Topologies for Three-Element Resonant Converters

Rudolf P. Severns

Abstract—**Many of the limitations of two-element resonant topologies can be overcome by adding a third reactive element. However, using three elements greatly increases the number of possible topologies, making it very difficult to explore this class of circuits on a trial-and-error basis. This paper presents an orderly search procedure that exposes a large number of new topologies, many of which are resonant and have useful properties.**

I. Introduction

The series (SRC) and parallel resonant converters (PRC) are examples of topologies with two reactive energy-storage elements. These circuits are widely used but have limitations that preclude their employment in many applications.

There are several reasons for studying three-element topologies. One is to remove these limitations. Another is that parasitic L and C is often present. This can force a nominally two-element topology to actually operate as three or four elements. It is necessary to understand the behavior of these higher order topologies, which can be substantially different from two-element circuits. A less obvious reason is the insight gained from studying a new family of converter topologies even when they may not have an immediate application. Every time we add another piece to the power converter puzzle we learn something about the whole.

The SRC displays very poor load regulation except when the switching frequency (f_s) is very close to the resonant frequency (f_0). At light loads the converter must be operated in the discontinuous mode to retain control. If both the load and the input voltage vary, the range of f_s can be large and good transient response is not easily obtained.

Chen and Bonert [1] demonstrated that the addition of a third element (C2 in Fig. 1) would modify the input to output transfer function ($M = V_o/V_i$) to be substantially load independent. This circuit is known as the LCC topology. In effect, a constant reactive preload has been added that swamps out the effect of the varying real load. This is a very simple way to look at this circuit, but there are other subtler changes. Fig. 2 compares the transfer functions of the SRC and LCC topologies. The only point of load independence for the SRC occurs at $f_n = 1 = f_s/f_0$. Away from that point the load regulation is very poor. With variable f_s control the sign of the feedback

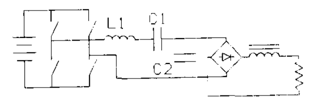

Fig. 1. LCC converter.

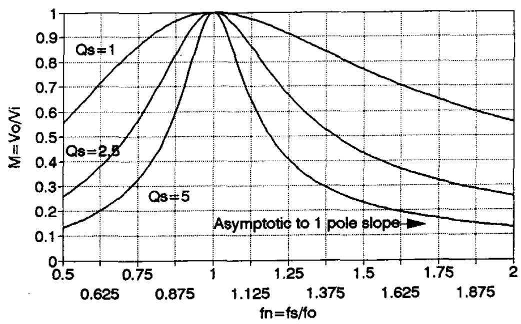

Fig. 2A. SRC transfer function.

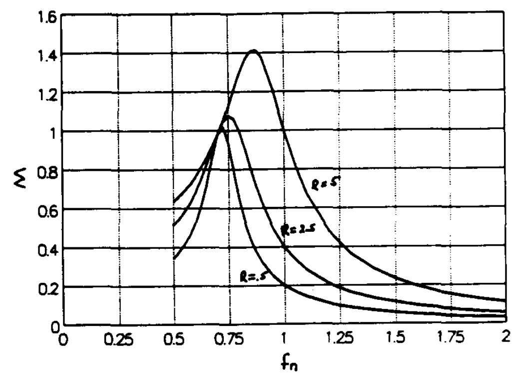

Fig. 2B. LCC transfer function.

loop changes as you pass through f_0 so that it is necessary to operate on one side or the other of f_0. In the LCC circuit the load-independent point is below resonance and by suitable choice of Q and the ratio of C1/C2, it is possible to operate from zero to full load with a small range of f_s. The added capacitor also increases the filter slope above f_0 from one to two poles, providing much better high-frequency ($>f_0$) filtering. These are significant improvements that allow the SRC to be used in many more applications. This and additional useful behavior is present in other three-element topologies, both above and below f_0.

Three-element topologies have their own set of drawbacks. First, there is the need for an additional reactive energy-storage component. This can effect the size, cost,

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and efficiency of the converter. The current in the additional element can increase the currents in other components, increasing losses. With care this problem can be minimized, but there will be a penalty. In some three-element topologies, the resonant point of the small-signal transfer function varies with load as indicated in Fig. 2(b). This can be a nuisance in a control circuit.

