

# A Comparison of Half-Bridge Resonant Converter Topologies

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**Abstract**—The half-bridge series-resonant, parallel-resonant, and combination series-parallel resonant converters are compared for use in low output voltage power supply applications. It is shown that the combination series-parallel converter, which takes on the desirable characteristics of the pure series and the pure parallel converter, removes the main disadvantages of those two converters. Analyses and breadboard results show that the combination series-parallel converter can run over a large input voltage range and a large load range (no load to full load) while maintaining excellent efficiency. A useful analysis technique based on classical ac complex analysis is also introduced.

## INTRODUCTION

TO REDUCE the size of power supplies intended for use in modern computing systems, it is desirable to raise the operating frequency to reduce the size of reactive components. To reduce the higher switching losses resulting from higher frequency operation, resonant power conversion is receiving renewed interest. This paper will compare the series-resonant topology, parallel-resonant topology, and a combination series-parallel resonant topology for use in low output voltage power supply applications. It is shown that the combination series-parallel converter, which takes on the desirable characteristics of the pure series and the pure parallel converter, removes the main disadvantages of those two converters. In particular, it will be shown by analyses and breadboard results that the combination series-parallel converter can run over a large input voltage range and a large load range (no load to full load) while maintaining excellent efficiency. In addition, a useful analysis technique based on classical ac complex analysis is introduced.

## CIRCUIT DESCRIPTIONS

Fig. 1 illustrates three types of resonant converters which may be used for high-frequency switching power supply applications. In the series-loaded circuit, the two capacitors  $C_s/2$  form a series-resonant capacitor of value  $C_s$ . In the parallel-loaded converter  $C_p$  is the only resonant capacitor, while the capacitors  $C_{in}/2$  serve only to split the input dc voltage. The series-parallel loaded converter has both series-resonant and parallel-resonant capacitors. All three of these converters result in low switching losses

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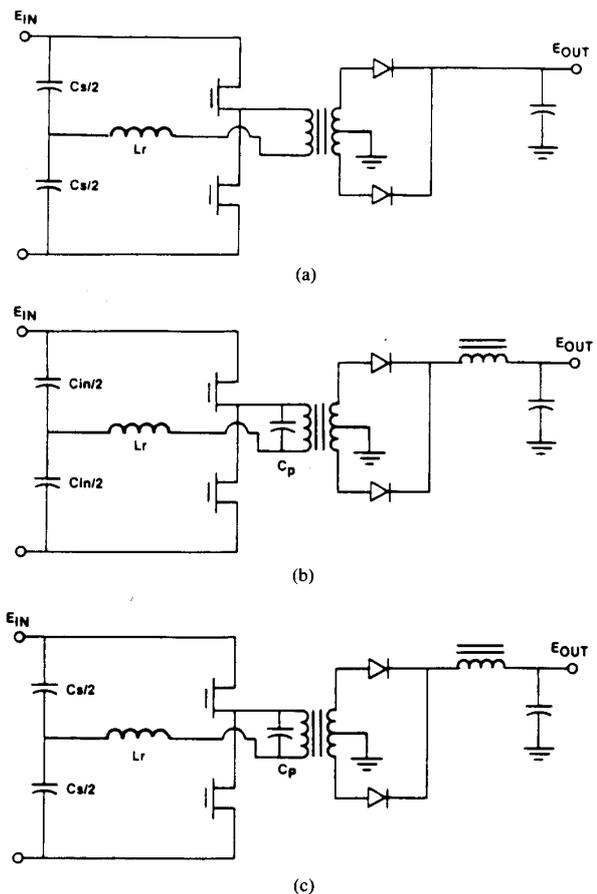


Fig. 1. Three types of half-bridge resonant converters. (a) Series loaded. (b) Parallel loaded. (c) Series-parallel loaded.

for the power devices. The circuits may be operated either above or below the resonant frequency of the resonant circuit. It is presently felt by the author that operation above resonance is preferred. This preference is explained by reference to Fig. 2.

In Fig. 2 waveforms are shown for a resonant converter operating above resonance. In all three converters the half-bridge applies a square wave of voltage to the resonant circuit, and due to the filtering action of the resonant circuit, approximate sine waves of current are present in the resonant inductor  $L_r$ . The fact that the circuit is operating

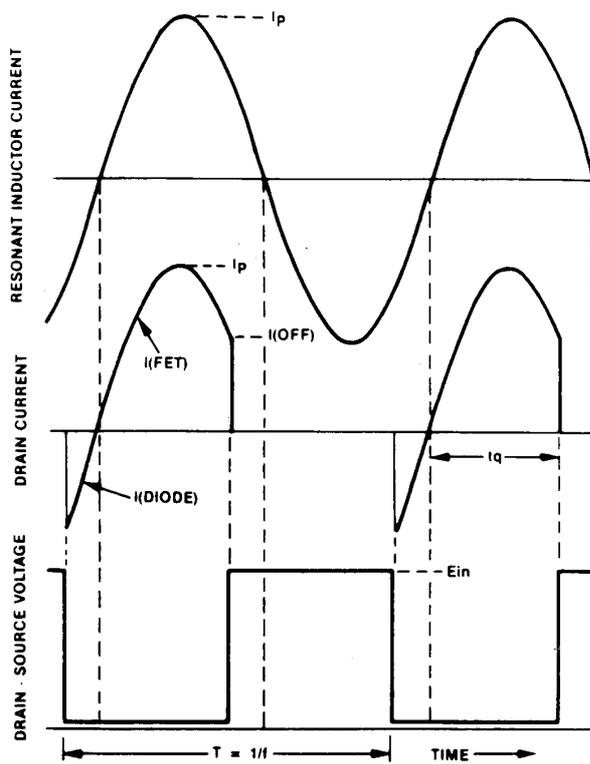


Fig. 2. Ideal resonant converter waveforms.

