10.0KFCT−ND 1.17$/10)

R407 keeps the HAP_off signal inactive even if the PIC is tri-stated, etc. 10k is a good value. It's sufficiently large compared to R408, yet should be small enough to keep the signal from false tripping.

The function of R407 changed in HAP v0.5, the old description follows :

Ideally, R407 would be dropped. It's function is to assure the \Shutdown signal to U400 is asserted when U403 is unpowered. U403 operates down to about 0.7V. It is possible that U400 would operate down to 0.5V so a window exists for unintended operation. This seems unlikely so no value for R407 was selected. If used the value should be fairly high if possible to minimize operating dissipation.

■ R408 1.00kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−1.00KFCT−ND 1.17$/10)

The role of R408 as of HAP v0.5 is to limit the current in the event of a manual power down of the HAP. 1k is a good value because it is low enough to prevent accidental tripping or slow edges, but large enough to limit the override current to a small value.

■ R409 20.0kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−20.0KFCT−ND 1.17$/10)

See RT400.

■ R410 32.4kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−32.4KFCT−ND 1.17$/10)

See R411.

■ R411 3.57kΩ , Any 1% or better film resistor in 0805 package (DigiKey P−3.57KCCT−ND 0.91$/10)

Panasonic P−3.57KCCT−ND
3.57kΩ nominal resistance
150V WV DC
1/10W Dissipation @ 70°C
0805 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

R410, 411 set the fast and trickle battery charging rates.

Here is the basic formula

```
Ibat == 1500 / Rset
```

The fast charge current is not very critical. The nominal value is 420mA, but if this was a problem for the supply is could be reduce to 300mA. For now assume 420mA is ok, then the programming resistor is

```
Rset → 1500 / 0.42 // N
```

```
Rset → 3571.43
```

Since this is a standard value, go for it.

The nominal trickle charge rate is C/10 or in this case 42mA. This can successfully be any similar figure but sticking to definite C/x values should give a charge−time profile that is independent of battery capacity, which might be good.

These assumptions enable a calculation for R410 to be made

```
Rset * (10 - 1) /. %
```

```
32142.9
```

Pick 32.4k as the next standard value.

```
1500 / (3570 + 32.4*^3) // N // EngineeringForm
```

41.7014×10^{-3}

■ R412−413 47.5kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−47.5KFCT−ND1.17$/10)

See RT406.

These are pull downs for the battery charge control lines.

■ R414 (Removed) 2MΩ , Any 5% or better film resistor in 1206 package (DigiKey P−2.00MFCT−ND1.17$/10)

It’s desirable to monitor the status of the input supply to the HAP. This should be done with a minimum of input pins on the PIC. The program current pin (PROG) on the Li+ charger (U404) can serve four functions. Connecting a resistor between the pin and ground can set the peak charging current. The true charging current can be determined by reading the voltage at the PROG pin. If the voltage rises above 2.25V(Max) the battery charge circuitry is disabled. The PROG pin is pulled up by a 3μA current source internal to U404. However, if the input voltage drops to zero, essentially zero current will flow from the PROG pin, and the action of R414 will be to pull the voltage at PROG down to zero volts. Thus, a reading of zero volts on the PROG pin indicates a main power failure except during charging, when it could indicate charge termination.
In the case of main power failure during charge, the loss of battery current could be mistaken for charge termination. For the anticipated algorithm, this is not the case, but even if a misidentification of the cause of a zero current reading was made, the ordinary response would be to terminate charging. If the mains were available, the PROG voltage would immediately increase. If the mains failed, the PROG voltage would remain zero, breaking the ambiguity.

The value of R414 must be large enough to allow the voltage to rise to the shutdown threshold under the action of the nominal 3μA current source. The minimum current is 1.5μA

```
1.5*^-6 * 2*^6
```

```
3.
```

3V is fine as the minimum threshold is 2.25V.

R414 must be small enough to pull the PROG pin low during "Sleep" mode, when the input voltage is ~ 0V. The leakage current in sleep mode is not specified on the data sheet. The maximum leakage for the whole chip over temperature is given as ±1μA.

Miscellaneous

■ B400 Sanyo UF611948, rectangular Li-Ion cell

Sanyo UF611948 (See Sanyo_Rectang_UF611948.pdf)
420mAHr nominal capacity
3.7V nominal voltage
4.2V maximum charging voltage
420mA maximum rated charging current
2.5Hr expected recharge time for fully discharged cell
273WHr/liter volumetric energy density
135WHr/kg gravimetric energy density
(0, 40)C allowable charging ambient temperature
(−20,60)C allowable discharge ambient temperature
(−20,50)C storage temperature range
48×19.5×6.1mm Case dimensions
11.5g cell mass

The main question here is how big a battery is required. A 1Hr mission durability is probably adequate. Assuming the low current requirements we have estimated are correct (<100mA) the capacity of this battery is clearly sufficient. The AA type are easier to come by but they seem harder to mount??? and have slightly worse gravimetric energy density???.

The 420mA recharge current may be somewhat high for the SPS target of 500mA (but maybe not) but if this was a problem, the current could probably be safely reduced 50% without complications except longer recharge times. Alternately the SPS inductor value could be raised, raising available output current.

By the way, the chemistry of the Sanyo cell is approximately
Lithium Cobaltate ($LiC_{???} O_2$), ρ==4.95, mp==1130C, black odorless powder
Graphite electrode (C) (as opposed to Coke), ρ==2.09~2.2, black odorless powder
Organic Solvent
Palyvynilidene difluoride (PVdF)

■ J400 2pin, Shrouded 2mm through-hole pin header (DigiKey H2094−ND0.24$ea)

This removable jumper allows the HAP to be powered down manually. The state can be checked with the push button on the PIC.

In HAP v0.5 J400 has these properties:
1) manually powers down HAP when connected while HAP is up.
2) dissipates no power if left connected in the unpowered state
3) HAP can be powered up by software with jumper connected (TP: i think)
4) software can detect if jumper is connected after a power up
5) HAP cannot be powered down by software with jumper connected
6) dissipation when powered up with jumper connected depends on the state of the PIC HAP_Off signal.

Flight software should report no-go if jumper connected following power-up.