Because of the improved filter rolloff characteristic it is often possible to operate three-element topologies with a Q different from two-element circuits. This can actually allow a reduction in component currents. A large variety [1], [6], [7], [9]–[19] of three-element circuits have been built and, in practice, efficiency has not proven to be a serious issue. Three-element topologies are certainly not a panacea for converter ills, but they can be useful.

II. Search Procedure

It has been suggested [2], [3] that power converters can be partitioned into several functional blocks, as shown in Fig. 3. For the purposes of a resonant topology search, the number of blocks can be reduced as shown in Fig. 4. The three blocks are an ac voltage or current source and an ac voltage or current load, connected by a network N that may contain any number ($n = 0, 1, 2, 3, \cdots$) of reactive energy storage elements. N could also contain switches, but that will not be addressed in this paper. An example of this concept is given in Fig. 5 for the SRC. The battery (V_i), C1, and the switches constitute an ac voltage source. L_r and C_r form a two-element network N. The rectifier bridge C2 and the load form an ac load.

The search procedure consists of first examining the different types of sources, loads, and interconnecting networks. Second, the possible combinations of these elements are explored. For resonant converters this approach has proven to be very useful. As will be shown shortly, it also provides interesting insights into switch mode and other topologies. However, no claim is made that this concept is universally applicable. With modification, the procedure may prove to be general, but that has *not* been established.

III. Sources and Loads

Practical sources and loads are never ideal generators, but the ideal generator does represent a reference against which practical generators can be judged. The wording in the following definitions is equivalent to that in a standard circuits text but has been restated to suit this discussion.

There are two classes of ideal sources or generators: voltage and current. An ideal voltage generator has the following properties:

1) V_o is independent of the load.
2) The amplitude and polarity of I_o is determined by the response of the load to V_o. I_o may be either + or −.
3) The output impedance (Z_o) is zero. This implies that when $V_o = 0$, the generator is a short circuit.

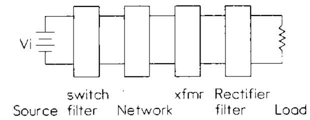


Fig. 3. Partitioned dc/dc converter.

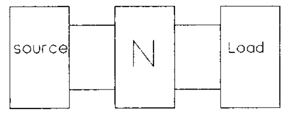


Fig. 4. Simplified block diagram.

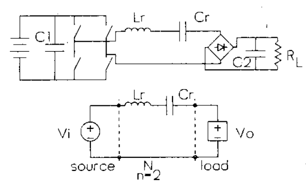


Fig. 5. Modeling example for SRC.

4) The generator may either source or sink energy.
5) Energy is not dissipated internal to the generator—it is 100% efficient.

An ideal current generator has the following properties:

1) I_o is independent of the load.
2) V_o depends on the load response to I_o.
3) Z_o is infinite. When $I_o = 0$, the generator is an open circuit.
4) The generator will source or sink energy.
5) The generator is 100% efficient.

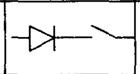
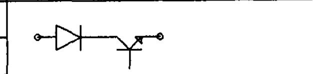
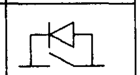
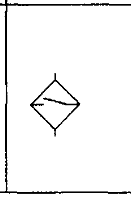
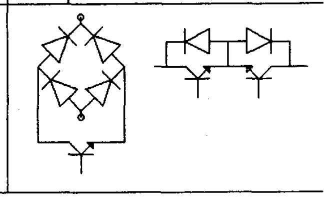
Practical generators can only approximate the ideal. Some are closer than others and the differences are important to circuit operation.

The requirement for high efficiency is universal in power converters. For this reason practical generators are realized with nondissipative components such as switches, capacitors, inductors, and voltage sources. Very few true current sources exist in practice. One example of a natural current source would be a solar array at high loads. However, because the array has a very large self-capacitance at frequencies of more than a few hundred hertz it becomes a voltage source. In practice current sources are approximated using a voltage source and an inductor.

A switch can only block voltage or conduct current. These actions may be either uni or bidirectional. This leads to the matrix shown in Table I. These are the four types of switches available for synthesizing generators. The table also gives the number of operating quadrants and the symbols that will be used. The figures in the table show an ideal switch. In reality this switch would be a BJT, MOSFET, etc.

