Introduction

The rising demand for portable, battery powered electronics such as handheld phones, notebook PCs, games, etc. brings with it a growing interest by designers in battery chargers and ac adapters. Rechargeable secondary type batteries have become important components in portable electronics. New, lightweight, and higher capacity secondary batteries are coming onto the market, with nickel cadmium, nickel metal hydrogen, and lithium-ion batteries being the most widely used in portable electronics. Because they are lightweight and have a relatively high energy density, lithium-ion batteries in particular are especially popular for higher end portable electronics. However, they differ from nickel cadmium and nickel metal hydrogen types in requiring the use of both constant current and constant voltage charging controls.

This application note examines such a lithium-ion battery charging circuit: a switched mode power supply (SMPS) built around a Fairchild Power Switch (FPS) IC. Table 1 lists Fairchild’s ICs applicable to SMPS chargers and shows the various output power capabilities available. The universal input (85-265 Vac) charger/ac-adapter presented in this application note uses the KA5M0265R Fairchild Power Switch, the block diagram of which is shown in Figure 1.

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<th>Frequency</th>
<th>Pin(max)</th>
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<td>50 kHz</td>
<td>10 W</td>
<td>8DIP</td>
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<td>25 W</td>
<td>TO-220FP-4L</td>
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</tbody>
</table>
1. Charger/AC Adapter for Lithium Ion Batteries

Figure 2 shows a KA5M0265R based lithium-ion battery charging circuit (SMPS) having constant voltage, constant current operating characteristics. Among the topics discussed in this application note are constant voltage, constant current control, design of the primary power section, protection, and design considerations.

1.1 Constant current, constant voltage control circuitry

A simple and accurate constant current, constant voltage control using the Fairchild LM393 comparator is shown in Figure 3. The comparator’s output terminal is open...
collector configuration. With the two comparator outputs directly connected, the lower output voltage is followed, so constant current, constant voltage control is possible without adding extra separate components.

**Figure 3.** Wired OR of the LM393 comparator output.

There are two types of control circuits, depending on whether the current sensing resistor is at the ground terminal or at the output terminal. With the sensing resistor in the ground terminal as in Figure 4, the output voltage changes with current due to the changing voltage drop across the sensing resistor. Hence, reducing the resistor value reduces the output voltage change. For well regulated current control, however, the resistor value cannot be reduced blindly. In general, it is best to decide on the appropriate sensing resistance by testing between approximately 0.10 - 0.15W. This control circuit configuration is relatively simple and low cost. The second type of control circuit with the sensing resistor at the output terminal is shown in Figure 5. The output voltage does not change with current, so the output is regulated more accurately. This circuit configuration, however, does require additional components.

**Figure 4.** Constant current control with the sensing resistor (R220//R221) at the ground terminal.

1.2 Output current control with sensing resistor at the ground terminal

1.2 Output current control with sensing resistor at the ground terminal

Figure 6 is a repeat of most of Figure 4, but shows current flows. Table 2 lists the component values. The circuit allows the output current to be controlled simply by the choice of the value of R211. The value of R211 shown in Table 2 sets the output current (Io) to 1A.

**Figure 6.** Constant current control with one end of the sensing resistor at ground.
Here’s how the circuit works. The voltage divider formed by R213 and R214 across the 2.5V reference voltage puts 1V on the comparator’s positive input, Pin 3. The control loop adjusts its negative input terminal, Pin 2, to 1V also. If the comparator input bias current is ignored (assumed to be zero), the current flowing through R212 is 1 mA (1.5V/1.5kΩ). Therefore, 1mA also flows through R211. This makes the voltage drop across R211 0.9V (1mA x 900Ω). The remaining 0.1V appears across the 0.1W sensing resistor (R220 in shunt with R221), thereby setting the output current to 1 A (0.1V/0.1W). Note the ease of current control: simply changing R211 to 800Ω, for example, puts a 0.8V drop across R211 and 0.2V across the sensing resistor, thereby changing the output current to 2A.

Table 2. Parts list for Figure 6.

