# L6598 BASED 12V/3A RESONANT APPLICATION

**by Eric Danstrom**

*Off-line switching power supplies have two major problems: high switching losses, and an operating environment, which is very sensitive to the radiation of Electromagnetic Interference (EMI) and Radio Frequency Interference (RFI). The switching losses are a largest creator of EMI (conducted) and RFI (radiated), so their control and reduction is a major benefit to the power supply designer. Resonant techniques offer a hope for greatly reducing the switching losses and hence the factors contributing to the EMI and RFI. This paper will present and organized approach to the design of a 36-watt, full-resonant, off-line converter that can be used for portable PC applications.*

**Operation of the Resonant Converters**

Full-resonant converters are based upon the half-bridge and full-bridge PWM topologies. They are driven with a symmetrical waveform whose frequency is changed to control the output voltage. There is a deadtime provided between the conduction of the upper and lower power switches to turn-on with zero voltage witching. Here the drain voltage of the MOSFET turning off immediately swings to the opposite voltage and causes the opposing antiparallel diode to conduct The opposing MOSFET can then turn-on with a very low drain-to-source voltage.

There are two major types of resonant converters: the parallel- resonant converter shown in figure 1a, and the series-resonant converter shown in figure 1b.

**Figure 1. The Two Types of Resonant Converters**



(a) Parallel

(b) Series
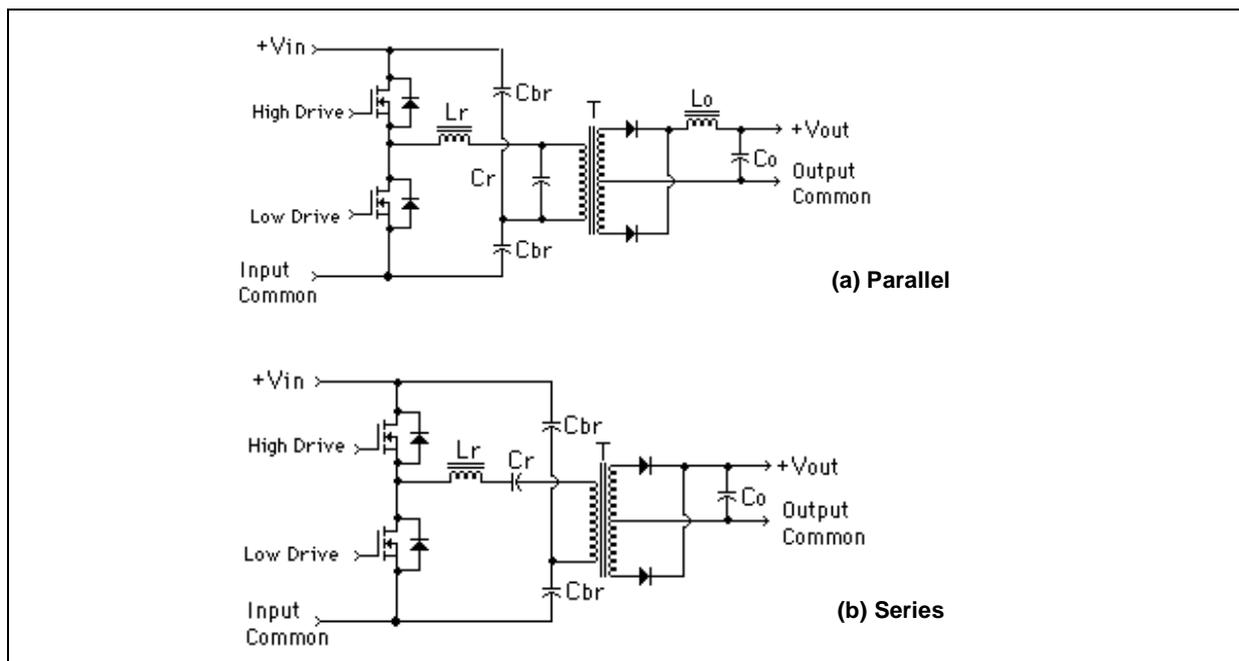
In the parallel-resonant converter, the load is placed in parallel with the resonant tank capacitor. This is done through a forward-mode transformer where the output rectifiers conduct simultaneously with the positive and negative excursions of voltage on the primary winding. The load though, must represent as high an impedance as possible so it does not load the tank capacitor too much which would excessively lower the tank circuit's "Q". This is done by the use of a choke-input (L-C) filter following the output rectifiers. The choke input filter is high impedance when viewed above its filter pole frequency. The output voltage is mathematically the area under the curve of the output voltage waveform. The number of turns in the secondary winding is therefore higher than in a series resonant converter.