Fig. 6 gives two examples of nearly ideal ac voltage sources. Because the impedance of real voltage sources

TABLE I

block voltage	conduct current	quads	device	
uni	uni	1		
bi	uni	2		
uni	bi	2		
bi	bi	4		

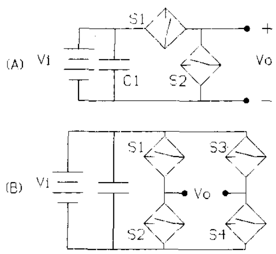


Fig. 6. Near ideal voltage sources.

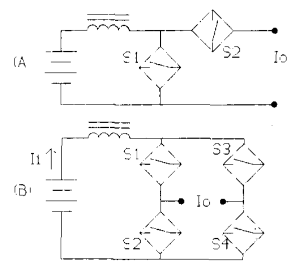


Fig. 7. Near ideal current sources.

(V_i) is usually low only at low frequencies and pulsating input current is not usually acceptable, a capacitor (C1) is added. In the unipolar generator (6A) either S1 or S2 will be on at any time. Both cannot be on simultaneously as that would short circuit the power source, a condition that is not usually allowed. If the duty cycle of S1 is D then the duty cycle of S2 is $D' = 1 - D$. An examination of this generator shows that it is very nearly ideal. When S1 is on $V_o = V_i$ independent of I_o. Because the switches are four quadrant, I_o may be either + or −. When S2 is on, $V_o = 0$ and $Z_o = 0$ within the limits of the components. The output waveform is limited to quasi square waves, but if a source other than the indicated battery were used it would be possible to have other waveforms such as chopped sinusoids, etc.

The bipolar source (6B) behaves in a similar fashion, but it allows the output to be either pure ac or pulsating dc of either polarity. In this circuit when $V_o = 0$, for $Z_o = 0$ it is necessary that either S1 and S3 or S2 and S4 be on. Within component and waveform limitations both of these sources are very nearly ideal. If this is not done when used in a resonant converter, some energy will be returned to the source during each switching cycle. This excess energy appears as higher switch, diode, and network currents.

Practical current sources can also be realized as shown in Fig. 7 by adding an inductor in series with V_i. These sources are also nearly ideal with one very important exception. On an average basis the energy delivered to the load is equal to the energy supplied by the source (ignoring circuit losses). The peak value for I_o is the same as I_i, which is determined by source power, which in turn is determined by the load! Averaged over several cycles, I_o is not independent of the load as it would be in an ideal generator. On a cycle-to-cycle basis it does act like an independent current source. In contrast, in a practical voltage source a high degree of load independence can be achieved.

In practice, near ideal sources are very often not required or even desirable. For example, controlled bidirectional power flow is frequently not required. This allows the use of two quadrant switches. Figs. 8 and 9 show simplified, but very practical, voltage and current sources. Fig. 8(a) could use two switches and diodes or could be as simple as a single switch, depending on the application. Other switch connections such as the half bridge and push–pull shown in Fig. 10 can also be used, but it is often not appreciated that these are inherently less ideal. For example in the full-bridge circuit (Fig. 8(b)) Z_o can be made zero by having S2 and S4 on while $V_o = 0$. In the half-bridge circuit there is no way to provide this characteristic because both switches cannot be on simultaneously. To force $Z_o = 0$ it is necessary to add another switch as indicated by the dashed lines. The same prob-

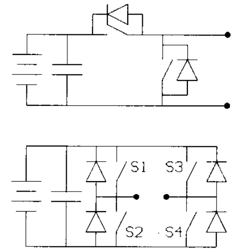


Fig. 8. Practical voltage sources.

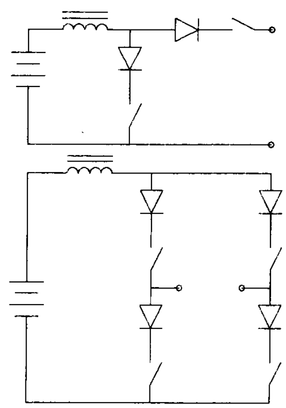


Fig. 9. Practical current sources.

