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PCI'82 - SEPTEMBER

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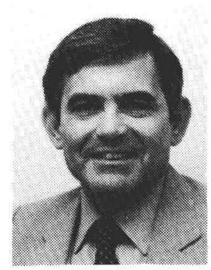
AN ANALYSIS OF RESONANT SWITCH MODE POWER SUPPLIES

Jozef H. Hendriks
TERADYNE, Inc.
183 Essex St.
Boston, MA 02111 USA



Jozef H. Hendriks

ABSTRACT

Resonant-type power converters generate wave shapes which are basically sinusoidal and exhibit all the advantages associated with sine waves in practical applications. An analysis of the single ended version of these resonant types of power converters reveals characteristics that could make such converters quite popular for domestic and industrial use.

AN ANALYSIS OF RESONANT SWITCH MODE POWER SUPPLIES

Jozef H. Hendriks
TERADYNE, INC.
183 Essex St.
Boston, MA 02111 USA

Resonant type switch mode power supplies are not new. My first encounter with such systems was in 1952, where the application was a 25kV supply with low enough energy storage to permit safe beam current limiting in a 6cm diameter TV projection tube. The switch was a single, class C driven, audio penthode running at 450 kHz. (See fig. 1)

A similar circuit was used by the founding engineers of Teradyne in 1961 for generating high test voltages for silicon rectifiers. (See fig. 2) The power switch acted as a controllable current source in the on state and the effective Q of the tank circuit modulated the operating frequency and the conduction angles of the rectifying diodes in the voltage doubling circuit. With an excellent reliability record; a push-pull version was built 10 years later(10W at 2kV). Full wave voltage doubling was used to achieve slew rates of 200V/ms and reduced output ripple.

In the early seventies, resonant mode switching found applications in P.S. systems for aerospace, telecommunications, and inductive heating. Self commutating of power SCR's was the primary reason and all the other advantages have only recently been described. (See ref. 1-3)

These are:
1) lower performance switches
2) improved reliability
3) less EMI conduction and radiation

What do we mean by resonant mode?

A method of power conversion where the rate of energy exchange is determined by at least one LC energy reservoir, the effective damping of which results in a Q >1.

For such circuits to be stable, the energy injection has to be properly timed, excess energy must be returned to the source and the stored dynamic energy should be small compared to the static energy stored at the load port, in order to achieve a good transient response.

Small displacement, high rpm engines in European cars makes them very attractive in dense traffic. This is a good analogy to clarify the previous statement.

450kHz Resonant Switch Mode PS (1952)