above resonance can be deduced from the fact that the current delivered to the resonant circuit (that is, the current in  $L_r$ ) is lagging the voltage applied to the resonant circuit (that is, the fundamental component of the square wave applied by the half-bridge circuit). The current carried by the power FETs is a  $180^\circ$  section of this sine wave of current as illustrated in Fig. 2. From Fig. 2 it is seen that no turn-on switching losses exist in the FET because its inverse diode carries current and the voltage across the FET is zero before the FET conducts forward current. Note that the inverse FET current is caused by the opposite FET turning off. For example, if the bottom FET in the half-bridge turns off, the current that was in this FET is transiently maintained by the inductive action of the resonant inductor, which forces the current to come up through the upper FET in the inverse direction (i.e., through its inverse parasitic diode). Note also that once the current reverses due to the resonant action of the circuit, the inverse diode which was conducting has a turn-off time  $t_q$  equal to the forward conduction time of the power FET (see Fig. 2) before forward voltage is applied to the diode. This fact results in no switching stresses being applied to the diode, and, in fact, the inverse parasitic diode associated with power FETs (or bipolar Darlington transistors if used) are of sufficient speed to be useful even at circuit operating frequencies of hundreds of kHz.

Therefore, a main advantage of operating above reso-

nance for the resonant converters is that there are no diode or FET switching losses and the diode can be of medium speed. Note, however, that to achieve those advantages the FETs must switch off current and are therefore subject to turn-off switching losses. However, lossless snubbers can simply be implemented by placing small snubber capacitors directly across the FET devices. No snubber discharge resistors are needed. This can be done because the capacitor is never discharged by turning the FET on but rather is discharged by turning off the opposite FET in the half-bridge. For example, when a bottom FET turns off, a capacitor which is placed directly across the upper FET will be discharged by the load current which will eventually be in the inverse parasitic diode of the upper FET.

Note also that considerable switching losses in power FETs operating at higher frequencies are actually due to storing energy in the FET drain-source and drain-gate capacitances and then discharging these capacitances internally (and losing the associated energy) the next time the FET turns on. This loss can be significant at higher voltages and frequencies. By operating the resonant converters above resonance, this loss is eliminated by the same argument put forth in the previous paragraph concerning lossless snubbers. That is, the energy stored in any capacitance directly across the device is returned to the dc source by virtue of the opposite FET turning off. In addition, the output and input filter sizes are minimized because the frequency is limited to a known lower limit (in operation below resonance the frequency is lowered to control output, and therefore the filters must be designed for the lowest frequencies encountered).

All of the aforementioned advantages are lost if the converter is operated below resonance. That is, below resonance operation results in FET turn-on switching losses, diode switching losses (high-speed diodes are needed), energy stored in device capacitances is discharged and lost internal to the FET's, and the input and output filters must be designed for the minimum switching frequency. FET turn-off does occur in a lossless manner when operating below resonance. However, because turn-off losses can be reduced using the lossless snubber technique when operating above resonance, this is not a major argument for operating the converter below resonance. For all of these reasons, it is felt that operation of resonant converters above resonance is the proper choice for most power supply applications operating at high frequencies. Therefore, the analyses to follow are all done for operation above resonance.

In the following sections an ac analysis technique is described, and the characteristics of each of the three resonant converters when operating above resonance are derived and compared. Further information can be found in [1] and [2], which give a detailed discussion and analysis of the parallel-resonant converter. In [3] a detailed analysis of the series-resonant converter is given. Small-signal stability considerations of resonant converters are given in [4]. The combination series/parallel resonant converter is introduced in [5]. An analysis of the combi-

nation series-parallel converter for the case of operation below resonance is given in [6].

RESONANT CONVERTER CIRCUIT ANALYSIS

For the three resonant converters considered here, the half-bridge converter applies a square wave of voltage to a resonant network. The resonant network has the effect of filtering the higher harmonic voltages so that, essentially, a sine wave of current appears at the input to the resonant circuit (this is true over most of the load range of interest). This fact allows classical ac analysis techniques to be used. The analysis proceeds as follows. The fundamental component of the square wave input voltage is applied to the resonant network, and the resulting sine waves of current and voltage in the resonant circuit are computed using classical ac analysis. For a rectifier with an inductor output filter, the sine wave voltage at the input to the rectifier is rectified, and the average value taken to arrive at the resulting dc output voltage. For a capacitive output filter, a square wave of voltage appears at the input to the rectifier while a sine wave of current is injected into the rectifier. For this case the fundamental component of the square wave voltage is used in the ac analysis.

It is important to note that the power supply load resistance is not the same load resistance which should be used in the ac analysis. The rectifier with its filter acts as an impedance transformer as far as the resonant circuit is concerned. This is due to the nonlinear nature of the rectifier.