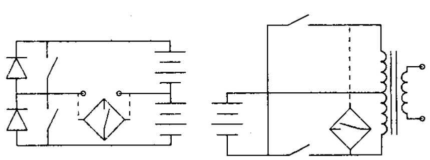


Fig. 10. Modified voltage sources.

lem exists for the push–pull connection. Many circuits do not require that $Z_o = 0$ when $V_o = 0$, but in many others, particularly resonant converters, it is desirable to prevent energy from recirculating back to the source.

Very similar arguments can be made for loads. Fig. 11 gives several examples of practical voltage and current loads. They may be nearly ideal as shown in (a) and (b) or very simple as shown in (c) and (d). All of these are seen in practice depending on the application. Loads with four-quadrant switches are particularly useful if active control of the load is desired or if bidirectional power flow is needed. A wide variety of other single- and polyphase rectifier connections are known [4] and could be used. By the use of the bilateral inversion procedure [5] all of the rectifier connections become switch connections.

For the purposes of this discussion the source and load symbols given in Fig. 12 will be used. The sources are independent generators and the loads are either a voltage controlled current source or a current-controlled voltage source. The dependency relationships are as indicated where K is an optional control factor.

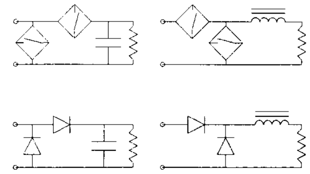


Fig. 11. Practical load examples.

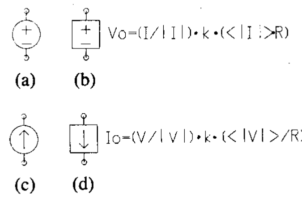


Fig. 12. Idealized loads and sources. (a) *V* load. (b) *I* load. (c) *V* load. (d) *I* load.

IV. Examples of Combinations

Given two types of sources and two types of loads there are four possible combinations with an arbitrary N as shown in Fig. 13. The combinations are *V-N-V*, *I-N-I*, *V-N-I*, and *I-N-V*.

Typical examples of $n = 0$ circuits using unipolar (or halfwave) sources and loads are given in Fig. 14. These are the well-known buck (*V-N-I*) and boost (*I-N-V*) topologies. Note that for $n = 0$ and unipolar sources and loads, the *V-N-V* and *I-N-I* combinations do not form functional converters.

I-N-I converters are of interest because of the nonpulsating nature of the input and output currents. Examples of $n = 1$, *I-N-I* topologies are given in Fig. 15. Fig. 15(a) is the Ćuk converter and (b) is a topology that uses a single-shunt capacitor for N. Both of these are *I-N-I*. It is widely believed that the Ćuk converter is the only topology that has an *I-N-I* character. This is not correct. Many other examples already exist [6] and the search for three-element resonant topologies generates many more. The SRC topology (Fig. 5) is an example of an $n = 2$, *V-N-V* topology.

V. Three-Element Networks

There are four possible classes of three-element networks containing only L and C: 1) $2L$ and $1C$, 2) $1L$ and $2C$, 3) $3L$, and 4) $3C$.

The 14 network topologies that do not obviously reduce to less than three elements are given in Fig. 16. Combining these topologies with the four combinations of components results in the networks shown in Figs. 17 and 18. These are the three-element networks available for interconnecting sources and loads. The network ID numbers are used in Table II to identify the combinations.

VI. Definitions and Restrictions

Before examining the three-element circuits it is necessary to make a few definitions. The purpose of this to-

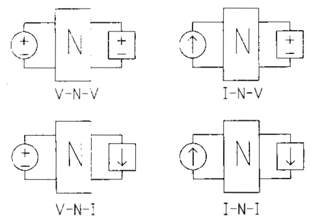


Fig. 13. Source-network-load combinations.

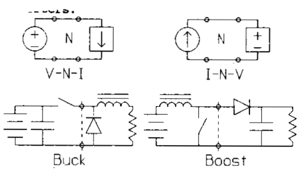


Fig. 14. $n = 0$ examples.