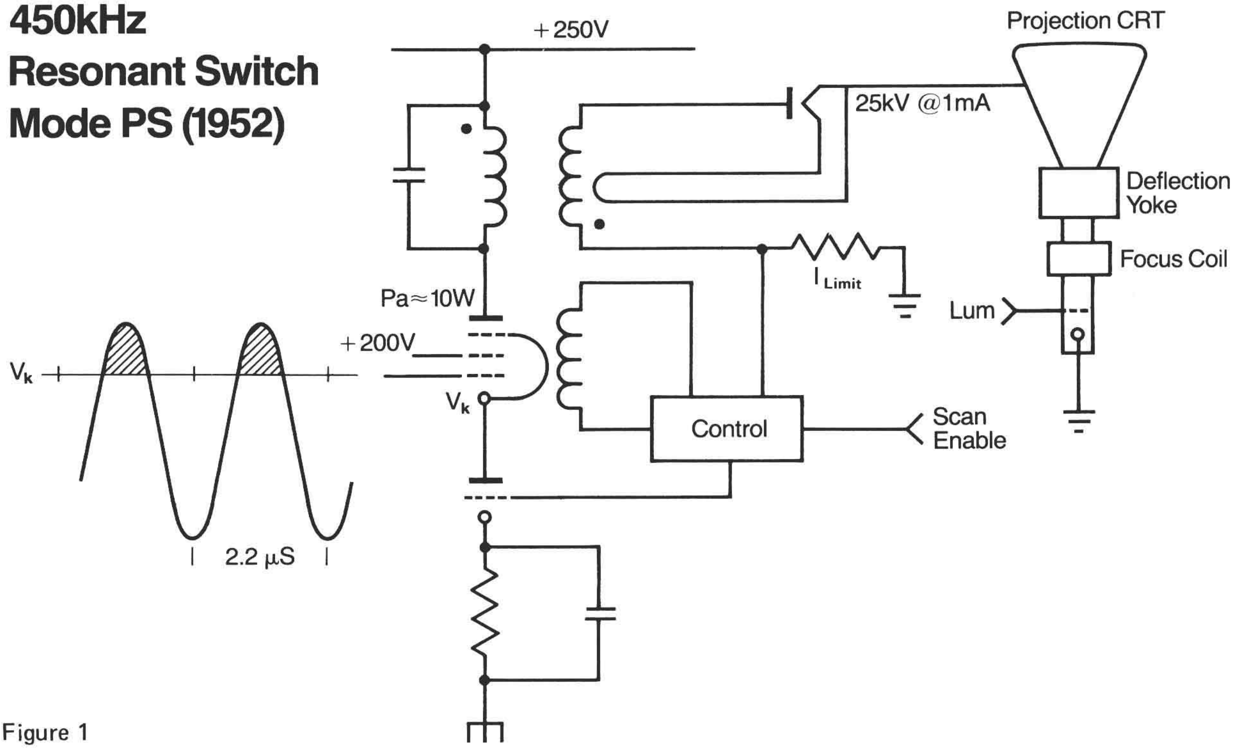


Figure 1

25 – 40kHz Resonant Switch Mode PS (1961)

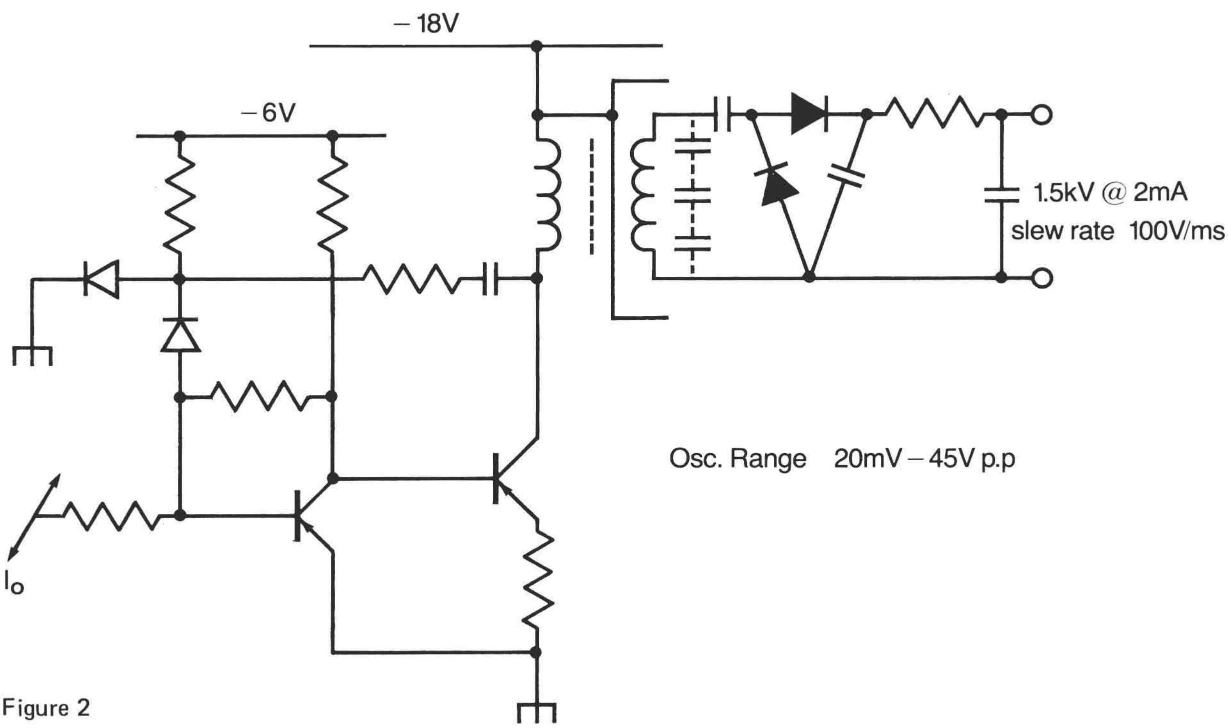


Figure 2

A single active switch makes it possible to transfer energy in both the flyback and the forward mode to lower either the output impedance, or to permit a wide range of input source voltages. A corresponding improvement of power factor will result when fed by full wave rectified AC with a minimum of smoothing.