■ RT400 Thermometrics NHQM103R10, negative temperature coefficient thermistor (DigiKey 235−1116−1−ND1.75$ea/10)

Thermometrics NHQM103R10 (DigiKey claims NHQM103B375T10)
10kΩ nominal resistance @ 25C
±10% tolerance
3750 β (25C, 85C)±200 K
0805 case size
<5s thermal time constant
1.5mW/K thermal dissipation factor
(−40,125)C operating temperature range
1.3 mm tall

This is the same unit used on the LV1b FC and IMU. The consistency is regarded as an advantage. Possibly an IC unit with linear response should be substituted.

Based on LV1b design, the series resistance should be 20kΩ.

■ W400 (eliminated due to space during layup)

This wire jumper allows a resistor to be inserted in case the 4.1V version of U404 is used on a battery with a nominal voltage limit of 4.2V. Possibly this jumper should be eliminated.

Pyrotechnic Sub-Node High Voltage Igniter Trigger (PYRO) (2100)

PYRO Goals

Reliable recovery (minimize point failure, provide redundancy, verifiable status)
Safety
Low mass
Small size

PYRO Specs

Mechanical interlock
Oxral igniters

PYRO Design notes

The PYRO system activates upto four Oxral "electric match" pyrotechnic igniters. This is done using a capacitive discharge system.

The advantages of the capacitive discharge system include
- Low current requirement from the HAP supply
- "Safe" state distinctively different from the "Armed" state
- Natural interlock by shorting capacitor

Issues

Should probably reconfigure the HV supply for flyback mode.

Could eliminate the separate supply for the floating drivers.

Connector(s)

Phoenix 2.54mm "Euroblock" screw terminals

Phoenix Contact 1725672
2.54mm pitch
4 circuits
6A current rating / contact
125V AC/DC voltage rating (UL)
20 - 30 AWG accepted wire sizes (UL)
(8.5 x 10.62 x 6.2)mm HxWxD
Green color

Andrew likes ’em

Integrated Circuit(s)

■ U2100 LT1310, 4MHz, 1.5A, DC-DC boost converter

The selection criterion include size, efficiency, complexity and frequency. This MSOP−10package can reach 85% efficiency and output upto 35V. It can be synchronized to above 4MHz, has under 1μA shutdown current, about 12mA quiescent current, and potentially above 1A output current while tolerating a 3-18V input range.

Particularly constraining here is the requirement of high voltage output, to minimize the number of stages in the voltage multiplier, and high frequency operation.

I really want to use the LT1310. The difficulty is that the maximum duty cycle is only about 80%, this is less than what’s required to get 100V off a tripled output. An LT1930 reaches 84%, which is sufficient, but this part cannot be synchronized.

If it’s allowable to resort to typical characteristics, the LT1310 will function marginally with a tripler at 100V. This might be a problem at low temperatures. If i can assume Schottky diodes at 0.25V, the implied duty cycle is

```
(100 + 5 * 0.25) / 3 (* voltage required *)
```

```
33.75
```

```
1 - VHAP / % (* required duty cycle *)
```

```
0.845926
```

This is almost the same duty cycle as would be achieved with silicon diodes.

```
(100 + 5 * 0.7) / 3
```

```
34.5
```

```
1 - VHAP / %
```

```
0.849275
```

I think this problem can be solved by using the discontinuous conduction mode, which will lower the effective duty cycle for a given output voltage. Alternatively a tapped inductor could be used (see an−19 p.43). Since a tapped inductor is a U-wind-it solution, it’s currently not the preferred method.

■ U2101−2104 IR2118S, Single channel inverting high voltage, high side driver

International Rectifier IR2118S
<600V maximum offset voltage
10-20V drive voltage range
<200ns propagation delay
SO8 package
8.4V UV lockout center voltage
(,120, 580)μA quiescent current
−40to +125C operating temperature range

This integrated circuit accepts a 10−20V power supply and drives a high side mosfet switch up to 600V. Unfortunately, the input is referenced to the power supply, and requires a minimum 6V input for a logic "1" at a 15V supply voltage. A more convenient driver may be available, this should be checked.

Selection criterion include quiescent current, voltage and drive capacity, and parts count.

Intersil ISL6801(HighSideDriver−fn9087.pdf)is close, but only good to 120V. The advantage of the ’6801 is its 5V logic level input, but intuitively skimping on high voltage rating is unwise.

■ U2105 SN74AHC1G126DBVR (DigiKey 296−8746−1−ND0.56$ea)

Texas Instruments SN74AHC1G126DBVR
5.5V maximum supply voltage
8mA recommended output current
<10ns propagation delay
SOT-23-5 package

Buffer drives a voltage tripler to supply boost power for the high side drivers.

Transistor(s)

■ Q2100 BSS−131SOT-23-3 N-FET (DigiKey BSS131INCT−ND0.36$ea@10)

Infineon BSS−131
SOT-23-3 package
240V V_DSS
16Ω R_DS(on)Max V_GS==10V, 25C
26Ω R_DS(on)Max V_GS==4.5V, 25C
400mA I_DMax (pulse)
100mA I_DMax (dissipation limited)
360mW Recommended dissipation

Zetex ZVN3320
SOT-23-3 package
200V V_DSS
25Ω R_DS(on)Max V_GS==10V, 25C
1000Ω R_DS(on)Max V_GS==3V, 25C
1A I_DMax (pulse)
60mA I_DMax (dissipation limited)
330mW Recommended dissipation

International Rectifier IRLL110 (Too big use ZVN3320 above)
SOT−223package
100V V_DSS
12mV/C Break down temp.co. 25C, I_D==1mA
760mΩ R_DS(on), V_GS=4V
<6.1nC total gate charge @ V_GS==5V
250pF typ. input capacitance
1.5A I_DMax (dissipation limited)

12A	Max pulse current
±10V	V_GSMax
2.5V	Max body diode forward voltage @ 25C, 1.5A, V_GS==0V
130ns	Max body diode reverse recovery time
6500nC	Max body diode reverse recovery charge
50mJ	Max pulse avalanche Energy
0.31mJ	Repetitive avalanche Energy
150C	T_JMax
60C/W	Max θ_JA on standard 1"sq. copper clad board, t < 1s.

This transistor passes current from the high voltage cap (C2100) to the voltage measurement divider. When on this allows the voltage on C2100 to be measured by the microprocessor. Also the measurement divider serves to discharge C2100 when that is desired. The current requirements for Q2100 are miniscule, but the voltage rating must be at least 100V.