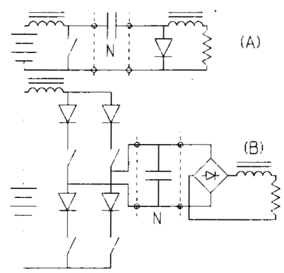


Fig. 15. $n = 2$ *I-N-I* examples.

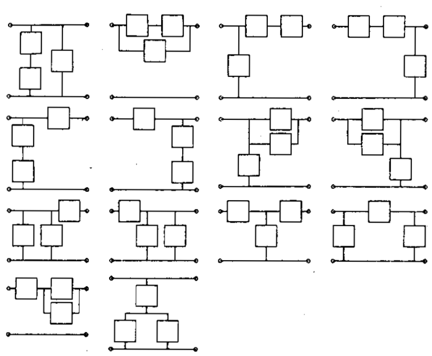


Fig. 16. Three-element network topologies.

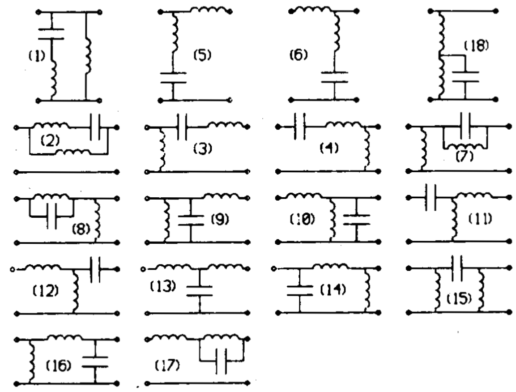


Fig. 17. Networks with $2L$ and $1C$.

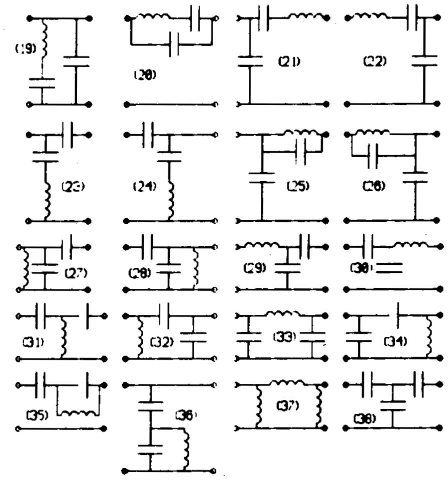


Fig. 18. Networks with $2C + 1L$, $3C$, and $3L$.

TABLE II
THREE-ELEMENT CONVERTER COMBINATIONS

Network Number	Topology	Network Number	Topology
1	*VNI*	20	*INV*
1	*INV*	20	*VNI*
2	*VNV-R*	21	*INV-R*
3	*VNV-R*	22	*VNI-R*
4	*VNV-R*	23	*INV*
5	*VNV-R*	23	*VNI*
6	*VNV-R*	24	*INV*
7	*VNI*	24	*VNI*
7	*INV*	25	*INI-R*
8	*VNI*	26	*INI-R*
8	*INV*	27	*INI-R*
9	*INV-R*	28	*INI-R*
10	*VNI-R*	29	*VNI-R*
11	*VNV-R*	30	*INV-R*
12	*VNV-R*	31	*INV*
13	*VNV-R*	31	*VNI*
14	*INV-R*	32	*INI-R*
15	*INV*	33	*INI-R*
15	*VNI*	34	*INI-R*
16	*VNI-R*	35	*INV*
17	*VNV-R*	35	*VNI*
18	*INV*	36	*INI-R*
18	*VNI*	37	*VNV*
19	*INI-R*	38	*INI*

pology search is to find three-element "resonant" topologies. Surprisingly, we do not have a single, generally accepted definition of what a "resonant" converter is. The following provisional definition will be used:

1) The power transfer from input to output is primarily via the fundamental component of f_s. The harmonics of the source and load contribute little to the power transfer. What this implies is that N acts as either a band-pass or low-pass filter, isolating the input and output at harmonics of f_s.

2) The waveform of the voltage or current response of N to the excitation of the source and the load is piecewise sinusoidal.

This definition will be used to decide which topologies are resonant. It is not implied that topologies that do not meet this criterion are not usable. They may well be useful but are another class of circuit. An example of this distinction is given in the section on other topologies.