Non-ideal storage type transformers can be fully utilized to virtually eliminate rectifier switching losses, and, by double tuning, will reduce considerably, the ESR losses in capacitors, and core losses in transformers. A 200kHz converter with power FETs is a recent example of brute force engineering rarely meeting its goal. (See ref. 4)

A simple circuit which takes advantage of non-ideal components has been described by Philips N.V. at PCI'80 in Munich. (See ref. 5 and fig. 3)

The objective of this paper is to derive closed form solutions for all design variables and relate the operation of the converter to well known existing circuits.

The inductor transformer is built with leakage inductances (k<1) in order to meet the following objectives:

1) to permit double tuning and to eliminate all switch snubber networks
2) to control the short circuit current
3) to make shielding between primary and secondary windings easier

As the output storage capacitor acts as a voltage source during its charging interval, the widest conduction angle can be achieved if the primary capacitor resonates with the leakage inductance at the third harmonic of the fundamental frequency of L_o and (C_1+C_2).

$$\omega_s^2 = \frac{1}{L_s C_1} = \frac{3^2}{L_o(C_1+C_2)} = 9\omega_o^2 \qquad 1)$$

L_o= open circuit inductance

L_s= short circuit inductance

ω = radial frequency

The equation for ω_o is not exact, but accurate enough for k >0.8 and its final expected tolerance. The relative error is $\frac{1}{2}\sqrt{1-k^2}$.

Using the T equivalent circuit for a 1:1 transformer and making $C_1=C_2=C$, we can derive:

$$\left(\frac{\omega_s}{\omega_o}\right)^2 = \frac{8}{3-2k^2-k^4} \qquad 2)$$

(continued...)