Fig. 3 illustrates the derivation of the equivalent resistance to use in loading the resonant circuit when using an ac analysis. The parallel and series-parallel resonant converters use an inductor output filter and drive the rectifier with an equivalent voltage source (i.e., a low-impedance source provided by the resonant capacitor). A square wave of current is drawn by the rectifier, and its fundamental component must be used in arriving at an equivalent ac resistance. For this case, the equivalent ac resistance is given by

$$R_{ac} = \frac{\pi^2}{8} R_L \tag{1}$$

Also given in the figure are the formulas for computing the fundamental ac components from the actual circuit values. The series-resonant converter uses a capacitive output filter and therefore drives the rectifier with a current source. A square wave of voltage appears at the input to the rectifier. For this case the equivalent ac resistance is given by

$$R_{ac} = \frac{8}{\pi^2} R_L \tag{2}$$

Also given in the figure are formulas for computing fundamental ac components from the actual converter waveforms. In summary, classical ac analysis techniques can be used to investigate the characteristics of the three resonant converters by taking the fundamental components

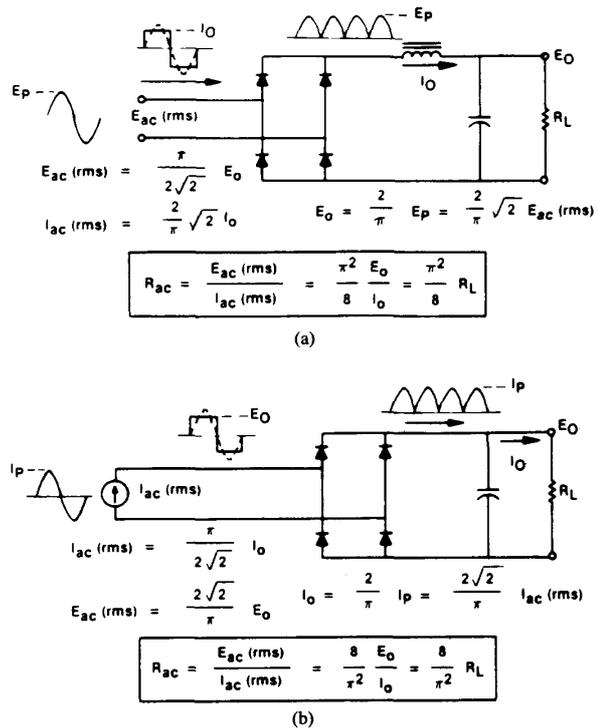


Fig. 3. Equivalent ac resistors as presented by rectifier loads. (a) Voltage source drive (applies to parallel and series-parallel converters). (b) Current source drive (applies to series converter).

of all the waveforms and by loading the resonant circuits with an equivalent resistance which takes into account the nonlinear behavior of the output rectifiers.

Analysis of Series-Resonant Converter

By using the equivalent load resistance \$R\_{ac}\$ and the ac analysis technique derived earlier, the characteristics of the series-resonant converter will be derived. The equivalent ac circuit of Fig. 4 will be used. The voltages designated by upper-case \$V\$'s are the ac fundamental components present in the circuits. They will be converted to square waves where appropriate at the end of the analysis. By using the equation for a voltage divider, it is a simple matter to write down the ac gain of the series-resonant circuit (see top of Fig. 4):

$$\frac{V_O}{V_{IN}} = \frac{1}{1 + j \left[ \frac{X_L}{R_{ac}} - \frac{X_C}{R_{ac}} \right]} \tag{3}$$

Note that \$V\_{IN}\$ is the fundamental component of the square wave of voltage applied to the resonant circuit by the inverter. This square wave of voltage is, for the half bridge converter, of magnitude \$E\_{IN}/2\$ (see Fig. 1). Because the input to the resonant circuit is a square voltage wave and the output is also a square voltage wave, the converter gain in terms of actual converter values is given by the

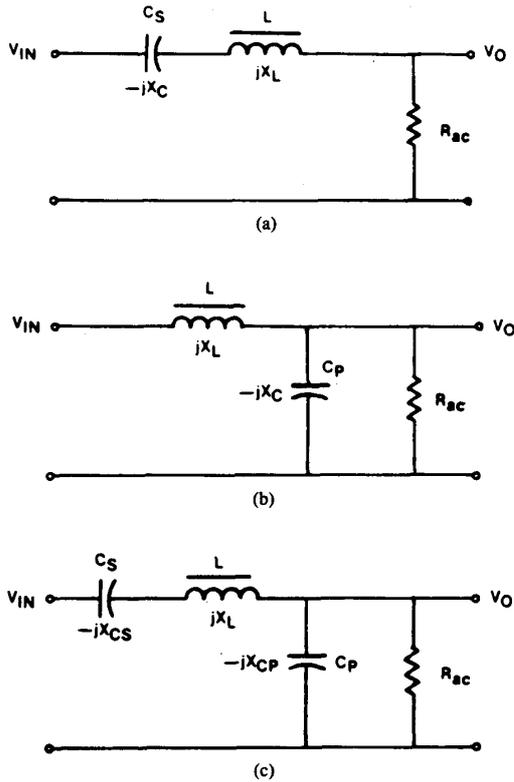


Fig. 4. AC equivalent circuits for resonant converters. (a) Series loaded. (b) Parallel loaded. (c) Series-parallel loaded.