One of the advantages of resonant converters is that they can be analyzed using the sine-wave approximation. This allows the circuit to be simplified as shown in Fig. 19. Even though the actual source may be square wave, because the harmonics supply little energy the source waveform can be approximated as a sine wave. Similarly, there is little power in the load harmonics so that the load can be replaced with a resistor of the appropriate value as shown by Steigerwald [7]. By using this approximation an equation for the transfer function in the S domain can easily be derived. These transfer functions can then be readily graphed using a spreadsheet program. The same procedure can be used to approximate the input and output impedances and to determine the component stresses.

Some restrictions must be placed on the input and output impedance of N for different types of sources and loads. For example, if a squarewave voltage source is connected directly to a shunt capacitor (part of N) large current spikes will flow during the switching transitions. If an inductor is connected in series with the output of a square-wave current source there will be large voltage spikes. Neither of these is usually acceptable. It will be shown later that this restriction can be eased in quasi- and multiresonant topologies due to the nonideal nature of the source. For resonant converters the restrictions on N will be:

1) For voltage sources or loads, the impedance seen by the source or load must be high at harmonics of f_s.
2) For current sources or loads, the impedance seen by the source or load harmonics must be low.
3) Only those networks for which the transfer function is either low or band pass will be considered as resonant.

The direction of the switch current at turn-on and -off is an important parameter for choosing a resonant topology. Vorperian [8] has defined two continuous conduction modes, plus (+CCM) and minus (−CCM). In +CCM operation the switch current is positive when the switch is turned on. −CCM operation implies that the switch current at turn-on is negative. For voltage source driven topologies, where N usually is inductive above resonance, the voltage leads the current for $f_s > f_0$. This means −CCM operation and low turn-on switching stress. For −CCM operation, the turn-off stress is usually high, comparable to an SMPC. The advantage of −CCM operation is the ease with which a low loss resonant transition snubber can be implemented. Very often the parasitic capacitance of the switches along with a drive-signal dead

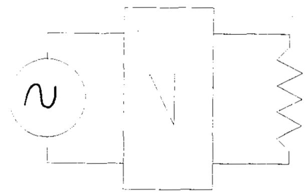


Fig. 19. Sine-wave approximation.

time is all that is required to have very low loss switching. This is very desirable for high-frequency operation. +CCM operation is characteristic of $f_s < f_0$ operation with voltage source driven topologies. It is particularly desirable for thyristor switches as the topology usually provides commutation without additional components.

VII. Three-Element Converter Topologies

Applying restrictions 1 and 2 to the combinations of Fig. 13 and the networks in Figs. 17 and 18 results in 48 functional topologies, which are listed in Table II. It is not probable that all 48 will have desirable or even useful properties, but they should all convert power. Note that 10 of the combinations have nonpulsating input and output currents (*I-N-I*).

Of the 48 topologies, 26 meet the criterion of "resonant." These are indicated in Table II by the -R suffix. Nine of the *I-N-I* topologies are resonant, and all 12 of the *I-N-I* topologies are resonant. Of the 26 resonant topologies, only a few have previously appeared in the literature. Those of which the author is aware are shown in Fig. 20 along with references to the papers in which they appeared.

VIII. Typical Three-Element Converters

It is not possible to address all 26 resonant topologies in this paper, but several examples that illustrate typical behavior are discussed. The LCC topology shown in Fig. 1 is sometimes referred to as "series-parallel" because it is a combination of the two. There are two ways to look at the topology: it can be viewed as an SRC to which a shunt capacitor (C2) has been added or as a PRC with an added series capacitor (C1). Each point of view gives a different insight to the circuit operation.