I think 5V should get R_DS down to ~100Ω, but the divider should be designed to accept 1kΩ with rated accuracy. The PIC ADC is 10bit. The maximum threshold is 3V. Compress 100V to 2V @ 5/1024V resolution. The voltage quanta is

$$\frac{100}{2} * \frac{5}{1024} \text{ V // N}$$

```
0.244141 V
```

About 1/4V. The 5V test voltage is thus 20 quanta. At best, if say 5V was scaled to 150V, 5V would read

$$5 \Big/ \left(\frac{150}{5} * \frac{5}{1024}\right) \text{ // N}$$

```
34.1333
```

Preliminarily, it looks like all is working.

Suppose we wish to discharge in 3 seconds to 95% ⇒ $\tau \approx$ 1s. For 10μF ⇒ R==100kΩ . Initial current ~1mA. This should be ok w/ V_GS > 2V using IRLL110. Map 100V to 2V.

$$\text{Solve}\left[\left\{\text{r2108} + \text{r2107} == \text{100*^3}, \frac{100}{\text{r2108} + \text{r2107}} \text{r2107} == 2\right\}, \{\text{r2108}, \text{r2107}\}\right]$$

```
{{r2108 → 98000, r2107 → 2000}}
```

■ Q2101-2104 IRLR3410 D-PAK N-FET (DigiKey IRLR3410−ND1.16$ea/10)

International Rectifier IRLR3410
TO-252AA (D-PAK) package
100V V_DSS
122mV/C Break down temp.co. 25C, I_D==1mA
105mΩ R_DS(on), V_GS=10V
<34nC total gate charge @ V_GS==5V
800pF typ. input capacitance
17 I_DMax (dissipation limited)
60A Max pulse current
±16V V_GSMax
1.3V Max body diode forward voltage @ 25C, 9A, V_GS==0V
210ns Max body diode reverse recovery time
1100nC Max body diode reverse recovery charge
150mJ Max pulse avalanche Energy
7.9mJ Repetitive avalanche Energy
175C T_JMax
50C/W Max θ_JA on standard 1"sq. copper clad board, t < 1s.

These transistors discharge the high voltage capacitor triggering the associated pyrotechnic charge. They require sufficient voltage and current ratings while maintaining low R_DSON. Unfortunately this make them huge.

■ Q2105-2108 See Q201

Diode(s)

■ D2100 BAV21W (DigiKey BAV21WDICT−ND0.32$ea/10)

See D2101.

Must withstand the full high voltage. Use the same as the high side drivers.

■ D2101-2104 BAV21W (DigiKey BAV21WDICT−ND0.32$ea/10)

Diodes Inc. BAV21W
SOD-323 package, 2 pin type possible to shoot a trace
200V V_R
0.1μA Max reverse current @ V_R
250mW Max power (assume 25C ???)
1V Max forward voltage @ 25C, 100mA
50ns Max reverse recovery time
5pF Max capacitance
(−40, 125)C T_J range (maybe a bit more)

These diodes provide current to the high side driver bootstrap circuits. They must withstand the full high voltage and be fast recovery. The current requirement is quite low, only 150mA (See C2103). The reverse recovery time should be under 100ns.

■ D2105-2109 1N4448HWS (DigiKey 1N4448HWSDICT−ND0.30$ea/10)

See CR201.

High voltage tripler diodes. Max voltage is ~35V, average current pretty low. These will work for the high side driver supply tripler too.

■ D2110-2114 1N4448HWS (DigiKey 1N4448HWSDICT−ND0.30$ea/10)

See D2105.

■ D2115 Hi−brightRed LED, 0603 (DigiKey 160−1473−1−ND2.82$/10)

LITEON LTST−S270CKT
0603(narrow) package
clear package color
660nm peak λ (Red)
1.8V Vf @ 20mA
16mC Peak Brightness
130° viewing angle
AlGaAs LED

Indicator lamp. Shows current flow into HV supply.

Inductor(s)

■ L2100 Coiltronics MP2A−1R5,1.5μH, See PM−4112.pdf

Coiltronics MP2A−1R5
1.5 μH nominal inductance
1.54 μH measured @ 100kHz, 0.25V RMS, 0.0A DC, ±20%
73 mΩ DC Ohms @ 20C typ.
2.02 A RMS current producing +40C temperature rise (excluding core loss)
3.22 A Saturation current defined by 30% loss of inductance. Measured at 20C
2.09 V·μs Volt·time product of 300kHz waveform which when applied across inductor produces core loss equal to 10% of power loss producing +40C temperature rise previously determined.
−40to +125C storage temperature range
−40to +85C operating temperature range
molybdenum permalloy core
rated to 500kHz ??? —I don't see any reason for this.
Body Length exclusive of leads 5.88 mm, Length with leads 7.5 mm. Width 5.2 mm, Height 1.8 mm.

The LT1310 has a switch current limit of 2.8A Max. Assuming a 33V output voltage and a 5.2V input voltage the following seem to apply.

Nominal output voltage is up to 100V, assuming silicon diodes in tripler configuration

```
Vboost = (100 + 5 * 0.7) / 3 (* input to voltage multiplier *)
```

```
34.5
```

```
D == 1 - VHAP / Vboost
```

```
D == 0.849275
```

The data sheet only guarantees 78% maximum duty cycle over temperature at 1.5MHz

```
 VHAP
────────
1 - 0.78
```

```
23.6364
```

That’s sort of pathetic. If the layout was modified to include a doubler on the HAP output, then even assuming a 2 diode drop loss, the resulting duty cycle is only

```
D == 1 - (2 VHAP - 3 * 0.7) / Vboost
```

```
D == 0.75942
```

On the other hand, changing to a quadrupler on the pyro output and trying for 100V means 7 diode drops

```
(100 + 7 * 0.7) / 4 (* required voltage using quadrupler *)
```

```
26.225
```

```
D == 1 - VHAP / %
```

```
D == 0.801716
```

This fails, barely. If schottky diodes were used in the quadrupler

```
(100 + 7 * 0.3) / 4
```

```
25.525
```

```
D == 1 - VHAP / %
```

```
D == 0.796278
```

The design still will not meet specs.

It therefore appears that the HAP should be modified for doubler operation. This is complicated by the need to completely shut off the HAP supply. Doing this would require another high-side switch, which is a bad deal.

Consider discontinuous mode operation.