<table>
<thead>
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<th>Part</th>
<th>Value</th>
<th>Part</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>R220</td>
<td>0.2 Ω</td>
<td>R214</td>
<td>1.0 kΩ</td>
</tr>
<tr>
<td>R221</td>
<td>0.2 Ω</td>
<td>C201</td>
<td>0.1 μF</td>
</tr>
<tr>
<td>R211</td>
<td>900 Ω</td>
<td>Vref</td>
<td>2.5 V (KA431, LM431)</td>
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<tr>
<td>R212</td>
<td>1.5 kΩ</td>
<td>IC202A</td>
<td>KA393 (LM393)</td>
</tr>
<tr>
<td>R213</td>
<td>1.5 kΩ</td>
<td></td>
<td></td>
</tr>
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</table>

However, since components are usually not perfect, variations particularly in the voltage reference, the comparator’s offset voltage and bias current, and the resistor values will cause deviations from the expected current value. These factors must be considered when designing. For instance, assuming 0.1V across the sensing resistor, a 5mV comparator offset voltage for feedback control will produce a 5% error in output current. Typically, a circuit such as Figure 6 when configured with 1% resistors and an LM393A will exhibit a maximum error range within about ±7%. Therefore when it is designed for a 1A constant current output, the actual output obtained will be between 0.93A and 1.07A.

1.3 Output current control with sensing resistor at the output terminal

Figure 7 is a repeat of most of Figure 5 and shows current flows. Table 3 lists the components shown. Again, the circuit allows the output current to be controlled simply by the choice of the value of R211. The value of R211 shown in Table 3 sets the output current (Iₒ) to 1A. Here’s how the circuit of Figure 7 works. Op amp IC202A’s positive input, pin 3, is tied directly to the reference voltage, 2.5V. Hence IC202A’s negative input, pin 2, is at the same potential due to the control loop. Thus the drop across R213 is also 2.5V and the current flowing through it is 1mA (2.5V/2.5kΩ). Ignoring IC202A’s input bias current and Q1’s base current, the 1mA current through R213 also flows, via Q1, through R211. So the voltage drop across R211 is controlled to 0.1V (1mA x 100Ω). Assuming IC201A’s two input pin potentials to ground are equal, and ignoring its input bias current, the drop across sensing resistor R220//R221 must also be 0.1V. Hence the output current (Iₒ) is controlled to 1A (0.1V / 0.1W). Again, note the ease of current control: simply changing R211 to 200Ω, for example, sets a 0.2V drop across it and also across the 0.1W sensing resistor, thereby controlling the output current to 2A.
Again, however, components are usually not perfect and the considerations apply as in Figure 6 above.

1.4 Protection against low voltage and short circuited battery loads

The load for the charger is a battery, and a battery’s terminal voltage varies with the battery’s charge condition; furthermore, if the battery is at the end of its life, or faulty, it may short circuit. In other words, at some point depending upon the battery (i.e., load) condition, it will be necessary to initiate current control protection circuitry to protect the charger. Figures 8 and 9 exhibit different control characteristics because of the behavior of the LM393 comparator’s $V_{cc}$ supply in each case.

For proper comparator operation, $V_{cc}$ must be 3V or more. In Figure 8, however, the comparator’s $V_{cc}$ is derived directly from the load voltage. Therefore, as the load voltage drops, so does the comparator’s $V_{cc}$. At the instant the load voltage becomes less than 3V the control loop breaks and the constant current control does not operate. Hence the output current cannot be controlled, and it increases. In the ultimate case of a shorted battery, an extremely large load current will flow. In Figure 9, in contrast, the comparator’s $V_{cc}$ is separated from the load voltage. The transformer winding ratio is designed such that the control bias voltage stays above 3V, even if the battery shorts and thereby allowing good constant current control and preventing an abrupt rise in output current. In summary, for reliable constant current control, it is necessary to design in such a way that the comparator’s control bias voltage remains above 3V regardless of the load conditions.