same expression as before (the factor relating the fundamentals to the square wave magnitudes cancel on the left side of the foregoing equation):

$$\frac{E_0}{E_d} = \frac{1}{1 + j \left[ \frac{X_L}{R_{ac}} - \frac{X_C}{R_{ac}} \right]} \quad (4)$$

where

$$E_d = \frac{E_{IN}}{2} \quad (5)$$

Now substituting the fact that

$$R_{ac} = \frac{8}{\pi^2} R_L \quad (6)$$

and defining

$$Q = \frac{\omega_0 L}{R_L} \quad (7)$$

and

$$\omega_0 = \frac{1}{\sqrt{LC_s}} \quad (8)$$

results in the expression for the converter gain finally being given by

$$\frac{E_0}{E_d} = \frac{1}{1 + j \frac{\pi^2}{8} Q \left[ \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right]} \quad (9)$$

Note that the upper-case  $E$ 's refer to the dc voltages which actually occur in the converter.  $E_d$  is the actual square wave voltage applied to the resonant circuit and for the half-bridge is equal to  $E_{IN}/2$ .

The previous equation is plotted in Fig. 5 for five values of  $Q$ . These curves may be considered accurate above resonance where the filtering action of the resonant circuit is sufficient to allow approximate sine waves of current to be in the circuit even though square waves of voltage excite the circuit.

#### Analysis of Parallel-Resonant Converter

A similar analysis can be carried out for the parallel-resonant converter. Using the equivalent resistance  $R_{ac}$  from Fig. 3 and the second equivalent circuit of Fig. 4, the ac gain of the circuit is given by

$$\frac{V_0}{V_{IN}} = \frac{1}{1 - \frac{X_L}{X_C} + j \frac{X_L}{R_{ac}}} \quad (10)$$

for this case (see Fig. 3)

$$V_0 = \frac{\pi}{2\sqrt{2}} E_0 \quad (11)$$

and

$$V_{IN} = \frac{2\sqrt{2}}{\pi} E_d \quad (12)$$

where

$$E_d = \frac{V_{IN}}{2} \quad (13)$$

Defining

$$Q = \frac{R_L}{\omega_0 L} \quad (14)$$

where

$$\omega_0 = \frac{1}{\sqrt{LC_p}} \quad (15)$$

the dc gain of the parallel resonant converter is finally given by

$$\frac{E_0}{E_d} = \frac{1}{\frac{\pi^2}{8} \left[ 1 - \left[ \frac{\omega}{\omega_0} \right]^2 \right] + j \frac{\omega}{\omega_0} \left[ \frac{1}{Q} \right]} \quad (16)$$

This equation is plotted in Fig. 6. Note that the maximum output occurs near resonance for  $Q$ 's above two or so and that the maximum output voltage can be computed from

$$\left( \frac{E_0}{E_{IN}} \right)_{\max} = Q \quad (17)$$

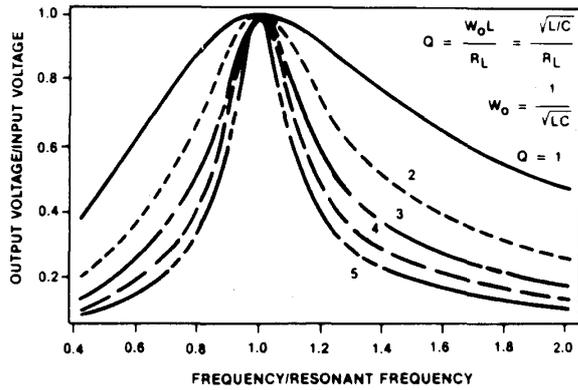


Fig. 5. Series-resonant converter gain (valid for frequencies at or above resonance only).

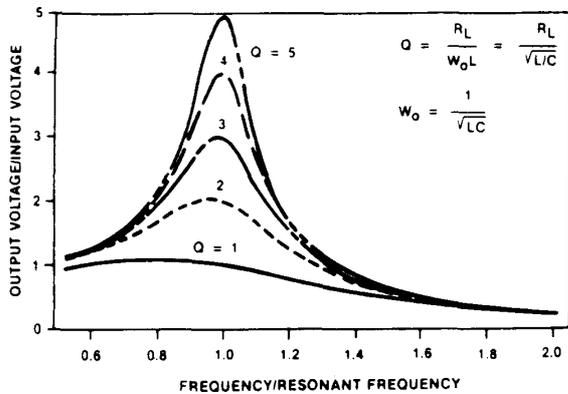


Fig. 6. Parallel-resonant converter gain (valid for frequencies at or above resonance only).

Again, these curves are accurate above resonance where the resonant circuit filters harmonics of the input square waves.