The current during the on time builds up to

$$\Delta I == \frac{Vi}{L} Ton$$

By definition, in discontinuous mode, the off time exceeds the time required for the inductor current to fall to zero. This time is given by volt · seconds balance

$$\text{Solve}\left[\text{Toffmin}\,\frac{\text{Vo}-\text{Vi}}{\text{L}} == \frac{\text{Vi}}{\text{L}}\,\text{Ton},\ \text{Toffmin}\right][[1,1]]\ //\ \text{FullSimplify}$$

$$\text{Toffmin} \to \frac{\text{Ton Vi}}{-\text{Vi}+\text{Vo}}$$

$$\text{Toffmin} == \text{Ton}\,\frac{\text{Vi}}{\text{Vo}-\text{Vi}}$$

The duty cycle concept ceases to be meaningful in discontinuous mode. The average current into the load is given by an integral (ignoring efficiency effects)

$$\text{Iavg} == \text{Io} == \text{f}\int \text{Ioff}\, dt$$

During off time the current is falling linearly

$$\text{Io} == \text{f}\,\frac{\text{I toffmin}}{2} == \frac{\text{f}}{2}\,\frac{\text{Vi}}{\text{L}}\,\text{Ton} * \text{Ton}\,\frac{\text{Vi}}{\text{Vo}-\text{Vi}} == \text{Ton}^2\,\frac{\text{f}}{2\,\text{L}}\,\frac{\text{Vi}^2}{\text{Vo}-\text{Vi}}$$

For a boost-mode supply, an−19gives these formula (note the 3rd formula is mis-stated in an−19and the 3rd equation is also plain wrong, see derivations below)

Smallest inductor that results in continuous mode operation

$$\text{Lcont} \geq \frac{\text{Vi}^2\,(\text{Vo}-\text{Vi})}{2\,\text{f Io Vo}^2}$$

The minimum inductance to supply a given current in discontinuous mode (ignoring efficiency)

$$\text{Ldis} \geq \frac{2\,\text{Io}\,(\text{Vo}-\text{Vi})}{\text{f Ip}^2}$$

Peak inductor current (Vf is diode forward voltage drop, R is switch on-resistance) (for continuous mode operation)

$$\text{Ip} == \text{Io}\,\frac{(\text{Vo}+\text{Vf})-(\text{Io Vo R}/\text{Vi})}{\text{Vi}-\text{Io Vo R}/\text{Vi}} + \frac{\text{Vi}\,(\text{Vo}-\text{Vi})}{2\,\text{f L Vo}}\ \text{(* WRONG! *)}$$

Here are some derivations of the above

$$\text{Lcont} == \frac{\text{Vi}^2\,(\text{Vo}-\text{Vi})}{2\,\text{f Io Vo}^2} == \frac{\text{Vi}^2\,\text{D}}{2\,\text{f Io Vo}} == \text{Ton}\,\frac{\text{Vi}^2}{2\,\text{Io Vo}}$$

$$2\,\text{Io}\,\frac{\text{Vo}}{\text{Vi}} == \text{Ton}\,\frac{\text{Vi}}{\text{Lcont}} == \Delta\text{I}$$

So the mode is just barely continuous when the current change during Ton is equal to twice the average current output times the voltage ratio. This derives the 1st equation.

Now find Io in terms of L_dis and frequency. Using what was done above

$$Io == f\,\frac{Ip\ toffmin}{2} == \frac{f}{2}\,Ip\ Ton\,\frac{Vi}{Vo - Vi}$$

In discontinuous mode, Ip==ΔI, so the ΔI formula can eliminate Ton

$$\text{Solve}\left[Io == \frac{f}{2}\,Ip\left(\frac{Ip\ L}{Vi}\right)\frac{Vi}{Vo - Vi},\ L\right][[1,1]]\ //\ \text{FullSimplify}$$

$$L \to -\frac{2\ Io\ (Vi - Vo)}{f\ Ip^2}$$

Now the 3rd formula

$$Ip == Io\,\frac{(Vo + Vf) - (Io\ Vo\ R / Vi)}{Vi - Io\ Vo\ R / Vi} + \frac{Vi\ (Vo - Vi)}{2\ f\ L\ Vo}$$
$$== Io\,\frac{Vo + Vf - Von}{Vi - Von} + Ton\,\frac{Vi}{2\ L}$$

The 1st term is the transformer relation for the average current corrected for diode drop and finite switch on voltage. The 1st term is wrong. The numerator should not contain the Von voltage. The second term is the ΔI contribution. The correct equation is

$$Ip == Io\,\frac{Vo + Vf}{Vi - Von} + \Delta I / 2 == Io\,\frac{Vo + Vf}{Vi - Io\ Vo\ R / Vi} + \frac{Vi\ (Vo - Vi)}{2\ f\ L\ Vo}$$

There’s no decent reason i see why the same inductor can’t be spec’d here as before.

The continuous mode current threshold is

$$Io \geq \frac{Vi^2\ (Vo - Vi)}{2\ f\ L\ Vo^2}\ /.\ \{Vi \to VHAP,\ Vo \to Vboost,\ f \to fnom,\ L \to 1.5*\wedge{-6}\}$$

$$Io \geq 0.144221$$

ΔI continuous mode

$$\Delta I == \frac{Vi\ (Vo - Vi)}{2\ f\ L\ Vo}\ /.\ \{Vi \to VHAP,\ Vo \to Vboost,\ f \to fnom,\ L \to 1.5*\wedge{-6}\}$$

$$\Delta I == 0.95685$$

This is within the LT1310 limit of (1.5, 2.1, 2.8)A

Capacitors(s)

■ C2100 American Capacitor Corp. DW2D106K

Electrocube 230B1B106K, <Electrocube 230 Series−MetallizedPolyester.pdf>
10 μ nominal capacitance
100V WV DC
Oval shape
<1.25% dissipation factor @ 1000Hz
metalized polyester dielectric
0.46 x 0.63 x 1.15 " case size
±10% tolerance
0.032" lead diameter (AWG #20)
−55to +85C operating temperature range (voltage derated to 125C)

Electrocube 230C1B106K, <Electrocube 230 Series−MetallizedPolyester.pdf>
10 μ nominal capacitance
100V WV DC
Flatish shape
<1.25% dissipation factor @ 1000Hz
metalized polyester dielectric
0.27 x 0.66 x 1.4 " case size
±10% tolerance
0.0253" lead diameter (AWG #22)
−55to +85C operating temperature range (voltage derated to 125C)

American Capacitor Corp. DW2D106K, <ACC_MetallizedMylar−DData.pdf>
10 μ nominal capacitance
100V WV DC
<1.25% dissipation factor @ 1000Hz
metalized polyester dielectric
0.46 x 0.63 x 1.17 " case size
±10% tolerance
0.032" lead diameter (AWG #20)
−55to +85C operating temperature range (voltage derated to 125C)

For details see DesignNotes???