### Table 3. Parts list for Figure 7

<table>
<thead>
<tr>
<th>Part</th>
<th>Value</th>
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<tbody>
<tr>
<td>R220</td>
<td>0.2 Ω</td>
<td>C201</td>
<td>0.1 µF</td>
</tr>
<tr>
<td>R221</td>
<td>0.2 Ω</td>
<td>Vref</td>
<td>2.5 V (KA431, LM431)</td>
</tr>
<tr>
<td>R211</td>
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<td>IC201</td>
<td>KA358A (LM358A)</td>
</tr>
<tr>
<td>R212</td>
<td>10 kΩ</td>
<td>IC202</td>
<td>KA393 (LM393)</td>
</tr>
<tr>
<td>R213</td>
<td>2.5 kΩ</td>
<td>Q1</td>
<td>C945</td>
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</table>

![Figure 8](image1.png)

**Figure 8.** The charger/ac-adapter’s constant current control circuitry’s bias source is the load voltage.

![Figure 9](image2.png)

**Figure 9.** The charger/ac-adapter’s constant current control circuitry’s bias source is separate from and independent of the load voltage.
1.5 Soft start circuitry

Upon initial power up of an SMPS, the PWM (pulse width modulation) control is not effective until the output voltage stabilizes at its normal value. Hence all switching during this initial interval is at the maximum duty cycle rate. The switch turns on again before the transformer magnetizing inductor energy completely resets, and, as a result, the peak inductor current increases as switching cycles pass: indeed, the transformer core may saturate. This is a voltage control mode phenomenon, as the current control is not yet effective. The abrupt increase in inductor current caused by the transformer saturation puts very high stress on the switching device. The so-called soft start technique improves the situation by increasing the duty cycle gradually and therefore preventing the transformer from saturating during the start up phase.

The soft start function is required:
(1) In the voltage control mode the transformer inductor current cannot be limited.
(2) For reducing switching device stress and audible noise. The device stress is due to an abrupt peak current increase, and the audible noise arises from the temporarily saturated transformer that follows the inductor current’s abrupt increase.

Figure 10. Soft start circuitry protects the switching device by allowing the switching duty cycle to increase gradually during the initial transient state.

1.6 Constant power control

In comparatively high power chargers, such as those for notebook and PC batteries, the current and voltage to be controlled are large. In this case, if the control circuit is configured only for a constant voltage /constant current mode, the design must allow for a very large maximum input power which requires a larger transformer and input capacitor and increases the external capacity demand. Figure 11 shows the characteristics of the constant current/constant voltage mode control (a, c) and constant current/constant voltage with constant power control (b, d). A design such as that of Figure 11a, with only constant current/constant voltage control and with limited power (as shown in c), has a lower charging efficiency as compared to a constant power control design. Constant power control is generally incorporated in chargers for comparatively high power notebook PCs.

Figure 11. Constant current/constant voltage control mode (a, c), and constant power control mode (b, d).

1.7 Charger/Ac-adapter using constant power control

Figure 12 (next page) depicts a charger/ac-adapter incorporating Fairchild’s patent pending, constant power control circuit. This is shown within the dashed lines, and operates as follows. Zener diode ZD1 remains off until the output voltage \( V_o \) across the (battery) load reaches ZD1’s breakdown voltage, \( V_{ZD1} \). With ZD1 off, IC201B’s positive input terminal bias is equal to the output voltage; since IC201B is operating as a voltage follower, its negative input terminal bias is equal to its positive input terminal bias (the output voltage), hence there is no voltage difference across R6 and no current flow through it.

As the output load voltage \( V_o \) increases, at some point it will cross and exceed ZD1’s breakdown voltage and current will flow through R6. The voltage difference \( V_6 \) across R6 is defined as \( V_6 = V_o - V_{ZD1} \). When \( V_6 \) is not zero the current flow through R6 is \( (V_o - V_{ZD1})/R6 \). As output voltage, \( V_o \), becomes greater than \( V_{ZD1} \) the current increases linearly and flows through Q2.

The voltage across R10 is held at \( V_{ref} \) which fixes the current through R10. Since the current through R10 is fixed, as the current through Q2 increases the current through Q1 must decrease. The decrease in the current through Q1 decreases the voltage drop across R4, which also decreases the voltage drop across Rsense , and thereby decreases the output current. (see, e.g., Figure 7). This decrease in output current with an increase in output voltage
(when \( V_o \) exceeds \( V_{ZD1} \)) maintains a fixed power level.