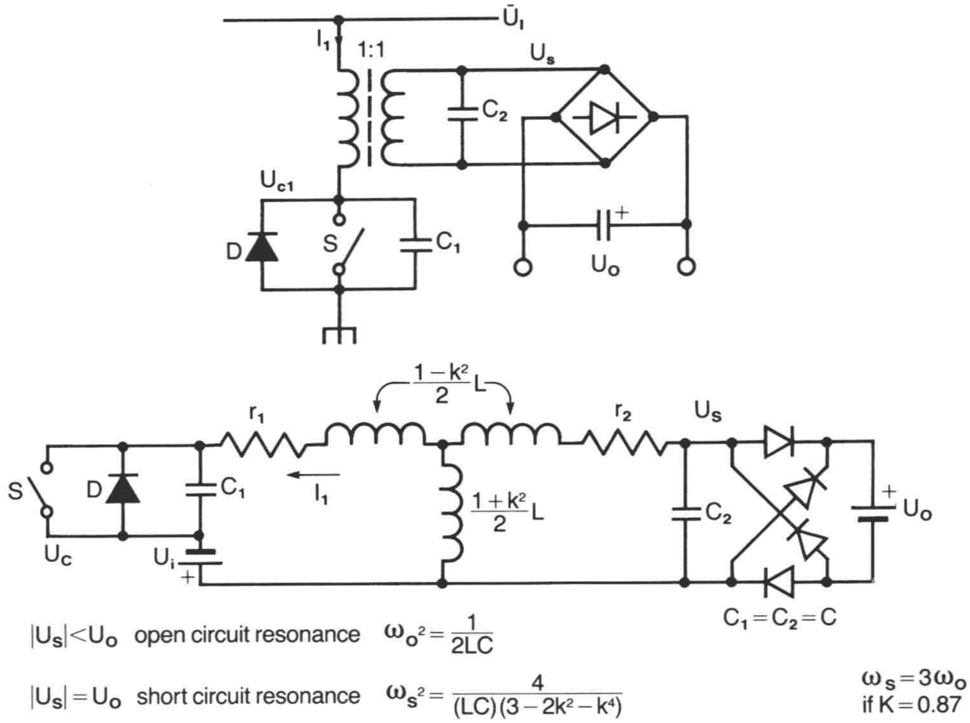


Figure 3

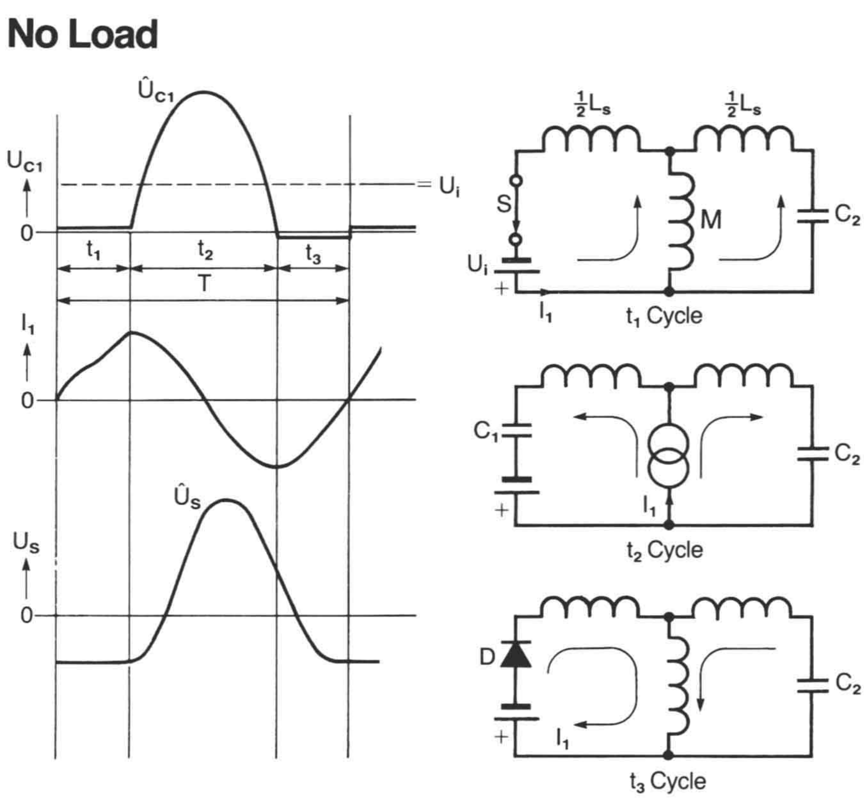


Figure 4

(...continued) making $\frac{\omega_s}{\omega_o} = 3$ and

solving for k yields

$$k = 0.87 \qquad 3)$$

Maintaing $\frac{\omega_s}{\omega_o} = 3$ with a practical $k \neq 0.87$ requires only a change of the $\frac{C_1}{C_2}$ ratio.

A resonant circuit becomes aperiodic if Q = 0.5. If we want the nominal load not to over damp the LC circuit:

$$\sqrt{\frac{L_o}{2C}} < \frac{R_L}{2} \qquad 4)$$

This assumes a constant output voltage (low ripple because of small conduction angles (<60°) and sinusoidal driving voltages. With quasi sine waves due to the non-linear damping effect of R_L, the reflected DC load before the rectifier is closer to:

$$R_{AC} \simeq \frac{2}{\pi} R_L \qquad 5)$$

Under short circuit (s.c.) conditions, the dominant frequency will be three times the fundamental frequency and the output rectifiers will conduct with a 180° conduction angle.

The s.c. current must be proportional to: $\sqrt{\frac{C_1}{L_s}}$ and U_i or

$$I_s = aU_i \sqrt{\frac{C_1}{L_s}} \qquad 6)$$

The open circuit secondary voltage U_2, will be equal to the primary voltage or:

(See fig. 4) $$\hat{U}_2 = -U_i + \hat{U}_{C1} \qquad 7)$$

$\hat{U}_{C1}$ can be calculated from the energy equation:

$$L_o\hat{I}_1^2 = C_1 U_{C1}^2 + C_2 U_2^2 \qquad 8)$$

$$2E = C_1 U_{C1}^2 + C_2 (U_i^2 + U_{C1}^2 - 2U_iU_{C1})$$

Solving for $\hat{U}_{C1}$ yields:

$$\hat{U}_{C1} = U_i \frac{C_2}{C_1 + C_2} \left\{ 1 + \left[-\frac{C_1}{C_2} + \frac{C_1 + C_2}{C_2^2} \cdot \frac{2E}{U_i^2} \right]^{\frac{1}{2}} \right\} \qquad 9)$$