Volume of 230B, 230C

```
{0.46 * 0.63 * 1.15, 0.27 * 0.66 * 1.4} // N
```

```
{0.33327, 0.24948}
```

Measured dimensions of 230B, 230C

$$b = \begin{pmatrix} .420 & .426 & .418 & .433 \\ .576 & .572 & .587 & .575 \\ 1.163 & 1.163 & 1.169 & 1.174 \end{pmatrix};$$

$$c = \begin{pmatrix} .269 & .274 & .272 & .264 \\ .703 & .697 & .696 & .700 \\ 1.386 & 1.383 & 1.391 & 1.386 \end{pmatrix};$$

```
Times @@ b (* Volume *)
```

```
{0.281353, 0.283391, 0.286833, 0.292297}
```

```
Times @@ c (* Volume *)
```

```
{0.262102, 0.264123, 0.263333, 0.256133}
```

So, interestingly, the ovals are a little smaller than advertised, the flats are a bit larger. Still the flats are smaller overall.

Mass of 4 together (full leads)

```
bm = 24.05 / 4
```

```
6.0125
```

```
cm = 22.05 g / 4
```

```
5.5125 g
```

For "C" flats, recommend 1.7" spacing on holes. For "B" ovals, recommend 1.5" spacing.

■ C2101 10μF, >6.3V

See C401.

This is an input capacitor for a high frequency boost switching power supply. The standard choice is a 1−5 μF ceramic capacitor. The Y5V or similar dielectrics are suitable. Voltage rating should be at least 6.3V. Choose the same type as used in the HAP to minimize inventory.

■ C2102, 2104, 2106, 2108 Any 25+V 0.1μF ceramic cap in 1206 package (DigiKey PCC104BCTCT−ND0.35$ea) (2108 is 0805 size)

See C202.

These capacitors are simple bypassing caps. 0.1μF values are adequate. Voltage rating should be at least 25V.

■ C2103, 2105, 2107, 2109 Any 25+V 0.33μF ceramic cap in 1206 package, Panasonic ECJ−3VB1E334K (DigiKey PCC1889CT−ND0.25$ea)

See C204.

These capacitors are bootstrap components for the high side switch. Ref. dt98−2.pdf.

Let Qbs be the charge delivered by the bootstrap capacitor per cycle, Qs be the total gate charge required by the mosfet, IqbsMax be the maximum quiescent current used in the bootstrap circuit, Qls be the charge required by the level shift circuitry, Icbsleak be the leakage current in the bootstrap capacitor, and f the operating frequency. The basic bootstrap equation is

```
Qbs == 2 Qs + IqbsMax / f + Qls + Icbsleak / f
```

Qs is specified by the mosfet manufacturer, in the case of the IRLR3410 this is approximately

```
34*^-9 + 5 * 800*^-12 (* nC *) // N // EngineeringForm
```

```
38. × 10^-9
```

Call this 40nC. The IR2117 datasheet gives the maximum bootstrap quiescent current as 240μA. And the level shifting charge is 5nC. Assume a low leakage capacitor, thus

```
Qbs → 2 Qs + IqbsMax / f + Qls + Icbsleak / f
   (* nC *) /. {f → 10*^6 / 6.5, Qs → 40*^-9, IqbsMax → 240*^-6,
    Qls → 5*^-9, Icbsleak → 0} // N // EngineeringForm
```

```
Qbs → 85.156 × 10^-9
```

Assume the minimum input voltage is 12V, and the maximum diode voltage is 0.7V.

The manufacturer recommends a capacitor value equal to

```
30 * Qbs / (Vcc - Vdiode) /. % /. {Vcc → 12, Vdiode → 0.7} // N // EngineeringForm
```

```
226.078 × 10^-9
```

Since we've already spec'd some 0.33μF parts, we may use them.

Incidentally, the diode forward current is

```
Qbs * f /. %% /. f → 10*^6 / 6.5 (* A *)  // N  // EngineeringForm
```

```
131.009 × 10^-3
```

■ C2110 KEMET C1206C224K5RACTU (DigiKey 399−1251−1−ND 4.18$/10)

KEMET C1206C224K5RACTU
0.22 μ nominal capacitance
50V WV DC
X7R dielectric
1206 case size
±10% tolerance
−35to +85C operating temperature range
Height 1.5 mm.

This is the 1st output capacitor for the pyrotechnic voltage tripler. Ripple can be very high compared to typical applications, but stability suggests a reasonable value should be used. The dynamic properties of this capacitor affect the loop stability so a moderately stable X7R or X5R type is typical. Voltage rating ≥ 35V.

Basic ripple formula

$$\text{Vpp} == \frac{\Delta\text{I Vi}}{\text{C Vo f}} \text{ /. \{}\Delta\text{I} \to 0.144,\ \text{Vpp} \to 50\text{*\^{}-3},\ \text{Vi} \to 5.2,\ \text{Vo} \to 34.5,\ \text{f} \to \text{fnom\} // N}$$

$$0.05 == \frac{1.41078 \times 10^{-8}}{C}$$

```
Solve[%, C][[1, 1]] // EngineeringForm
```

```
C → 282.157 × 10^-9
```

This is a very modest value

■ C2111 - C2113 See 2110

These are the remaining ladder caps of the pyro voltage tripler. I don’t see why the same 0.22μF values can’t be used. Recharge time should still be sub-second.

■ C2114 Any 25+V 0.01μF ceramic cap in 1206 package (DigiKey PCC104BCT−ND0.35$ea)

Filter capacitor for Pyro voltage measurement. 0.01μF gives ~200μS time constant.

■ C2115 See C2116

See C204. — Superceeded, see C2116

This is the output capacitor for the high side driver supply. The maximum steady load is

```
(240 + 340) 1*^-6 * 4 (* A *) // N // EngineeringForm
```

```
2.32 × 10^-3
```

At the operating frequency this represents a small charge of

```
% / f /. f → 10*^6 / 6.5 (* nC *) // N // EngineeringForm
```

```
1.508 × 10^-9
```

If a the ripple is set to 100mV, the required capacitance is

```
% / .1 (* F *) // N // EngineeringForm
```

```
15.08 × 10^-9
```

This is tiny. Triggering the driver will pull 40nC

```
15 * 40 / 1.5 (* nF *) // N
```

```
400.
```

Since the 0.33μF caps have already been spec’d keep using ’em.