1.7.1 Determining component values for the charger/ac-adapter of Figure 12

1) Select \( R_{\text{sense}} = 0.1 \, \Omega \) for minimal power loss.

2) Use \( V_{\text{ref}} \) (LM431) of 2.5V.

3) Assign \( R11 \) and \( R12 \) values assuming an output voltage of 20V.

4) Assign a value to \( R10 \) for a 1mA current flow (1mA selected for calculation convenience).

5) Calculate \( R4 \) assuming a 4A output current: i.e., use 1mA current flow through it and assign a value to it of 400W, for a 0.4V voltage drop.

6) Select \( ZD1 \) as a 10V Zener, since it is desired to initiate constant power control at an output voltage of 10V.

7) Calculate \( R6 \) for a voltage drop across it of 10V, when the output voltage is 20V at a constant current of 2A. Note that setting the current through \( R6 \) to 500\( \mu \)A fixes the current through \( R10 \) at 1mA, which in turn sets the current through \( R4 \) to 500\( \mu \)A. Since the voltage drop across \( R4 \) is controlled to 0.2V, the constant current output is controlled to 2A.

Table 4. Parts list for Figure 12

<table>
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<th>Part</th>
<th>Value</th>
<th>Part</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( R_{\text{sense}} )</td>
<td>0.1 ( \Omega )</td>
<td>( R11 )</td>
<td>17.5 k( \Omega )</td>
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</table>
2. Primary Side Regulation

It is best to include some form of primary regulation in a charger/ac-adapter, to allow the SMPS to handle load (output voltage) variations that would otherwise be outside its capabilities. In Figure 13, the Zener diode at the Fairchild Power Switch (FPS) Vcc terminal regulates the primary side Vcc to a fixed voltage. This plus the transformer’s characteristics achieve the desired regulation of the secondary side output voltage.

Therefore, for good regulation, the transformer should be designed with a high coupling coefficient between the primary and secondary; for this, the sandwich winding method is commonly used.

2.1 Limiting peak power in a low power adapter

Figure 14 shows that a resistor series connected to the Fairchild Power Switch (FPS) feedback terminal appears to be in parallel with the Fairchild Power Switch (FPS)’s internal resistance. This lowers \( V_{fb} \) and therefore lowers the transformer peak inductor current, thereby limiting the input power. The 1mA current source in the figure controls the Fairchild Power Switch (FPS)’s internal feedback voltage \( V_{fb^*} \). The decrease in peak \( V_{fb^*} \) and the limited value of the maximum SenseFET current are indicated through \( R_c \). Note that even though the attached resistor breaks the feedback loop, the feedback voltage does not increase; consequently, the latch circuit does not operate.

Protection such as this is made possible by the Fairchild Power Switch (FPS)’s internal UVLO (under voltage lockout) block. When the output voltage drops, \( V_{cc} \) decreases. Below 10V the Fairchild Power Switch’s internal control IC stops; \( V_{cc} \) goes back and forth between the UVLO limits, and protection begins.

![Figure 13. Adapter using primary regulation.](image-url)
3. Primary Side power terminal design

The primary side power terminal design makes use of the Fairchild Power Switch (FPS), which incorporates within it a high power SenseFET and control IC. When designing, it is helpful to know and understand the Fairchild Power Switch (FPS) internal blocks.

3.1 Feedback terminal design

Design of the Fairchild Power Switch (FPS) feedback terminal is straightforward. First, understand that in normal operation the feedback terminal voltage ranges between approximately 0.2V and 3V. From 3V to 7.5V the device operates at maximum duty cycle. As soon as the feedback voltage reaches 7.5V, internal circuitry shuts down the feedback circuit.