In the series resonant converter, which is proposed here, the load is placed in series with the resonant inductance and capacitance. Here the load must be low impedance. This is best done by the use of a capacitor-input filter on the transformer's secondary, where the tank views the low impedance of the output capacitor(s) reflected to the primary winding of the transformer.

Both resonant topologies are driven with a symmetrical waveform above that tank circuit's resonance frequency. The output power is controlled by where on the tank circuit's gain Vs frequency curve the converter is operating. The closer, to the tank's resonance, the higher is the output power.

The series-resonant half-bridge causes the current to resonate in sinusoidal fashion when the tank circuit's voltage is excited at its resonance frequency. The tank circuit is made-up primarily of the resonant inductor ($L_r$) and the Capacitor ($C_r$), but also incorporates the transformer parasitic elements which also include to varying degrees, the primary and secondary leakage inductances and the inter-turn and inter winding capacitances. The load impedance on the secondary of the transformer is also reflected to the primary circuit and becomes a portion of the tank circuit. By incorporating these parasitic elements into the tank circuit, the noise typically generated by their unpredictable behavior within PWM converters is harnessed for real work by the converter.

The half-bridge capacitors (C1, C2) are in series with the tank circuit. Traditionally these are high enough in value such that the center node stays at a fixed voltage of approximately one-half the DC input voltage. These capacitors though could be reduced in value and used as the resonant capacitor itself, thus eliminating the resonant capacitor. Going yet another step further, these capacitors are electrically in parallel within the resonant circuit, and one of them can be eliminated with no detriment to the converter's operation. The upper capacitor would be the logical choice to eliminate from a reliability standpoint.

The external resonant inductor can also be eliminated, if one realizes that the leakage inductance is in series with the primary winding, as is the external resonant capacitor. If one purposely makes a transformer with high primary leakage inductance, one can then eliminate the external resonant inductor in some cases. For the power systems (220-240 VAC), where the transformer has more turns on the primary winding, developing a leakage inductance of around 100 $\mu$H is possible. This is done by using a 2-sectioned bobbin where the primary winding is placed in one-half of the bobbin and the secondary in the other half. This lowering of the coupling between the primary and secondary and the core, raises the leakage inductance. For the low voltage AC power systems (100 - 120 VAC), it can be more difficult to create the large leakage inductance, so an external inductor may still need to be placed within the tank circuit.

The tank circuit's approximate resonance frequency, neglecting the dead time period, can then be then calculated by:

$$f_o = \frac{1}{2\pi\sqrt{L'r \ C'r}} \qquad \text{[Eq. 1]}$$

Where: Lord is the series combination of Lelk + Lr(ext) if used

C'r is the AC value of the bridge capacitor(s)

This frequency will be the primary point of reference for the converter. The control circuit must always stay above this point to maintain the advantage of zero current switching in the semiconductors.

The output load determines the "Q" of the power stage. Its equivalent resistance is reflected through the transformer. This is given by equation 2.