■ C2116-2120 16V 0.47μF 0805 package, Panasonic ECJ−2YB1C474K (DigiKey PCC1818CT−ND0.2.45$/10)

Used to be: Any 25+V 0.33μF ceramic cap in 1206 package, Panasonic ECJ−3VB1E334K (DigiKey PCC1889CT−ND0.25$ea)

See C2115.

Thinking about changing to 0805. May need to go 16V or 0.22uF (DKp.648)

I fact had to go to 0805. The easy thing to do is to drop to 16V. This should be ok. This allows increasing the value to 0.47μF.

■ C2121-2124 Frequency compensation components, see R2110.

■ C2125 1500pF ceramic cap in 1206 package (DigiKey PCC152BCT−ND1.17$/10)

Panasonic ECU−V1H152KBM

1500pF	nominal capacitance
50V	WV DC
X7R	dielectric
1206	case size
±10%	tolerance
−55to +125C	operating temperature range

Height 0.6 mm.

LPF for PLL. Manufacturer recommended value is 1500pF.

■ C2126 Not Present

LPF for PLL. Should not be required.

■ C2127 100pF NPO 0805 (DigiKey PCC1964CT−ND2.14$/10)

Panasonic ECJ−2VC2A101J

100pF	nominal capacitance
100V	WV DC
NPO	dielectric
0805	case size
±5%	tolerance
−55to +125C	operating temperature range

Height 0.85 mm.

Sets the center frequency of the PLL. For synchronization, set ~33% higher than target frequency.

$$\frac{9+3}{9}\,\frac{10^{*}\verb|^|6}{6.5}\ (*\ \mathrm{Hz}\ *)\ //\ N\ //\ \mathrm{EngineeringForm}$$

$$2.05128\times 10^{6}$$

The recommended capacitor for this frequency from the datasheet is 100pF. This value is cheap enough in NPO.

Resistors(s)

■ R2100 47.5kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−47.5KFCT−ND 1.17$/10)

See R406.

Pull down for Pyro-Charge input signal. Pretty much standardized on 47k.

■ R2101-2104 Any 5% or better film resistor in 1206 package of around or about 10Ω's (guess). (DigiKey P−10.0FCT−ND 1.17$/10)

These resistors slow down the transitions of the mosfet switches. Typical values are between 3.3Ω and 47Ω. Typical indications for there use are excessive supply bounce or VHF oscillation in the mosfet gate. Just for fun we have spec'd 10Ω's.

■ R2105 39.2kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−39.2KFCT−ND 1.17$/10)

Panasonic P−36.5KFCT−ND
39.2kΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

R2105, 2106 form the feedback divider for the PYRO boost converter. The reference voltage is 1.255±0.016V. The desired output voltage is 33V. The error amp input current is under 150nA, so a moderately high divider impedance ~500kΩ, could be used, but noise considerations suggest working closer to the 100kΩ level. Somewhat arbitrarily set the target divider impedance to 150kΩ then the values are

```
Solve[{Vfb == Vo * R2105 / (R2105 + R2106), 1 / Rt == 1 / R2105 + 1 / R2106},
   {R2105, R2106}][[1]] // FullSimplify
```

$$\left\{R2106 \to \frac{Rt\ Vo}{Vfb},\ R2105 \to \frac{Rt\ Vo}{-Vfb + Vo}\right\}$$

```
% /. {Rt → 150*^3, Vfb → 1.255, Vo → 33} // N // EngineeringForm
```

$$\left\{R2106 \to 3.94422 \times 10^{6},\ R2105 \to 155.93 \times 10^{3}\right\}$$

Scale R2106 to 1MΩ

```
R2106 / 1*^6 /. %
```

```
3.94422
```

```
{R2106, R2105} / %% /. % // N // EngineeringForm
```

```
{1. × 10^6, 39.5338 × 10^3}
```

Try 1M and 39.2k

```
Vfb → Vo * R2105 / (R2105 + R2106) /. {Vo → 33, R2105 → 39.2*^3, R2106 → 1*^6} // N
```

```
Vfb → 1.2448
```

```
Vfb / 1.255 /. % // N
```

```
0.991875
```

Seems OK, though not great.

■ R2106 1.00MΩ, Any 1% or better film resistor in 1206 package (DigiKey P−1.00MFCT−ND 1.17$/10)

Panasonic P−1.00MFCT−ND
1.00MΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

See R2105

■ R2107 2.00kΩ, Any 1% or better film resistor in 1206 package (DigiKey P−2.00KFCT−ND 1.17$/10)

Panasonic P−22.1KFCT−ND
22.1kΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

Calculation superseded, see Q2100

R2107 along with R2108 form the high voltage measurement divider. For convenience at the μP, it’s desirable to keep the impedance less than 20kΩ. We wish to compress a zero to ~150V range down to 0−5.2V. Perhaps the most natural division is into 1/2 volt quanta (full scale 128V). Here are the equations.

```
Solve[{1/Rt == 1/RA + 1/RB, Vo/Vi == RA/(RA + RB)}, {RA, RB}][[1]] //
 FullSimplify
```

$$\left\{RB \to \frac{Rt\,Vi}{Vo},\ RA \to \frac{Rt\,Vi}{Vi - Vo}\right\}$$

```
% /. {Rt -> 20*^3, Vi -> 128, Vo -> 5.2} // N // EngineeringForm
```

$$\left\{RB \to 492.308 \times 10^{3},\ RA \to 20.8469 \times 10^{3}\right\}$$

An OK compromise would be RA → 22.1k , RB → 523k, this is open to discussion.

■ R2108 97.6kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−97.6KFCT−ND 1.17$/10)

See R2107.

■ R2109 4.99kΩ , Any 1% or better film resistor in 0805 package (DigiKey P−4.99KCCT−ND 1.17$/10)

Panasonic P−4.99KCCT−ND
4.99kΩ nominal resistance
150V WV DC
1/10W Dissipation @ 70°C
0805 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

This resistor charges the high voltage discharge capacitor to the supply voltage (5.2V nominal). The minimum permissible value is that resulting in the no-fire igniter current, which is about 100mA. The maximum value is that which results in a time constant of ~1/2s. These two constraints are

```
R > 5.2 / .1 // N (* Ohms *)
```

$R > 52.$

```
R < 0.5 / 10*^-6 // N (* Ohms *)
```

$R < 50000.$

5000Ω seems a convenient figure.