Figure 15 shows the Fairchild Power Switch (FPS) internal feedback block. The sum of the resistances R1 and R2 is 3kΩ, and the current flowing through D2 towards them is 1mA. At 1mA, \( V_{fb} \) is 3V. Therefore, when the voltage on \( C_{fb} \) (from the 5µA current source) exceeds 3V, D1 turns off. From this point on, even if \( V_{fb} \) continues to increase, \( V_{fb}^* \) (the voltage across R1) remains fixed. Consequently the SenseFET current limits the peak current according to the voltage fixed by R1. This peak current is the (max) value listed in the specification of every Fairchild Power Switch (FPS). Therefore generally the peak input power is set by the inductor current regardless of the input voltage.

Therefore, power is limited by Fairchild Power Switch (FPS) internal block alone, without any need for separate, external circuits.

3.2 Under voltage lockout circuit

The Fairchild Power Switch’s current mode PWM control IC (Figure 1) features a UVLO (under voltage lockout) capability, which minimizes power loss in the start-up resistor. Before the operating voltage of the internal control IC reaches its normal level, most of its functional blocks are off. In this initial state it requires only a minimum start up current. The smaller this current, the larger the start-up resistor can be. A larger start-up resistor dissipates less power during normal operation. The start up current charges the \( V_{cc} \) terminal capacitor. When the voltage reaches typically 15V, all the Fairchild Power Switch’s internal blocks start to operate. From this point on, the Fairchild Power Switch (FPS) draws a steady operating current of 10mA and is switching.
When designing with the Fairchild Power Switch, a peculiarity of the UVLO capability should be kept in mind. A charger/adapter using a Fairchild Power Switch (FPS) starts up normally with a small load. With no load, however, it is unable to start when the input power is turned on. The usual cause for this phenomenon is that \( V_{cc} \) drops to the cutoff voltage level before the charger/adapter reaches normal operation. This resets the Fairchild Power Switch (FPS)'s internal control IC. This is easily verified by applying 16V, from a separate power supply, to the Fairchild Power Switch \( V_{cc} \) terminal and then performing the same test. Apply the 16V after input power has been applied to the charger/adapter. If \( V_{cc} \) is applied with no power applied to the Fairchild Power Switch (FPS) shutdown circuit will continue to operate even if input power is applied. If the above test verifies that the start up trouble in the no-load state is due to a drop in \( V_{cc} \) to the cutoff level before the charger/adapter reaches its normal state, then two solutions are available: (1) increase the size of the \( V_{cc} \) capacitor; or, (2) increase slightly the number of turns on the transformer’s \( V_{cc} \) winding.

3.3 Fairchild Power Switch(FPS) protection functions

Fairchild Power Switch(FPS) protection functions fall into three major categories: overcurrent protection; output short circuit protection; and overvoltage protection. These capabilities are built into the Fairchild Power Switch(FPS) and do not require any additional components. Output current limiting (overcurrent protection) is provided through the Fairchild Power Switch(FPS) internal overload protection circuitry. When the flyback converter (Figure 16) is designed for discontinuous current mode (DCM) over all inputs voltages, the output current can be limited regardless of the input voltage. Charger circuits require a more precise constant current control. Therefore, instead of designing the flyback converter based simply on DCM for all input voltages, design it instead using an appropriate division of DCM and CCM (continuous current mode). It should operate in CCM when the input voltage is low, and in DCM when the input voltage is high. Efficiency and thermal characteristics are better in CCM than in DCM.

The Fairchild Power Switch (FPS)'s own internal overvoltage circuit monitors \( V_{cc} \) and shuts down if \( V_{cc} \) reaches a specified upper level. Overvoltage protection, is implemented since a voltage change in the output winding of the flyback converter brings about a voltage change in the \( V_{cc} \) winding. Protection against a short circuit across the output terminals is available through Fairchild Power Switch(FPS) internal circuitry that stops the switching action in the primary side control. The short circuit protection works in two ways, both of which stop the IC’s internal operations: (1) shutdown due to the increase in voltage on the feedback terminal (per the internal latch circuit operation); and, (2) a drop in \( V_{cc} \) to the UVLO cut off level.