$$Q = \frac{\sqrt{\dfrac{L_r}{C_r}}}{R_{pri} + R_{refl}} \qquad \text{[Eq. 2]}$$

**Figure 2. The Control-Transfer Function of the Series-Resonant Converter**
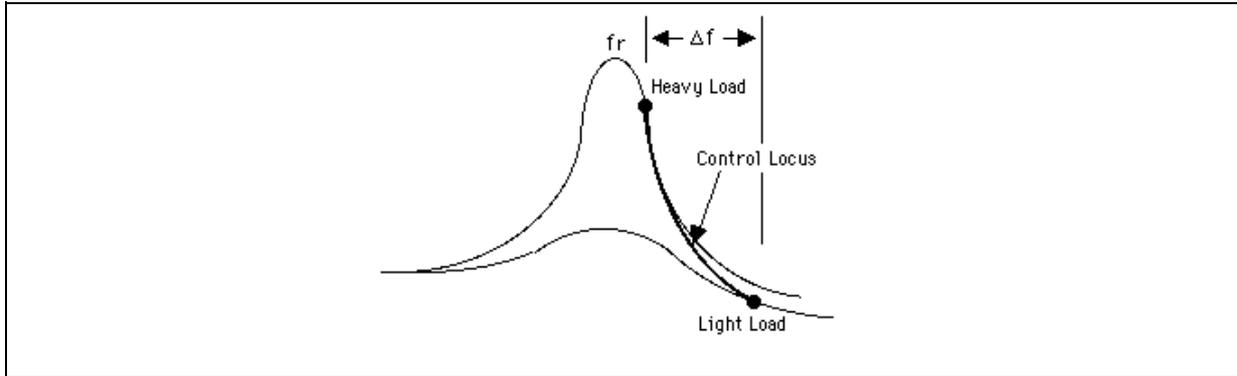


The damping resistance is the sum of the resistances within the primary circuits and the reflected load resistance. The equivalent resistance is reflected through the transformer. This is given by equation 3.

$$R_{refl} = \left(\frac{N1}{N2}\right)^2 R(load) \qquad \text{[Eq. 3]}$$

The typical transfer function of the resonant current within the tank circuit is shown in figure 2. It shows when the converter provides more output power by lowering its frequency. The amplitude of the current grows significantly. Conversely, when the load lightens, the control loop increases the frequency away from the tank circuit's resonance. Simultaneously, the Q of the tank circuit decreases slightly.

Figure 3 shows the converter operating at 110% of rated load and the operating frequency is permitted to go under the resonant frequency of the tank. This condition may occur when the input voltage is at its minimum rating and the output is at rated or slightly over-rated load and the frequency lower limit is not set properly.

Period 1 is the deadtime of the power switches when the center node voltage between the MOSFETs swings from one voltage rail to the opposite rail. After the voltage has transitioned, the primary current flows through the opposing MOSFET's antiparallel diode.

The rate of change in the voltage waveform is dictated by the sum of the parasitic capacitances on the node. These would be primarily the output capacitances of the MOSFETs.

Period 2 begins when the opposing MOSFET turns-on. Current is already flowing through the antiparallel diode when the MOSFET turns ON. The current will continue its decline driven by the tank circuit and then reverse direction. During periods 1 and 2, the secondary voltage on the forward winding transitions to the output voltage and the winding's current begins a sinusoid-like increase. the peak and decline again towards zero where another deadtime is encounderd and the period repeats in the opposing direction.

When the input voltage is low and the output power draw is at or greater than its rated putput, one begins to notice a current valley begin to appear just past the peaks of the primary current. The supply may be operating below resonance. This "inflection" in the primary current waveform increases as the input voltage is further lowered or the output load is increased.

Waveforms at light load can be seen in figure 4. Here, the two normal operating periods (1 & 2) are seen.