#### Analysis of Series-Parallel Converter

The analysis of the series-parallel resonant converter proceeds in a manner similar to the earlier ac analyses although more algebra is involved. Using classical ac analysis techniques, it can be shown that the gain of the third circuit of Fig. 4 is (using  $R_{ac}$  from Fig. 3)

$$\frac{V_0}{V_{IN}} = \frac{1}{1 + \frac{X_{CS}}{X_{CP}} - \frac{X_L}{X_{CP}} + j \left[ \frac{X_L}{R_{ac}} - \frac{X_{CS}}{R_{ac}} \right]} \quad (18)$$

Defining

$$Q_s = \frac{X_L}{R_L} \quad (19)$$

where

$$\omega_s = \frac{1}{\sqrt{LC_s}} \quad (20)$$

and using the facts (see Fig. 3)

$$E_0 = \frac{2\sqrt{2}}{\pi} V_0 \quad (21)$$

and

$$V_{IN} = \frac{2\sqrt{2}}{\pi} E_d \quad (22)$$

the gain of the series-parallel converter may be written as

$$\frac{E_0}{E_d} = \frac{1}{\frac{\pi^2}{8} \left[ 1 + \frac{C_p}{C_s} - \omega^2 LC_p \right] + jQ_s \left[ \frac{\omega}{\omega_s} - \frac{\omega_s}{\omega} \right]} \quad (23)$$

As seen from the previous expression, the gain will depend on the choice of the ratio of  $C_p$  to  $C_s$ , which also determines the parallel- or series-resonant characteristics of the circuit. The choice of this ratio will be discussed later. Here, the characteristic gain curves for two cases will be plotted. For  $C_p = C_s$ , the gain is given by

$$\frac{E_0}{E_d} = \frac{1}{\frac{\pi^2}{8} \left[ 2 - \left[ \frac{\omega}{\omega_s} \right]^2 \right] + jQ_s \left[ \frac{\omega}{\omega_s} - \frac{\omega_s}{\omega} \right]} \quad (24)$$

If  $C_s = 2C_p$ , the gain is given by

$$\frac{E_0}{E_d} = \frac{1}{\frac{\pi^2}{16} \left[ 3 - \left[ \frac{\omega}{\omega_s} \right]^2 \right] + jQ_s \left[ \frac{\omega}{\omega_s} - \frac{\omega_s}{\omega} \right]} \quad (25)$$

Figs. 7 and 8 show graphs of the previous two equations. Each curve is for a different value of  $Q_s$ , the series  $Q$  of the circuit. As seen, for series  $Q$ 's above three or four, the peak of the resonant curves appear at approximately the same frequency given by the resonant frequency of the series capacitor and the series inductance. In other words, for these values of series  $Q$ 's, the load resistance is low enough to approximately "short out" the parallel resonant capacitor, which results in the circuit approximating a series-resonant converter. As the converter is unloaded, the series  $Q$  expression decreases and the resonant peak moves higher in frequency. This is due to the fact that the equivalent resonant capacitance at light load is given by the parallel combination of the series- and parallel-resonant capacitors (see Fig. 4 for the case when  $R_{ac}$  is large). Finally, at light loads and no load, the resonant peak will occur at a frequency given by

$$f_{02} = \frac{1}{2\pi \sqrt{\frac{LC_p C_s}{C_p + C_s}}} \quad (26)$$

Note that as the series  $Q$  expression results in a smaller value, the parallel  $Q$  expression (which is equal to the reciprocal of the series  $Q$  expression) results in a larger value. That is, as the load resistance goes from a small value to a large value, the circuit characteristics go from

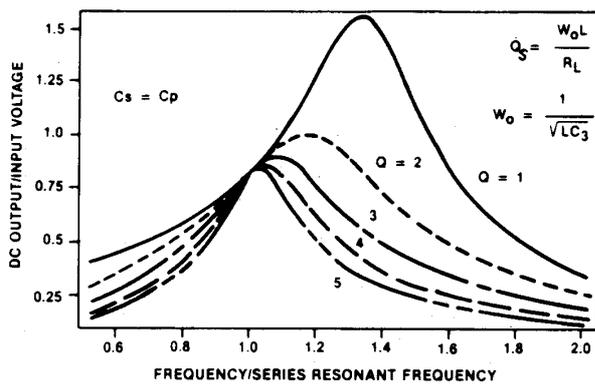


Fig. 7. Combination series-parallel converter gain (valid for frequencies at or above resonant peaks only).

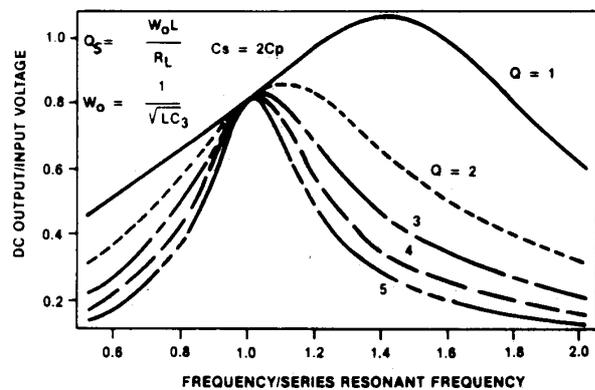


Fig. 8. Combination series-parallel converter gain (valid for frequencies at or above resonant peaks only).

those of a series-resonant to those of a parallel-resonant converter.

COMPARISON OF RESONANT CONVERTER TOPOLOGIES

In the following discussion the characteristics peculiar to each of the three resonant converter topologies analyzed will be discussed with a view toward selecting the proper converter for a given application.