■ R2110 frequency compensation component

This resistor, along with C2121-2124 form the feedback compensation network. As explained for the HAP, the component values are to be determined.

■ R2111 3.01kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−3.01KFCT−ND 1.17$/10)

Low pass filter for PLL. Manufacturer recommended value.

■ R2112-2120 47.5kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−47.5KFCT−ND 1.17$/10) (R2119 needs 0805 package.)

See R406.

Pull down for high side driver input signal. Pretty much standardized on 47k.

■ R2121 10kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−10.0KFCT−ND 1.17$/10)

Gate resistor for Q2100 Pyro measurement divider switch.

The gate resistor isn’t a necessity, but by slowing the transitions it eliminates any unwanted high frequency effects. Also in the event of oxide failure, the gate resistor will limit the current to the microprocessor. 10k isn’t too big to be slow, and will limit the peak current to around 10mA.

DTMF Demodulator for 2m radio (DTMF) (2200)

Connector(s)

■ CM2200 2mm pitch 2x5 pinheader Sullins PRPN052PAEN (DigiKey S2210−05−ND1$ea)

Sullins straight 2mm pitch pinheader
2mm pitch
10 circuits, double row
1A current rating / pin
50V AC/DC voltage rating
500VAC withstanding voltage, 1 minute
20mΩ contact resistance Max
4mm pin length on top
2.8mm pin length on bottom
2mm thickness of plastic housing
Polyester, UL94V−0,Brass w/selective gold/tin plating

This is the connector from the recovery node to the 2m board. It is double pinned for redundancy. A 2mm pinheader was chosen. It is effectively locked by the mass of the 2m board above it. The intention is to use a plain ribbon connector w/IDC from the 2m board to this header.

A 5 signal connector was selected. 4 signals are required. An extra ground for DTMF signal shielding has been added.

Integrated Circuit(s)

■ U2200 HT9170D, DTMF Receiver

A single chip dual tone decoder. Requires an external 3.579545 MHz crystal or ceramic resonator.

Holtek HT9170D
2.5 - 5.5V operating range
(, 3, 7)mA operating current
(, 10, 25)μA standby current
SOP18 package
3.579545±0.004 MHz Input frequency range
±1.5% frequency acceptance range
±3.5% frequency rejection range
(−29,,−6)dBmInput signal range — dBm are referenced to 1mW, The conversion to voltage assumes (we think) 600Ω’s

Solve for power

```
Solve[dBm == 10 * Log[10, P / P0], P][[1, 1]]
```

$$P \to 10^{dBm/10}\ P0$$

Find the voltage

```
Solve[Evaluate[P == V^2 / R /. %], V]
```

$\{\{V \to -10^{dBm/20}\sqrt{P0}\sqrt{R}\},\{V \to 10^{dBm/20}\sqrt{P0}\sqrt{R}\}\}$

Assuming 600Ω's find the voltage

```
%[[2, 1]] /. {R → 600, P0 → 1*^-3} // N // FullSimplify
```

$V \to 0.774597\, e^{0.115129\, dBm}$

```
vdBm[dBm_] := 0.7745966692414833` e^(0.1151292546497023` dBm)
```

```
vdBm[-29]
```

0.0274837

```
vdBm[-6]
```

0.388218

So the correct voltage range for the input (GS pin) is ~ 50 - 300 mV, a good target would be 100mV.

Capacitors(s)

■ C2200, C2201 1.8nF, Panasonic ECJ−2VC1H182J(DigiKey PCC182BCT−ND1.17$/10)

Panasonic ECU−V1H182KBM(not for new design)
1.8 nF nominal capacitance
50V WV DC
X7R dielectric
1206 case size
±10% tolerance
−55to +125 C operating temperature range
Height 0.6 mm

Part of the DTMF input active filter. These caps should have equal value. NPO 1206 are not digikey-able; 0805's are available in NPO, but layout needs the 1206 size.

For Calculations see Diff−BP−FilterDesign

■ C2202 10nF, Panasonic ECU−V1H1103KBM(DigiKey PCC103BCT−ND1.47$/10)

Panasonic ECU−V1H1103KBM(not for new design)
10 nF nominal capacitance
50V WV DC
X7R dielectric
1206 case size
±10% tolerance
−55to +125 C operating temperature range
Height 0.6 mm

Part of the DTMF input active filter. C2200 is the capacitor for the passive low pass pre-filter.

For Calculations see Diff−BP−FilterDesign

■ C2203 eliminated

The original idea was to have a differential amplifier, but the noise didn't seem to justify the complexity.

■ C2204 Any 25+V 0.1μF ceramic cap in 0805 package (DigiKey PCC1840CT−ND1.61$/10)

See C202

C2204 along with R2204 sets the time constant for the tone detector. The value chosen is straight off the data sheet.

■ C2205, C2206 27pF, Panasonic ECU−V1H270JCM(DigiKey PCC270CCT−ND1.21$/10)

Possible oops, HT9170 has $6.\overline{6}$ pF of internal capacitance (see LV2−DTMF−0.2.fig).

Panasonic ECU−V1H270JCM(not for new design)
27 pF nominal capacitance
50V WV DC
NPO dielectric
1206 case size
±5% tolerance
−55to +125 C operating temperature range
Height 0.6 mm

These capacitors form part of the resonator for the TV crystal oscillator. The crystal is spec'd at 17pF. Figure 3pF stray capacitance. The target is 28pF. 27pF are available. This should give

```
27 / 2 + 3 // N
```

```
16.5
```

Just fine

■ C2207 10μF, >6.3V

See C401

Low pass filter cap for RSSI analog input.

Assume 10μF, given R2207 == 1kΩ the break frequency is

```
1 / (10*^-6 * 1*^3) / (2 π) // N (* Hz *)
```

```
15.9155
```

This is fine. Use the HAP input capacitor ’cuz we got ’em.

■ C2208, C2209 Any 25+V 0.1μF ceramic cap in 1206 package

These are bypass caps for the DTMF decoder, and the 2m board power run. Could sub 0.33μF if we have them

Resistors(s)

■ R2200 750kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−750KFCT−ND 1.17$/10)

See R406.