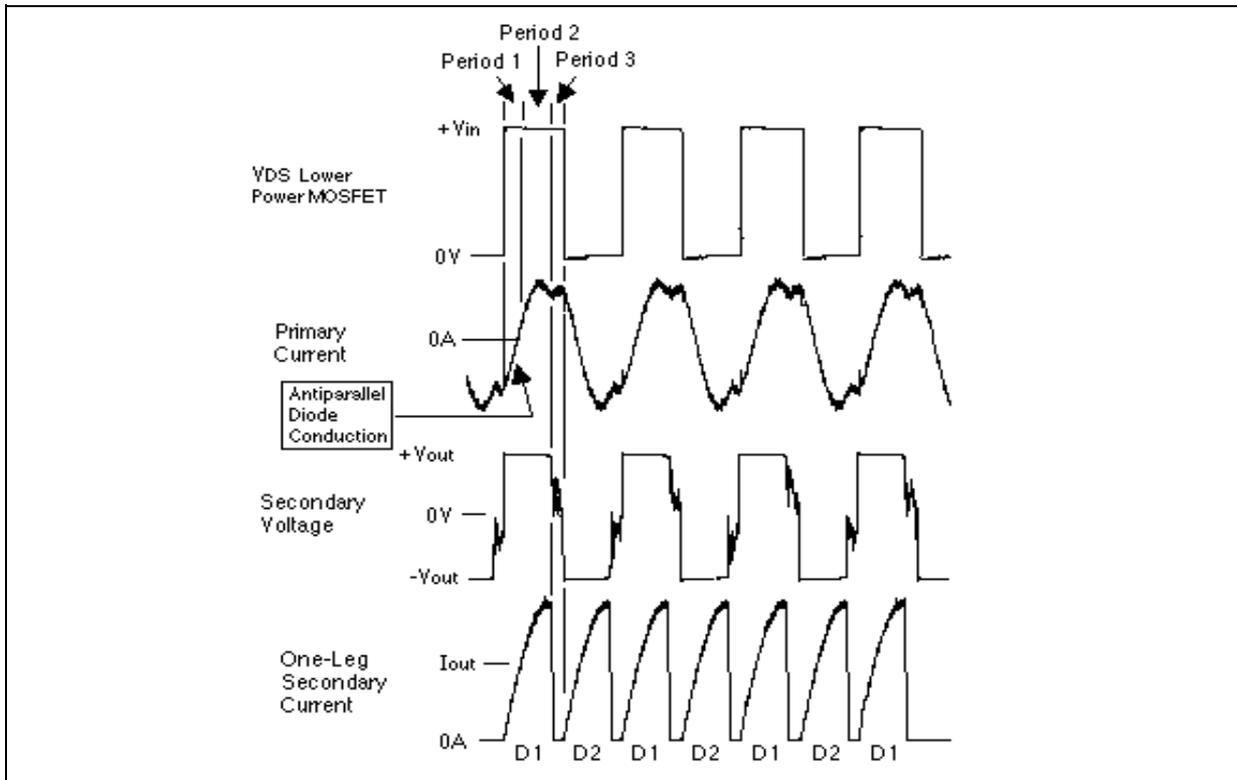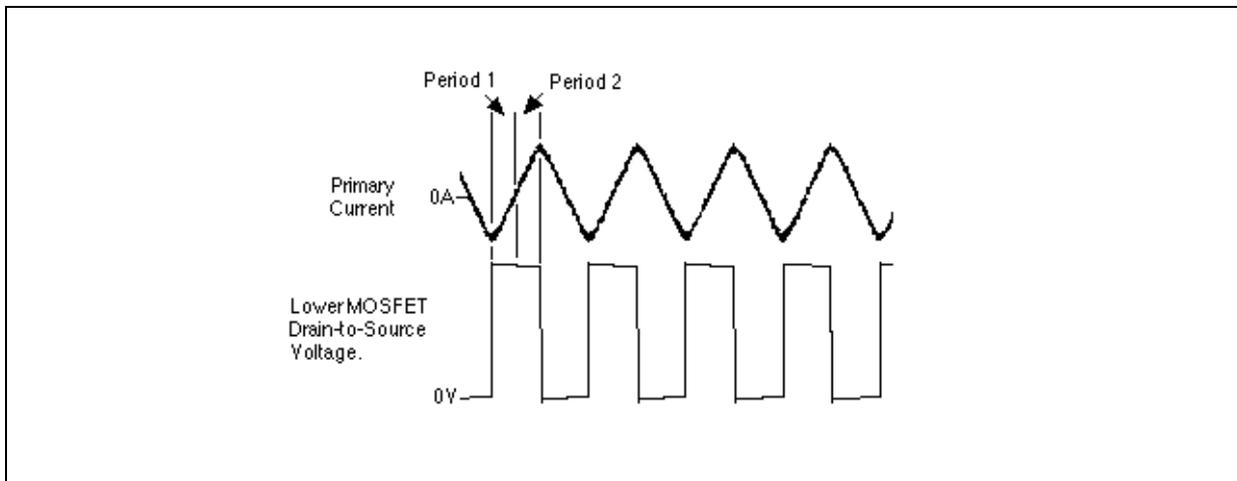**Figure 3. The Waveforms of a Series-Resonant Supply slightly Overloaded.**



**Figure 4. The Primary Waveforms of the Series-Resonant Supply at Light Load.**

**DESIGNING 36 WATT SERIES RESONANT CONVERTER**

**General Specifications:**

Input Voltage:                         85 - 130 Vrms

Output Voltage:                       12 VDC +/- 2%

Rated Output Current:            3 A

Lightest Output Current:         0.5 A

The tank circuit will operate at a resonant frequency of 100 kHz.

**Transformer Design**

The E-E core size for a forward converter of this output power is about 25 mm. I will use TDK part number EE25/19 core using the H7C1 material in this example.