Series-Resonant Converter

The series-resonant converter (top of Fig. 1) has the main disadvantage that the output voltage cannot be regulated for the no-load case. This can be seen from the characteristic resonant curves of Fig. 5. At  $Q = 1$ , for example, the curves have very little "selectivity," and, in fact, at no load the curve would simply be a horizontal line. This means that this converter would only be used "as is" in applications where no-load regulation was not required. A means of getting some no-load regulation would be to turn the converter on and off in a time ratio control fashion at a frequency much lower than the resonant frequency of the converter.

Another disadvantage of this converter is that the output dc filter capacitor must carry high ripple current (equal

in magnitude to 48 percent of the dc output current). This is a significant disadvantage for applications with low output voltage and high current. For this reason the series-resonant converter is not considered suitable for low-output-voltage high-output-current converters but rather is more suitable for high-output-voltage low-output-current converters. For the high-output-voltage case no magnetic components are needed on the high-voltage side of the converter.

The main advantage of the converter is that the series-resonant capacitors on the primary side act as a dc blocking capacitor. Because of this fact the converter can easily be used in full-bridge arrangements without any additional control to control unbalance in the power FET switching times or forward voltage drops (i.e., dc currents are kept out of the transformer). For this reason the series-resonant converter is suitable for high-power applications where a full-bridge converter is desirable.

Another advantage of the series-resonant converter is that the currents in the power devices decrease as the load decreases. This advantage allows the power device conduction losses (as well as other circuit losses) to decrease as the load decreases, thus maintaining high part load efficiency. As will be seen in the next section, this is not the case for the parallel-resonant converter.

Note that if the converter is operating near resonance (i.e., at heavy load) and a short circuit is applied to the converter output, the current will rise to high values. To control the output current under such conditions, the frequency of the converter is raised by the control. Making the converter short circuit proof is relatively easy because it takes a few resonant cycles for the current to rise. This fact allows considerable time for the control circuit to take action.

Parallel-Resonant Converter

The characteristic gain curves for the parallel-resonant converter are given in Fig. 6. From these curves it is seen that, in contrast to the series-resonant converter, the converter is able to control the output voltage at no load by running at a frequency above resonance. Note also that the output voltage at resonance is a function of load and can rise to very high values at no load if the operating frequency is not raised by the regulator.

The main disadvantage of the parallel-resonant converter is that the current carried by the power FETs and resonant components is relatively independent of load. By way of illustration, Fig. 9 shows calculated values of input current to the resonant circuit (i.e., the current in the resonant inductor—which is also in the power FETs) as a function of load resistance. Also shown in the figure is the phase of the current relative to the fundamental of the square wave of voltage applied to the resonant circuit and the frequency of operation. As seen, as the load resistance increases (load decreases), the frequency of operation increases to regulate the output voltage, but the current into the resonant circuit stays relatively constant. The consequence of this behavior is that the conduction losses in

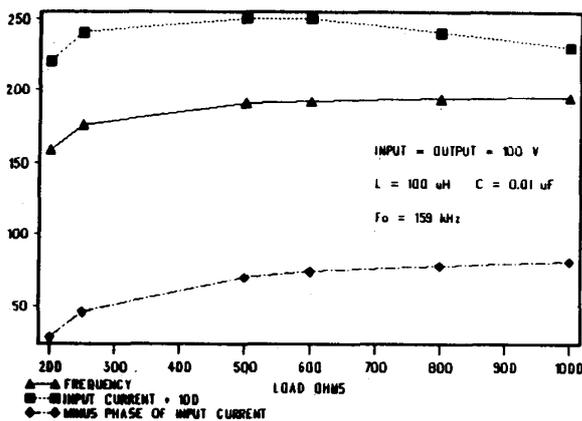


Fig. 9. Characteristics of parallel-resonant converter.

the FETs and the reactive components stay relatively fixed as the load decreases so that the light-load efficiency of the converter suffers. In addition, this circulating current increases as the input dc voltage to the converter increases. Thus this converter is less than ideal for applications which have a large input voltage range and which require it to operate considerably below its maximum design power while maintaining very high efficiency. Conversely, the converter is better suited to applications which run from a relatively narrow input voltage range (e.g., plus or minus 15 percent) and which present a more or less constant load to the converter near the maximum design power (e.g., 75 percent of maximum design power). Of course, the power converter must be designed thermally for the maximum power and, therefore, has no problem running at reduced power thermally—only the part-load efficiency is less than the full-load efficiency.

The parallel-resonant converter is suitable for low-output-voltage high-output-current applications. This is due to the fact that the dc filter on the low-voltage-output side of the transformer is of the inductor input type and, therefore, dc output capacitors capable of carrying very high ripple currents are not needed. The inductor limits the ripple current carried by the output capacitor. Note also that the transformer leakage inductance could be used as the resonant inductance by placing the resonant capacitor across the total span of the secondary winding. This is normally not ideal for low output voltages because the capacitor would have to carry too much ac current. However, for higher output voltage converters this placement of the resonant capacitor may be desirable. Also, the resonant capacitor can be placed on a tertiary transformer winding. These alternate arrangements are discussed more fully in [1].