3.4 Fairchild Power Switch(FPS) EAR characteristics

Assuming no snubber circuit is used, when a high spike voltage is applied to the MOSFET drain, it is clamped at the MOSFET breakdown voltage. The energy above the breakdown voltage is absorbed by the MOSFET. The amount of energy sufficient to destroy the MOSFET is defined by a specification called EAR (Repeated Avalanche Energy). The EAR specification is very important to an SMPS switching device and is closely connected to its destruction during abnormal operating conditions. Note that the EAR specification for Fairchild Power Switches is much better than that of the competition’s small capacity SMPS products. This is because the SenseFET is constructed with DMOS process.

Several abnormal conditions on the SMPS output may result in large spikes being generated on the primary side of the transformer caused by the transformer leakage inductance. These conditions include output terminal shorted, output...
diode shorted, over voltage protection (OVP) active due to an open feedback terminal, or the input source switch is in a repeated on/off test state. Under such conditions, the switching device’s EAR specification can affect the reliability of the entire SMPS. In normal operation, a product having a weak EAR does not cause a problem because the Vds applied to the switching device does not exceed its breakdown voltage.

However, as just described, in the transient state an applied spike voltage just slightly greater than the switching device’s breakdown voltage could possibly destroy the device. For such a switching device, a snubber circuit, composed of a diode and a high voltage Zener (Figure 17b), must be used to clamp the high spike voltage below the breakdown voltage. All of the energy stored in the leakage inductance is dissipated in the Zener. This kind of snubber unavoidably increases cost and decreases reliability. In contrast, a Fairchild Power Switch (FPS) EAR is superior, and its RCD snubber circuit (Figure 17a, previous page) is reliable and inexpensive.

Nevertheless, it is still possible to boost the effective EAR by using the circuit of Figure 18, which adds R105 and reconnects the Zener ZD102 to the junction of R103 and R102. These changes affect the Fairchild Power Switch (FPS) start time so that the transient state duty cycle is reduced (see Figure 19). It is therefore possible to use a switching element with a lower breakdown voltage.

Fairchild Power Switch (FPS) offers two families rated at 650V and 800V.

Figure 17. Switching devices with weak EARs require relatively costly and unreliable snubber circuits as in (b); Fairchild Power Switches have superior EARs and can use the inexpensive, reliable snubber in (a).

![Figure 17](image1.png)

![Figure 18](image2.png)

![Figure 19](image3.png)

4. Example Transformer Design for Charger

1. Define system specifications:
   - Output power = 4.2V/800mA 3.36W
   - Vac input = 85 - 265Vac (universal input) f = 60Hz
   - Charger efficiency n = 65%

2. Determine minimum dc input voltage (Vmin):
   - When the SMPS operates at the same output power for all ac inputs, the maximum peak drain
current occurs at the minimum input voltage \( V_{\text{min}} \). Also, \( V_{\text{min}} \) will exhibit the largest ripple voltage \( \Delta V \) at that time. The dc link capacitor \( C_{\text{in}} \) is charged and discharged at 120Hz (Figure 20).

Figure 20. If power output stays constant as the ac input varies, peak current drain will occur at \( V_{\text{min}} \). Also, the largest ripple on \( V_{\text{min}} \) occurs at this point; dc link capacitor \( C_{\text{in}} \) charges/discharges at 120Hz.

2a. Calculate energy discharge time, \( T_d \):

\[
T_d = \frac{1}{f_s} \times \frac{1}{4} \left( \frac{\text{arc} \sin \left( \frac{V_{\text{min}}}{V_{\text{min, peak}}} \right)}{\pi} + 1 \right)
\]

2b. Calculate dc link capacitor, \( C_{\text{in}} \):

\[
C_{\text{in}} = \frac{2 \times \text{Win}}{V_{\text{min, peak}} - V_{\text{min}}} = \frac{2 \times 0.035}{(\sqrt{2} \times 85)^2 - (\sqrt{2} \times 85 - 20)^2} = 16 \mu \text{F}
\]

However, 16\( \mu \)F is not a standard value of capacitor. Hence, to calculate the true \( V_{\text{min}} \), select the nearest standard value for \( C_{\text{in}} \) (6.8\( \mu \)F\times2=13.6\( \mu \)F) and substitute it above, solving for \( V_{\text{min}} = 100V \).