The number of turns needed for the primary, operating at about 40 percent of the saturation flux density (1800 G), is about 49 turns. We will use a 2 section bobbin to "enhance" the primary's leakage inductance, though, is a very complex phenomenon, and this result will only be a rough estimation.

$$L_{leak} \approx \frac{K1(L_{mt})nx^2}{100W_l}\left(T_{ins} + \frac{b_w}{3}\right) \qquad \text{[Eq. 4]}$$

Where:

K1 is equal to 3 for a simple primary and secondary winding

$L_{mt}$ is the mean length of a turn around the bobbin for the whole winding (in).

$n_x$ is the turns contained in the winding being analyzed

W1 is the length of the winding from end-to-end (in)

$T_{ins}$ is the thickness of the wire insulation (in)

$b_w$ is the build (or thickness from the bobbin center-leg) of all the windings of the completed transformer (in)

This relationship shows the largest factors in the amount of leakage inductance a winding will exhibit, is most influenced by the length of the winding ($W_l$), the turns ($n_x$) and the number of layers of turns ($b_w$).

Plugging in the wire and core-related dimensions and solving, one arrives at a preliminary leakage inductance of 44 $\mu$H. This will be used to calculate a preliminary value for the resonant capacitor.

The number of turns needed for the auxiliary and output windings will be 10 turns. I will use litz wire on the secondary winding of 12 strands of 30 AWG to enhance the flexibility of the wire.

**Design of the Tank Circuit**

Calculating the value of the resonant capacitor to produce a resonance frequency of 100 kHz, one uses:

$$C_r = \frac{1}{(2\pi rf)^2 L_r} \qquad \text{[Eq. 5]}$$

Solving for this, using the leakage inductance found above, one gets a resonant capacitance of 0.056 $\mu$F.

**Design of the Output Circuit**

The value of the output capacitance to produce the desired level of ripple voltage is found by:

$$C_o = \frac{I_{out(max)}[1-d]}{f(V_{ripple(p-p)})} \qquad \text{[Eq. 6]}$$

This makes the output capacitance equal to 100 $\mu$F, which is needed to produce a ripple voltage of 30 mV when the power supply is fully loaded at an output of 3 amperes.

**Design of the Input Circuit**

The input filter capacitor's value is determined by the amount of ripple one desires on the DC input voltage of the converter. The ripple voltage of 5-10V is typical. The input capacitor can then be estimated by:

$$C_{(in)} \approx \frac{0.3 P_{(in)}}{f_{in}(V_{in(min)})(V_{ripple(p-p)})} \qquad \text{[Eq. 7]}$$

The approximate value of the input capacitor is then 100-150µF

The 60Hz input rectifiers for the typical capacitor input filter would be determined by the current that is higher than the average input current but lower than the estimated peak current. The input current rating is then approximated by Eq. 8.

$$I_{in(rect)} \approx \frac{P_{in(rated)}}{0.4 V_{in(min)}} \qquad \text{[Eq. 8\}}$$

The input current rating of the input rectifier should then be equal to or greater than 0.8A. The reverse voltage rating would then be 2 times the peak input voltage or 400V. The input rectifiers should then br 1N4004.

**Selection of the Semiconductors**

*The power Switches:*

Since this is a half-bridge 155 VAC application, the power MOS FETs need only have a breakdown voltage ($V_{DSS}$) of between 200 to 250 volts. The choice of N-channel MOSFET is the IRF630S.

*The output Rectifiers:*

The minimum reverse voltage that the output rectifiers will see will be about 25 volts.  The RMS current that each output rectifier will see will be about 1.5 amperes. The diode current waveform will be a half-rectified sine, which will have an approximate peak value of 6 amperes. Since its peak current is much greater than its average current, it is a good idea to select a schottky rectifier that is about the value of the peak value, to keep the voltage drop during the conduction time to a minimum. Therefore a rectifier of 5 or more amperes would wise.  A good choice would be STPS8L30B. This rectifier is an 8 A schottky diode with a reverse voltage of 30 V.

**Design of the Control Circuitry:**

The functions that need setting-up are the excursion limits of the frequency control. These are controlled by the 4 pins; $R_{fmin}$, $R_{fstart}$, $C_{ss}$ and $C_f$. The frequency parameters of these pins must be referenced to the resonant frequency of the tank circuit. Some adjustment will be necessary after the converter is built, to set the control loop to the actual resulting resonance frequency of the physical tank circuit. In a production setting, these control limits should be checked with the production transformer and resonant capacitor.