This is the feedback resistor for the input opamp. It is also part of the input active filter.

For Calculations see Diff−BP−FilterDesign

■ R2201 100kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−100KFCT−ND 1.17$/10)

R2201 is part of the input network of the opamp. We want to do 3 things with the opamp network. 1) Provide a fixed gain. 2) Block DC components, 3) Filter high frequency AC components.

DTMF covers a frequency range from 697 - 1633 Hz.

For Calculations see Diff−BP−FilterDesign

■ R2202 9.31kΩ , Any 1% or better film resistor in 1206 package (DigiKey P−9.31KFCT−ND 1.17$/10)

Panasonic P−9.31KFCT−ND
9.31kΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

R2202, along with C2202, form the low end of the input pre-filter low pass voltage divider.

For Calculations see Diff−BP−FilterDesign

■ R2203 1kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−1.00KFCT−ND 1.17$/10)

Panasonic P−1.00KFCT−ND
1.00KΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

R2203 merely connects Vp to Vref on the input opamp of the DTMF decoder. THis could be just a wire, but a resistor seems to be more spiffy.

■ R2204 300kΩ, Any 5% or better film resistor in 0805 package (DigiKey P−300KCCT−ND 0.91$/10)

R2204 along with C2204 sets the time constant for the tone detector. The value chosen is straight off the data sheet.

■ R2205 1kΩ , Any 5% or better film resistor in 0805 package (DigiKey P−1.00KCCT−ND 0.91$/10)

See R2203

This resistor enables the decoding of DTMF codes A−D.It could be a zero ohm jumper. A 1k seems spiffier.

■ R2206 1kΩ , Any 5% or better film resistor in 0805 package (DigiKey P−1.00KCCT−ND 0.91$/10)

See R2203

This resistor disables the decoding of DTMF codes A−D.It could be a zero ohm jumper. A 1k seems spiffier.

■ R2207 1kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−1.00KFCT−ND 1.17$/10)

See R2203

R2207 is a series current limiting resistor. The RSSI signal comes from the 2m board. The recovery node must be protected from any weird off-board events. 1k seems a good compromise, and won’t impair ADC accuracy (PIC data sheet sez 2.5kΩ ok).

Miscellaneous

■ X2200 3.579545MHz parallel resonant crystal (DigiKey X184−ND0.98$ea)

ECS ECS−35−17−4DN (industrial temperature range)
17 pF target load capacitance
7pF shunt capacitance
±100ppm accuracy over temperature
±30ppm accuracy at 25°C
±5ppm aging, first year
HC−49/US case size
−40to +85 C operating temperature range
Fundamental mode

Required crystal for DTMF decoder oscillator

PIC processor (PIC) (300)

Connector(s)

■ CF300 JST−08R−JED

JST <www.jst−mfg.com>08R−JED,straight through female (receptacle, straight-through)
1.25mm pitch
8 circuits, double row
1A current rating / pin
50V AC/DC voltage rating
500VAC withstanding voltage, 1 minute
(20, 40)mΩ contact resistance Max (initial, after environmental testing)
300MΩ insulation resistance
(4.6 x 5.95 x 6.15)mm HxWxD
Glass filled nylon 66, Brass w/copper undercoat, tin plated

See CM200

This connector was intended as a de-bug / programming connector. However it is used in the recovery node to control the PLL on the 2m board.

Integrated Circuit(s)

■ U300 PIC18F458

Microcontroller

■ U301 PCA82C250

Physical CAN interface circuit

Diode(s)

■ D300 Fast diode (DigiKey 1N4148WCT−ND3.0$/10)

Diodes Inc. 1N4148
SOD-123 package, 2 pin type possible to shoot a trace
75V V_R
2.5μA Max reverse current @ V_R @ 25C
400mW Max power (assume 25C at leads)
0.855V Max forward voltage @ 25C, 10mA
4ns Max reverse recovery time
2pF Max capacitance
(−55, 125)C T_J range

Allows pullup, but blocks programming voltage.

■ D301 Hi−brightRed LED, 0603 (DigiKey 160−1473−1−ND2.82$/10)

LITEON LTST−S270CKT
0603(narrow) package
clear package color
660nm peak λ (Red)
1.8V Vf @ 20mA
16mC Peak Brightness
130° viewing angle
AlGaAs LED

"Error" lamp from PIC, also confirms button press.

Capacitors(s)

■ C300, C301 27pF, NPO, 1206 case

See C2205, C2206

Resonating capacitors for the PIC oscillator

■ C302-304 0.1μF bypass cap

See C202

0.1μF seems a reasonable value

Resistors(s)

■ R300 1kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−1.00KFCT−ND 1.17$/10)

See R2203.

This is a pull-up for the debug / programming connector. Why this value ???

■ R301 Not used

■ R302 750Ω , Any 5% or better film resistor in 1206 package (DigiKey P−750FCT−ND 1.17$/10)

Panasonic P−750KFCT−ND
750Ω nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

R302 is part of the omni-button 2003. Say 5mA for the PIC line, 5mA on the button. The pull down can be the standard 47kΩ .

Assume 1.5V for the LED

```
R302 == (5 - 1.5) / 5*^-3 (* Ohms *)
```

```
R302 == 700.
```

Since the Vf is probably a little high, use 750Ω (1k would work too).

■ R303 47.5kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−47.5KFCT−ND 1.17$/10)

See R404

Pulldown for logic level input.

■ R304 1.82kΩ , Any 5% or better film resistor in 1206 package (DigiKey P−1.82KFCT−ND 1.17$/10)

Panasonic P−1.82KFCT−ND
1.82KΩ nominal resistance
200V WV DC
1206 case size
±1% tolerance
±100ppm temp.co.
−55to +125C operating temperature range
Height 0.6 mm.
thick film composition

This is the omni-button pullup. Bumped the value to 1.8k because want the led to be noticeably darker if the PIC is holding the line down.

Miscellaneous

■ X300 10MHz parrallel resonant crystal (DigiKey X443−ND0.8$ea)

ECS ECS−100−18−4
18 pF target load capacitance
7pF shunt capacitance
±50ppm accuracy over temperature??
±30ppm accuracy at 25°C
±5ppm aging, first year
HC−49/US case size
−10to +70 C operating temperature range
Fundamental mode

■ SW300 6mm x 3.5mm, 160g (DigiKey P8058S−ND0.77$/ea)

160g chosen to prevent false triggers. 250g available. Not sealed.