3. Set the maximum duty cycle, \( D_{\text{max}} \):

A typical current mode SMPS has a \( D_{\text{max}} \) under 50\%. However, a \( D_{\text{max}} \) over 50\% causes sub harmonic instabilities. Therefore a \( D_{\text{max}} \) of 40\% is recommended.

4. Calculate primary inductance, \( L_p \), and select the Fairchild Power Switch (FPS):

\[
\text{Pin} = \text{In} \times \text{Vin}, \text{In} = \frac{1}{2} \times D_{\text{max}} \times \text{I}_{\text{peak}}
\]

From (2c) above, \( \text{Pin} = 5.169W \) and, from (2d), \( V_{\text{min}} = 100V \). From these values, calculate \( \text{I}_{\text{in}} = 51.69 \text{mA} \) and \( \text{I}_{\text{peak}} = 258.5\text{mA} \) in the discontinuous current mode. The value of \( \text{I}_{\text{peak}} \) indicates that a suitable selection for the Fairchild Power Switch (FPS) is the KA5H0165RN. Knowing \( \text{I}_{\text{peak}} \), calculate \( L_p = 1.55\text{mH} \) from the following formula:

\[
L_p = \frac{D_{\text{max}} \times \text{Vin} \times \text{min}}{I_{\text{peak}} \times f_{\text{switch}}}
\]

To improve EMI and thermal characteristics, \( L_p \) may be modified changing the mode to continuous current mode (CCM) as below.

5. Choose the ferrite core according to the output power: For a charger output power under 5W, an EE1614 core is a suitable choice. According to the TDK core databook, the specifications for the EE1614 (PC40EE16-Z) core are:

\( A_e \) (effective cross sectional area), 19.2\( \text{mm}^2 \)

\( A_l \) 1140nH/\( \text{N}^2 \) 25\%

\( B_s \) (flux density at saturation), 340 - 370mT at 100\( \times \)C

Now recalculate \( A_e \) from the minimum required effective core area, \( A_{\text{min}} \):

\[
A_{\text{min}} = \frac{(I_{\text{peak}} - 1\text{)} \times \sqrt{A_l \times L_p}}{\Delta B_{\text{max}}}
\]

For a design margin, set \( \Delta B_{\text{max}} = 180\text{mT} \) and \( A_{\text{min}} = 15.4\text{mm}^2 \) (<19.2\( \text{mm}^2 \)). To improve temperature and EMI characteristics, increase \( L_p \) to 2.2mH. This, decrease \( I_{\text{peak}} \) to 220mA. Drain current \( I_1 = 38.34\text{mA} \) because the SMPS changes to continuous current mode (CCM) operation during turn on.

5a. Average current will be the same in DCM and CCM if the output has the same value for the same input.

\[
\text{Win} = \frac{C_{\text{in}} \times (V_{\text{min, peak}} - V_{\text{min}})}{\text{I}_{\text{in}}^2}
\]

\[
V_{\text{min, peak}} = 85\sqrt{2}, V_{\text{min}} = 85\sqrt{2} - 20
\]

\[
\text{Win} = \frac{P_{\text{out}}}{\eta} \times \text{Td} = 5.169(7.692) \times 6.773\text{ms} = 0.035\text{j}
\]

\[
C_{\text{in}} = \frac{2 \times \text{Win}}{V_{\text{min, peak}} - V_{\text{min}}} = \frac{2 \times 0.035}{(\sqrt{2} \times 85)^2 - (\sqrt{2} \times 85 - 20)^2} = 16\mu \text{F}
\]
5b. Determine \( L_p = 2.2 \text{mH} \), slope of the drain current can be changed as follows.

\[
\frac{V_{\text{min}}}{L_p(1.55 \text{m})} = \frac{I_{\text{peak}1}}{\text{Ton}}[\Delta C M]
\]

\[
\frac{V_{\text{min}}}{L_p(2.2 \text{m})} = \frac{I_{\text{peak}2} + I_1}{\text{Ton}}[\Delta C M]
\]

\( I_{\text{peak}2} = I_{\text{peak}1} - I_1 \)

5c. calculate \( I_1 \) in CCM.