There is some interaction between the frequency-setting pins. This is necessary for consistent operation of the VCO.

The soft-start capacitor is determined by:

$$C_{SS} = \frac{(0.15 \sec/\mu F)}{T_{SS}} \qquad \text{[Eq. 9]}$$

For a 0.5 second soft-start period the soft-start capacitor is approximately 0.3µF.

One then determines the minimum operating frequency, which should be set at approximately $1.1f_r$. Since there is one equation and two unknowsns. I will select the oscillator capacitor as 360pF. The $R_{fmin}$ resistor is found by:

$$R_{fmin} = \frac{1.41}{f_{min}(C_f)} \qquad \text{[Eq. 10]}$$

Rfmin is then approximately 36kΩ.

From a graph, the value of $R_{fstart}$ to produce a +80kHz start frequency or a start frequency of 190KHz, one should use a 33k$\Omega$ resistor on pin 2.

**Design of the Bootstrap Start-up Circuit**

The bootstrap start-up circuit is intended to allow a small amount of current from the input voltage tothe control IC when power is first applied. It is also intended to restart the converter in the event of an output overcurrent situation. The selected circuit allows a current of about 1mA to charge the IC bypass capacitor should have enough stored energy to start the converter, after which IC can then be powered directly from the auxiliary winding on the power transformer. The input start-up current is not high enough to allow sustained operation of the IC and MOSFET drive. This results in a "hiccup" overcurrent start-up. Another nice operational mode is the input start-up current is turned off after the converter has started thus saving several watts of power during operation.

The selection of the small signal NPN transistor for the 115V AC line should hinge mainly upon thr $V_{CEO}$ of the device. For higher voltage AC lines, the $P_D$ and arc-over voltage of the package become more of an issue. The transistor that meets the needs of the converter is the SO642.

The start-up resistor (collector Resistor)

R(start) is found by:

$$R_{start} = \frac{(120V - 12V)}{1mA} = 108k \text{ or } 100k$$

The base-bias resistor R(bias) must supply both the base current and the minimum amount of current needed to operate the zener diode (225mW) at its "knee" voltage:

$$R_{bias} = \frac{(120V - 12V)}{\left(250\mu A + \frac{I_c}{h_{FE}}\right)} = 379k$$

Make this resistor 360k.

**Compensating the feedback Loop:**

The bandwidth of the control circuit is being set to 15kHz which produces a very good response time for almost all switching power supplies.

The resonance frequency of the tank circuit presents a peculiar set of problems for stability. Approaching the resonance point from below, the gain drastically increses and the gain begins to decrease. Above the resonance frequency, both the gain and the phase are frequency, both the gain and the phase are decreasing rapidly. Since the resonance point of the tank circuit and the ESR of the output capacitor are not fixed, it is advisable to roll-off its gain below to the tank circuit's resonance frequency. One can do this by calculating the gain of the open-loop circuit at DC. The gain is essentially the minimum input voltage divided by the change needed in the control voltage (2V) which produces a +35.5dB gain at DC. By closing the loop, one must have an attenuation of 35.5dB at 15kHz to produce the desired closed-loop gain and phase. There are of course other methods of closing the loop with mixed results. This is the most conservative, thus yelding a guaranteed stable system.

To set the frequency of the crossover frequency of the error amplifier one, essentially "substracts" the DC gain of the control-to-output" characteristic or:

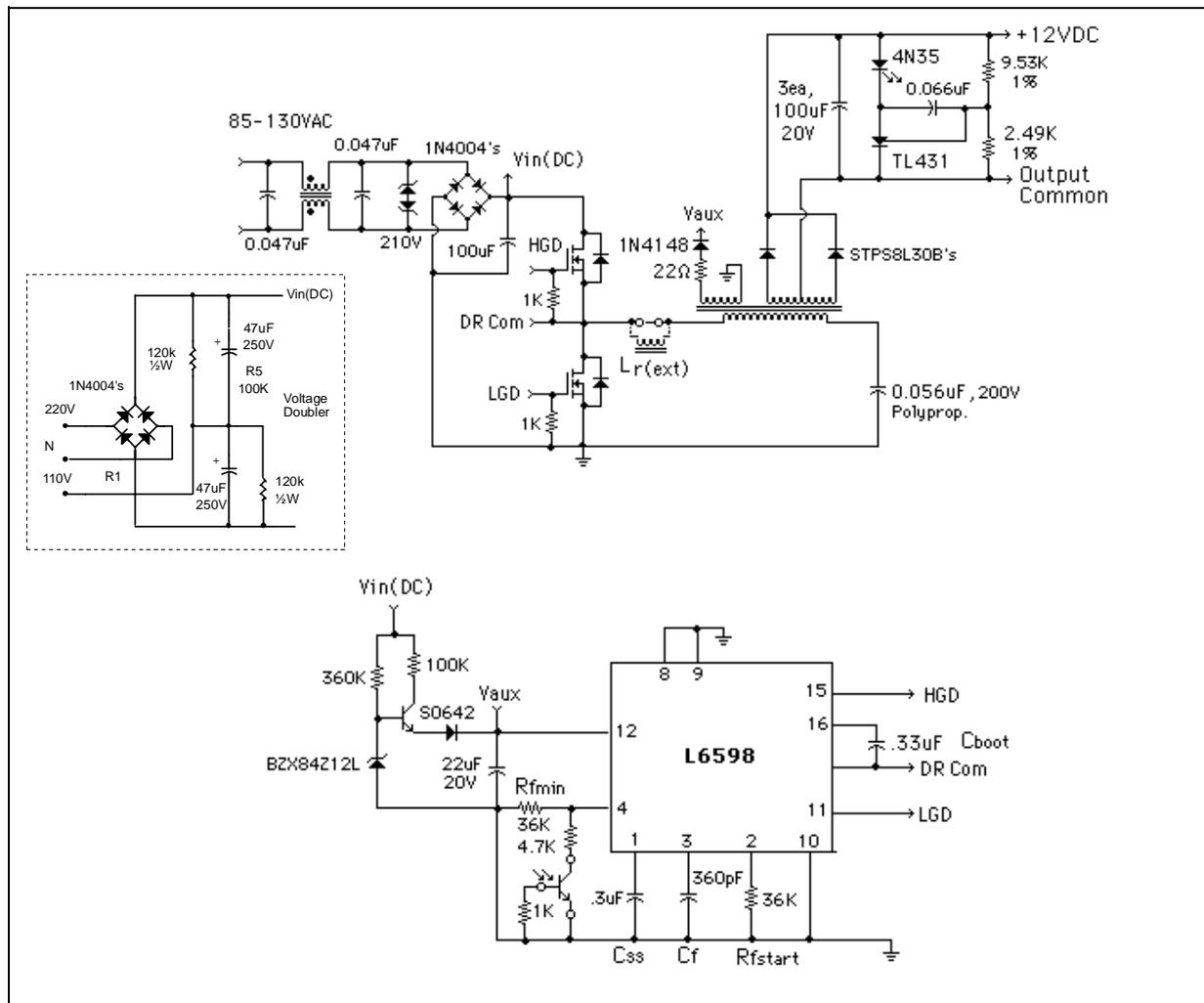$$-35dB = 10Log\left(\frac{f_{ep}}{f_{xo}}\right) \qquad \text{[Eq. 11]}$$

using a 15kHz closed loop bandwidth and solving, results in an error amplifier bandwidth of 252Hz. Then solving the equation;

$$C_f = \frac{1}{2\pi R f_{ep}} \qquad \text{[Eq. 12]}$$

This yields a feedback capacitor of 0.066$\mu$F

The final schematic is shown in figure 5.

**Figure 5. The 36W, Off-line Series-Resonant Converter**



**Summary:**

At the production prototype stage, the resonant frequency of the tank circuit should be checked over a reasonable popolutation of transformers and resonant capacitors. the final transformer in this unit had a leakage inductance of 72mH (as opposed to 45 as calculated). A minor adjustment to the value of the resonant capacitor was required.

The leakage inductance can be held within a fairly tight range, if the transformer manufacturer uses an automated winding process. This should be discussed with the perspective transformer manufactures prior to production.

The final efficiency was found to be between 78% at light loads (0.5A) to 91% at rated load (3A).

This type of converter would be the best choice for consumer appliances where the load presented to the converter is fairly constant and where environment is sensitive to EMI.

**References:**

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