With $C_1 = C_2 = C$ and enough energy supplied so that the term $\frac{C_1}{C_2}$ under the root sign may be ignored:

$$\hat{U}_{2o} = - \frac{U_1}{2} + \hat{I}_1 \sqrt{\frac{L_o}{2C}} \qquad 10)$$

If S is being closed when I_1 goes through 0 and we wait long enough for C_2 to charge to $-U_1$ via S and L_s we can write:

$$\hat{I}_1 = \frac{U_i}{L_o} t_{on} \qquad 11)$$

Combining 10) and 11) yields:

$$\hat{U}_{2o} = U_i \left(\frac{t_{on}}{\sqrt{2L_oC}} - 0.5 \right) \qquad 12)$$

From 1), we define:

$$T_o = 2\pi \sqrt{2L_oC} \qquad 13)$$

$$\text{and } t_{on} = \delta \, T_o \qquad 14)$$

Now,

$$\hat{U}_{2o} = U_i \, (2\pi\delta - 0.5) \qquad 15)$$

As long a $(2\pi\delta - 0.5) > 1$ δ min. = 0.24, only half wave rectification takes place and the circuit has properties similar to a flyback circuit, i.e.,:

a) $U_o > U_i$

b) high output impedance $\simeq Q\sqrt{\frac{L_o}{2C}}$

c) energy used per cycle $P_o \cdot (T_o + t_{on})$ must be stored in L_o first before it can be transferred

Major differences, however, are the finite values for $\frac{dU2}{dt}$ and $\frac{dI2}{dt}$ due to resonance, which in turn means lower dynamic losses in rectifiers. Also, the switch S is operated when the energy flow through it is minimum (either U_{C1} or I_1 equals 0).

The advantage of C_2 is two fold:

a) it reduces $\hat{U}_{C1}$ and allows a lower BV limit of S

b) it splits I_1 in two equal parts ($C_1 = C_2$) and reduces total losses in ($C_1 + C_2$) due to their ESRs.

Once $U_o < U_i$, the circuit allows energy transfer in the forward mode as well. While S is closed, part of I_1 is diverted to the secondary circuit of transformer T and its voltage source type sink U_o, which clamps the secondary voltage U_2. (See fig. 5)

Hence: $$|U_2| \leq U_o$$

C_2 is charged rapidly via S and L_s at a rate related to:

$$\omega_s^2 = \frac{1}{L_s\ C_2} = (3\omega_o)^2 \qquad 1)$$

The other part of I_1 can energize M and becomes available during the off period of S to charge C_1, reverse charge C_2, and to supply current to U_o. The total amount of energy required:

$$E_t > P_o t_{on} + 2C_2U_o^2 + \tfrac{1}{2}\ C_1\hat{U}_{C1}^{\ 2} + P_o t_{off} \qquad 16)$$

Part of E_t is available to resonant charge U_o via U_{C1} and L_s. It increases I_o and reduces U_{C1} by flattening its top due to the third harmonic $\omega_s = 3\omega_o$. The remaining part will reduce I_1 to zero or a finite negative value via D. This mode of operation where $U_o \simeq U_i$ permits more energy extraction out of U_i, due to energy transfer to U_o during both cycles t_1 and t_2. (Compare 2 stroke engines versus 4 stroke engines.)

The output impedance approaches a value equal to:

$$Z_o \rightarrow \frac{1}{Q}\sqrt{\frac{L_s}{C_1}} \qquad 17)$$

The circuit is sufficiently overdamped that the energy in C_2 ($=\tfrac{1}{2}C_2U_o^2$) is not sufficient to return U_{C1} to 0, i.e., S has to dissipate the energy left in C_1 ($=\tfrac{1}{2}C_1U_{C1}^2$).

It is advantageous to use a slow device for S to reduce excessive current spiking during that moment.

A third operation mode to be considered is when the output is short circuited, which effectively eliminates C_2. While S is closed, I_1 will be primarily limited by U_i, t_{on}, and the leakage inductance L_s; assuming its initial value equals zero:

(See Fig. 6) $$\hat{I}_1 = \frac{U_i}{L_s}\ t_{on} \qquad 18)$$

Nominal Load

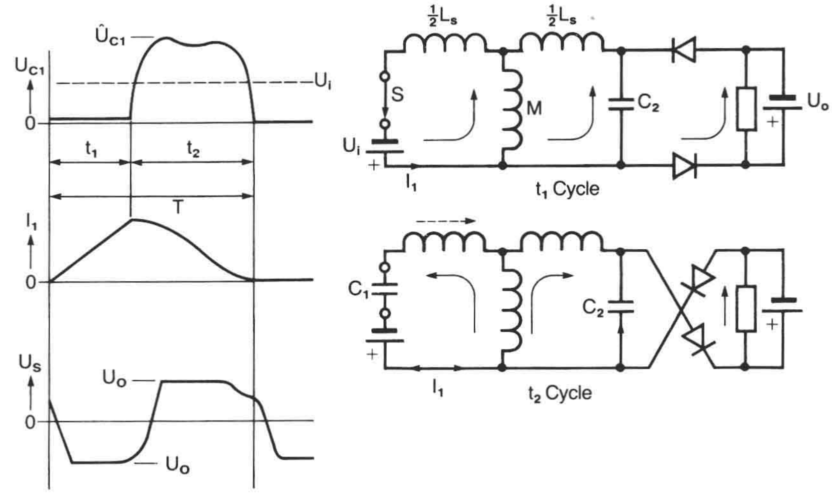


Figure 5

Short Circuit Load

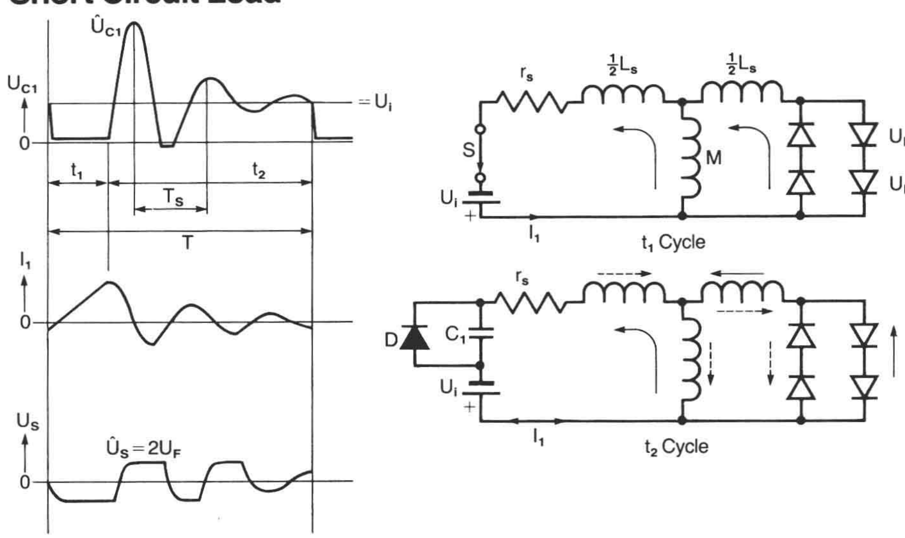


Figure 6

Low Leakage Floating Source

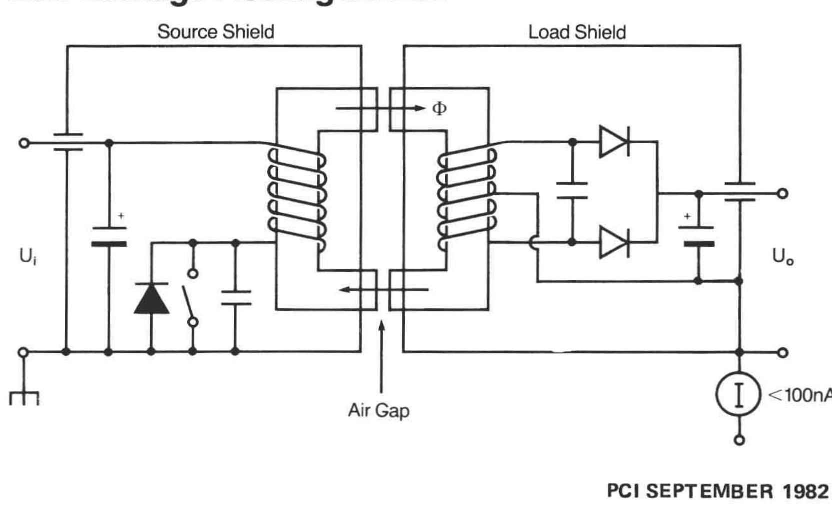


Figure 7

When S opens (t_1), this current I_1 will resonant charge C_1 to a voltage:

$$\hat{U}_o = \hat{I}_1 \sqrt{\frac{L_s}{C_1}} \qquad 19)$$

combining 18) and 19) with $T_s = 2\pi\sqrt{L_s C_1}$

$$\text{yields: } \frac{\hat{U}_{C1}}{U_i} = 2\pi \times \frac{t_{on}}{T_s} + 1 \qquad 20)$$

C_1 will subsequently discharge via the load circuit in a resonant fashion, i.e., I_1 will be cosinusoidal until $U_{C1} = 0$ and $I_1 = -\hat{I}_1$. Thereafter, it will rise linearly $\left(\frac{dI}{dt} = +\frac{U_i}{L_s}\right)$ until it crosses zero.