The parallel-resonant converter is naturally short circuit proof. This property can be seen by applying a short directly across the resonant capacitor. For that case, the entire square wave voltage applied by the inverter is directly across the resonant inductor and, therefore, the current is limited by this impedance. This property makes the par-

allel-resonant converter extremely desirable for applications with severe short circuit requirements.

#### Combination Series-Parallel Converter

The combination series-parallel converter attempts to take advantage of the best characteristics of the series and the parallel converter while eliminating their weak points (lack of no-load regulation for the series-resonant converter and circulating current independent of load for the parallel-resonant converter). As will be shown, this goal is met by proper selection of the resonant components but a somewhat wider frequency range of operation is needed.

By viewing the characteristic gain curves of Figs. 7 and 8, it is clear that the converter can operate and regulate at no load provided that the parallel-resonant capacitor  $C_p$  is not too small (if  $C_p$  is zero, then the circuit reverts to the series-resonant converter). It is seen that the smaller  $C_p$  is, the less "selectivity" is available in the resonant curves. That is, the converter resembles a series converter more and more as  $C_p$  gets smaller and smaller. However, for reasonable values of  $C_p$  the converter will clearly operate with no load, which removes the main disadvantage of the series-resonant converter. In doing this, the converter takes on some of the characteristics of the parallel-resonant converter.

It is desirable that the main disadvantage of the parallel-resonant converter (constant circulating current independent of load) not be present in this converter. This will be true but only for certain component values. This can be seen in the calculated curves of Fig. 10 which give the input current to the resonant circuit (i.e., the current in the resonant inductance) as a function of load resistance. As seen for these sample component values, the input current decreases as the load decreases (resistance increases) as desired. This curve may be compared with that of Fig. 9 for the straight parallel-resonant converter. It was noted that for other values of the circuit components no current decrease was achieved.

The effect of decreasing  $C_p$  was shown in Figs. 7 and 8, respectively. As  $C_p$  gets smaller relative to  $C_s$ , the curves have less "selectivity." For example, if it is desirable to maintain the output voltage at a normalized value of 0.6 at a light load of  $Q = 1$ , Fig. 7 shows that the frequency of operation needed is approximately 1.7. On the other hand, Fig. 8 shows that a frequency of 2 is needed. In other words, as  $C_p$  gets smaller, the upper frequency needed at light loads increases. This is the limiting factor in reducing  $C_p$  to reduce circulating current.

To have the circulating current decrease with load to maintain high part-load efficiency, it is desirable to select the converter components so that the full load  $Q$  is in the neighborhood of 4 or 5. For these values of  $Q_s$  the converter appears essentially as a series-resonant converter and the circulating current will decrease as the load decreases. As the load decreases further, the converter takes on the characteristics of a parallel-resonant converter, and the circulating current no longer decreases with load.

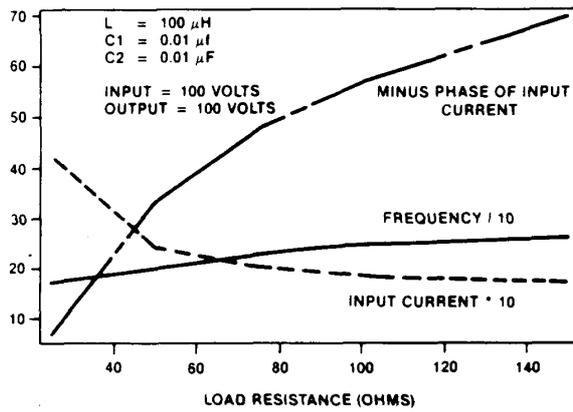


Fig. 10. Characteristics of combination series-parallel converter.

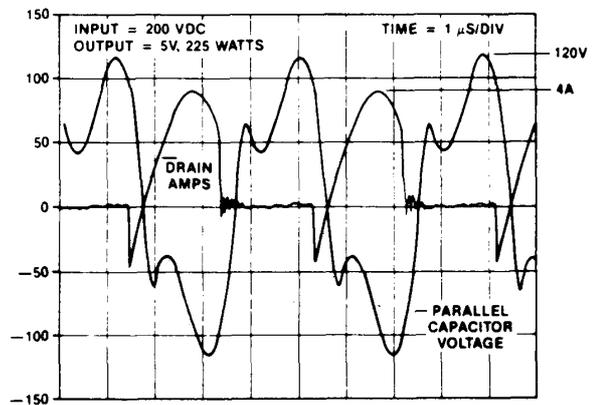
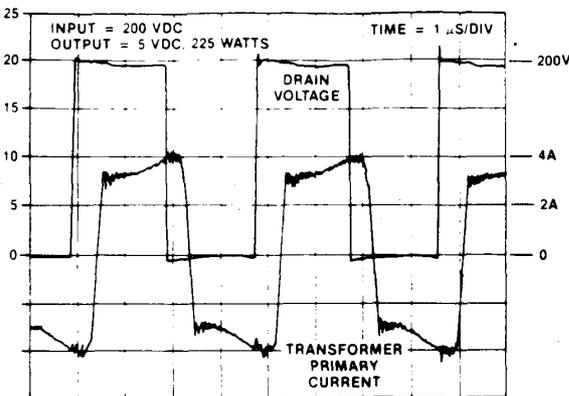
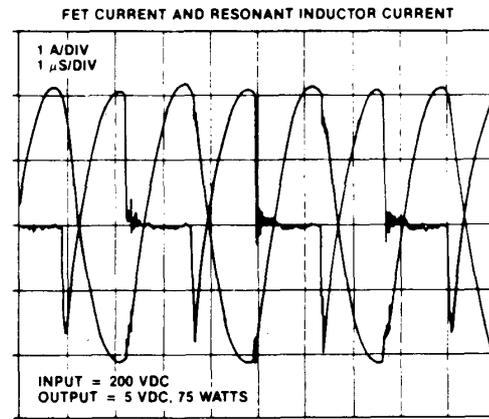
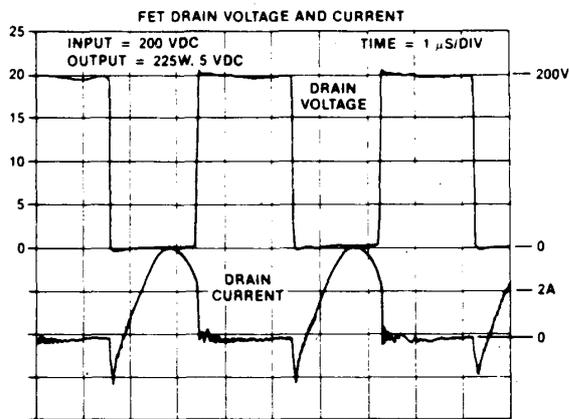