If C1 is large ($k = C2/C1 = 0$) the circuit reduces to a two-element PRC. The transfer function for $k = 0$ is given in Fig. 21. It can be seen that for $f_n = 0.5–1.5$ the load regulation is relatively poor, especially near $f_n = 1$. Note that the lowest value of f_n is 0.5. Values of $f_n < 0.5$ lead either to multiple cycle ringing or discontinuous-mode operation [8]. Neither of these is generally desirable. If good load regulation is desired then this topology is normally operated with $F_n > 1.5$. Although this is often useful at low power levels, when the power level rises to the point where thyristor switches are needed, operation with $f_n > 1$ requires the use of impulse commutation because the operating mode is −CCM. This form of com-

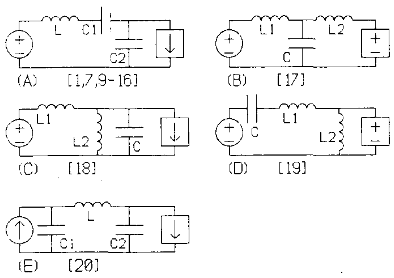


Fig. 20. Published topologies.

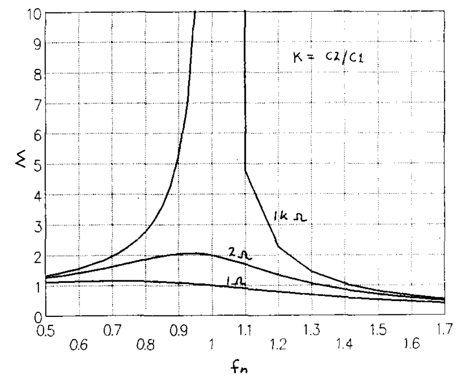


Fig. 21. M for LCC topology.

mutation is undesirable due to the size and number of commutation components. If $f_n < 1$, then the circuit is naturally commutated (+CCM) without additional components but the load regulation is poor. For $f_n < 1$, the load regulation can be improved by making C1 smaller. Figs. 22–24 show the effect of different values of k. For $k = 0$ (C1 >> C2) the regulation is very poor but as C1 is made progressively smaller ($k = 0.5 - 1$), load-independent operation appears. Note that the load range is 1000 : 1! For some applications this is very useful behavior.

What price is paid for these "improvements"? First, there is an additional power capacitor that must carry the full current in L. Second, as k is increased (C1 made smaller) the voltage across C1 rises. For $k = 1$, $V_c > V_o$ and this can limit the usable values of k. Certainly there will be a cost, and perhaps efficiency, penalty that must be justified by the improved circuit performance.

If the topology is viewed as an SRC to which a shunt capacitor is added, then the penalty is more obvious. Assuming that V_o is constant, increasing the value of C2 increases the effective load current. This in turn appears as higher current in L, C1, and the input switches. In general C2 is made as small a possible to minimize the increase in current. Fortunately the load and C2 currents are in quadrature so that the inductor current is not the linear sum of the two. By minimizing k and carefully selecting the full-load circuit Q, it is possible to keep the increase in switch current to a factor of 1.5 or less. Similar effects

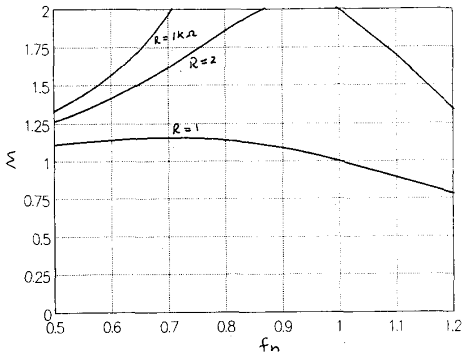


Fig. 22. LCC, $k = 0$, $R = 1, 2, 1000\ \Omega$.

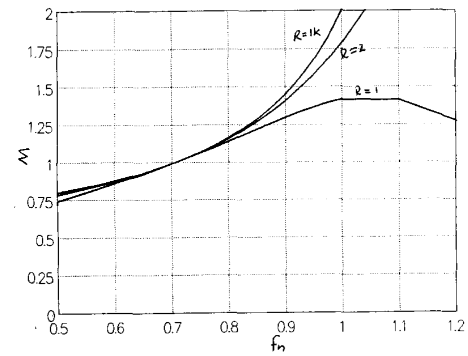


Fig. 23. LCC, $k = 0.5$, $R = 1, 2, 1000\ \Omega$.