If we leave S open, the cycle will repeat itself and, due to a finite Q of the circuit, both U_{C1} and I_1 will exponentially decay at a rate determined by the circuit Q ($Q = \pi n$ where n is number of cycles where $|U_C| = \frac{\hat{U}_C}{e}$).

If we want to repeat the first cycle exactly, we close S again for the duration of t_{on}.

High output currents can be achieved this way at the expense of high voltage swings across S and C_1 (equation 20).

It should be recognized that this mode of operation has been used in almost every TV set to scan horizontally the electron beam of the picture CRT.

L_s represents the deflection coil and the mutual inductance M is replaced by the HV flyback transformer. The total cycle time is $T \simeq 60\mu s$ with a retrace time (S open) of 16%. This means that $\frac{U_{C1}}{U_i} = 6$ and it explains why HV transistors or SCRs are so important for 220V AC mains voltages. Due to a misunderstanding about the proper operation of this resonant type switch mode inverter, diode D is called "damper diode" in the English literature. It must be obvious that this was a misnomer. To reduce U_{C1}, third harmonic tuning was used in the HV transformer of some German designs (see ref.6) by using a $k < 0.9$ coupling factor in the design of the HV transformer.

To summarize, the single ended resonant mode converter has the following advantages:

1) most waveforms are sinusoidal reducing EMI
2) energy injection takes place at the most advantageous moment (U = low, I is low)
3) it combines the flyback with the forward mode of operation when $U_o \leq U_i$.
4) it has inherent current limiting due to L_s
5) by selecting the proper turns ratio, U_o can be properly matched to U_i
6) "snubbing" is integrated into the design
7) easier shielding due to lower k

Disadvantages are:

1) optimization is more complex due to the fact that the switching frequency must be related to the resonance frequencies of the energy transfer network
2) peak voltages and currents are high in relation to U_{in}, U_o, and I_o due to resonance phenomenae
3) core losses are difficult to relate to published data because of the unsymmetrical exitation voltage
4) fixed frequency control narrows dynamic range of P_o at a particular U_o.

Controlling the energy flow:

As we have seen, it is vital to close S after U_{C1} goes to zero or after a fixed off time period of S, which ever comes first, and before D stops conducting. ($\frac{dI_1}{dt} = +$, $I_1<0$). A current sense transformer in series with D and S will sense both conditions. The on time period has to come from the regulator circuit which senses P_o (U_o and or I_o). This method is called constant off time control and can be used on a cycle by cycle basis to improve transient response.

Constant frequency with variable but limited duty cycle can also be used. Ringing will occur during short circuit operations because the fixed switching frequency by definition does not itself adapt to the short circuit frequency, i.e., S will close after a time T_s when $U_{C1} \rightarrow U_i$. This simpler method was used during the evaluation of the resonance mode technique by using the NE 5561 control circuit (see fig. 8 to 12).

Applications:

One of the advantages of any switch mode type inverter is that reactive "power" (VAR) does not have to be dissipated, as in a linear amplifier. Both results are a consequence of Tellegen's theorem ($\Sigma U_a \cdot I_a = 0$). This means that close to unity power factors can be realized in AC motor drives and lamp ballasts. The resonant mode is particularly attractive for the latter application due to its lower EMI and a magnetic structure which permits separation of load and source which might improve heat sinking. Also, an efficient "world lamp" (100 - 230V AC) is well within its capability. In instrumentation, this type of converter is very attractive for floating power sources. (See fig. 7)

There is also a huge replacement market for constant voltage ferro resonant transformers. These machines utilize non-linear magnetics in conjunction with a fixed frequency. A resonant mode converter accomplishes the same by using linear magnetics with a variable frequency which, due to its magnitude considerably reduces weight for a given output power (1 kW/kg).