Fig. 11. Breadboard waveforms.

However, for the case of  $C_p = C_s$  the circulating current decreases approximately 2 to 1 as the load decreases from its full-load value. Because the losses due to the circulating current are proportional to the square of the current, the conduction losses are decreased 4 to 1 over their full-

load value. This decrease is sufficient to maintain good part-load efficiency. Therefore, it is felt that  $C_s = C_p$  is a good compromise design which gives good part-load efficiency while allowing regulated operation at no load with a reasonable upper frequency.

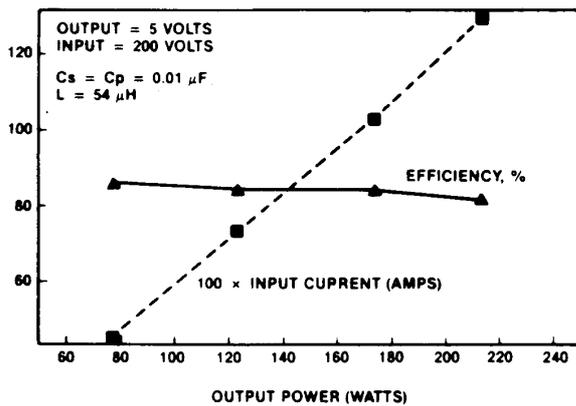


Fig. 12. Performance of combination series-parallel converter bread-board.

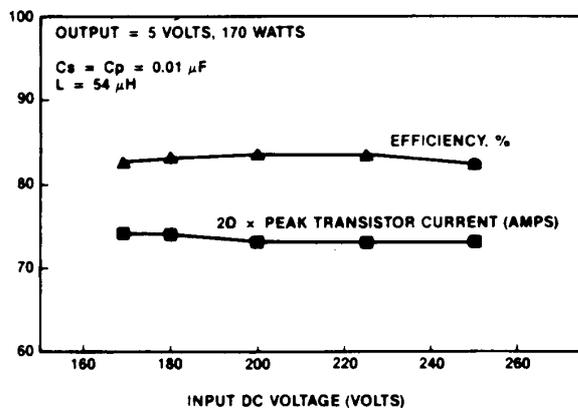


Fig. 13. Performance of combination series-parallel converter bread-board.

#### EXPERIMENTAL RESULTS

A combination series-parallel converter (Fig. 1) was breadboarded to verify the theoretical predictions. A  $48\text{-}\mu H$  resonant inductor was used. The parallel capacitor and equivalent series capacitor were set equal to each other ( $0.01 \mu F$ ). The circuit was designed to deliver 250 W to a 5-V load. The input voltage was 180–360 V dc. The curves of Fig. 7 were used in the design. Fig. 11 gives measured waveforms. These may be compared with the ideal waveforms of Fig. 2. As seen, lossless turn-on is demonstrated, and the current waveforms are sinusoidal in nature. Fig. 12 gives the measured efficiency (not counting the control power) and the peak current to the resonant circuit (i.e., the peak FET current or circulating current). As seen, the circulating current decreases with load, which accounts for the excellent hold-up of efficiency at light load.

Fig. 13 gives the efficiency and peak circulating current (peak transistor current) as a function of input dc voltage

to the converter at approximately 70-percent load. As seen, these quantities remain approximately constant as the input voltage varies. The frequency varied between approximately 250 and 380 kHz for a load variation of 75–225-W output and a constant 200-V dc input.

#### SUMMARY AND CONCLUSION

An ac analysis method was used to derive the characteristics of the half-bridge series-resonant, parallel-resonant, and series-parallel resonant dc-dc converters for the case of super-resonant operation. Using these analytical results, as well as experimental results, it was shown that the combination series-parallel converter takes on the desirable characteristics of the pure series and the pure parallel converter while removing the main disadvantages of those two converters. In particular, it was shown that the combination series-parallel converter can run over a large input voltage range and a large load range (no load to full load) while maintaining excellent efficiency.

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