\( I_{\text{peak}2} = 258.5 \text{mA} - I_1, \frac{100 \text{V}}{2.2 \text{m}} = \frac{(258.5 \text{m} - 2I_1)}{4 \text{ms}}. \)

\[ I_1 = 38.34 \text{mA}, I_{\text{peak}2} = 220 \text{mA}. \]

Therefore, from these substitutions in the formula for \( A_{\text{min}} \), \( AL = 105.8 \text{nH/turn}^2 \).

Therefore, \( I_1 \) becomes 38.34 mA and \( I_{\text{peak}2} \) becomes 220 mA.

6. Determine primary turns, \( N_p = 145 \) turns, from:

\[
N_p = \frac{L_p}{A_L} \text{ or } N_p = \frac{L_p \cdot (I_{\text{peak}2} + I_1)}{A_{\text{min}} \cdot \Delta B_{\text{max}}}
\]

and the core gap as follows

\[
Ag = \frac{N_p^2 \mu_0 A_e}{L_p} \times 10^3 = \frac{145^2 \times 4\pi \times 10^{-7} \times 15.4 \times 10^{-6}}{2.2 \times 10^{-3}} \times 10^3
\]

\[ = 0.185 \text{mm} \]

\( \mu_0 = 4\pi \times 10^{-7} \)

7. Determine secondary turns \( N_s \) as follows:

Consider Figure 21: a portion of the flyback converter (see Figure 16): with the energy of the core \( L_p \) totally discharged, Area \( A \) for charging will be the same as area \( B \) for discharging \( (V_s = V_{\text{out}} + V_d) \).

8. Determine IC bias turns, \( V_{\text{CC}} \):

The selected Fairchild Power Switch (FPS) will operate properly with a \( V_{\text{CC}} \) from 9V to 25V. For this charger, \( V_{\text{CC}} \) can be 24V at maximum output loading of 4.2V, 800mA. Therefore:

\[
\frac{N_{VCC}}{N_{s(10T)}} = \frac{21 \text{V}}{4.7 \text{V}} \quad N_{VCC} = 45 \text{Turns}
\]

9. Determine coil thickness:

Using the copper wire's current capability of 5A/mm², the rms current through \( N_p \) is 88.17 mA, calculated as follows:

\[
I_{\text{rms}} = \sqrt{(I_{\text{peak}}^2 + I_{\text{peak}} \cdot I_{1} + I_{1}^2) \cdot D_{\text{max}}}{3}
\]

The wire thickness \( \phi \) of \( N_p \) is 0.15 mm, from:

\[
\pi \left(\frac{\phi}{2}\right)^2 = I_{\text{rms}} / 5 \text{A}
\]

Use a coil thickness of 0.2 mm. Obtain the thickness of the secondary winding \( N_s \) in the same way.
5. Fairchild Power Switch (FPS) Charger Adaptor Demo Board Circuit Transformer Specification

**Transformer Specifications**

- Ferrite Core: EE1614
- Bobbin: EE1614
- Winding diameter and turn number
  - Primary winding: Ø 0.2mm; 145T
  - Vcc winding: Ø 0.2mm; 40T
  - Secondary1 winding: Ø 0.65mm; 10T
  - Secondary2 winding: Ø 0.2mm; 11T
- Primary self-inductance: 2.2m± 50µH
- Primary maximum leakage inductance: 100± 5µH
- Transformer winding structure
  - Layer 1: 1 → 3, 145T, Ø 0.2mm; Primary winding;
    two edge margins of 2mm in the end of the width
  - Layer 2: 4 → 5, 45T, Ø 0.2mm; Vcc winding;
    two edge margins of 2mm in the end of the width
  - Layer 3: 7 → 9, 10T, Ø 0.65mm; secondary1 winding;
    two edge margins of 2mm in the end of the width
  - Layer 4: 9 → 11, 11T, Ø 0.2mm; secondary2 winding;
    two edge margins of 2mm in the end of the width
**Component List**

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<th>NO</th>
<th>Symbol</th>
<th>Value</th>
<th>Remarks</th>
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**Bobbin**

\[ V_o = 4.2 \text{ V} \]
\[ I_o = 800 \text{ mA} \]
References