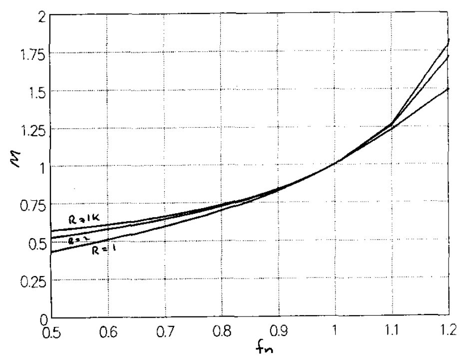


Fig. 24. LCC, $k = 1$, $R = 1, 2, 1000\ \Omega$.

are present in some other three-element topologies, but in general, the current increase can be controlled. The tradeoff is the ability to operate with zero load in a continuous conduction mode, with better HF filtering, versus the penalty of increased component current.

In the LCC topology the region of load independence is around $M = 1$. For some applications it is desirable to have this happen for $m > 1$. The topology search has turned up a variation of the LCC topology that does this. An example is shown in Fig. 25. In this topology the shunt and series capacitors are interchanged. The transfer function for this circuit is given in Fig. 26. The minimum

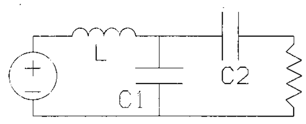


Fig. 25. Alternate LCC topology.

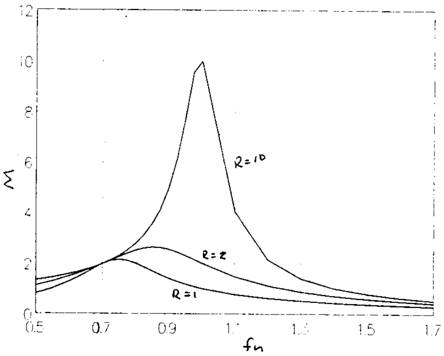


Fig. 26. M vers f_n, $k = 1$, $R = 1, 2, 10\ \Omega$.

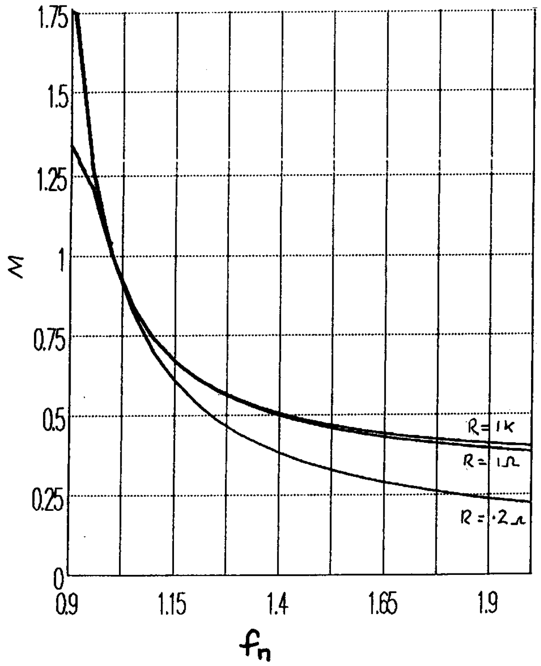


Fig. 27. LLC, $k = 2$, $R = 1, 2, 1000\ \Omega$.

load-dependence point occurs at $M = 1 + 1/k$, rather than $M = 1$.

For applications where thyristor switches are not used, $-$CCM operation is often preferred. The LCC topology and the variation in Fig. 25 are not attractive for $-$CCM operation. The topology shown in Fig. 20(d) would be a better choice. In that topology, an inductor is added rather than a capacitor.

A typical graph of M is given in Fig. 27. We see that this topology also displays load-independent operation, but now it is on the high side of resonance. For a load range of 1000 : 1 the regulation is very good. By the right choice of topology, load independence can be had either above or below f_o.

When a transformer is added to a PRC the transformer magnetizing inductance (L_m) is added in parallel with the resonating capacitor, forming a three-element topology like that shown in Fig. 20(c). A graph of M (L1 = L2) is shown in Fig. 28. The primary effect of the shunt inductance (L2) is to raise f_0 and reduce the peak amplitude of the response. It is unlikely that L_m would be this small, however. A more normal effect would be to push f_0 up by a small amount.

What about the effect of transformer leakage inductance? Adding L2 to the PRC network results in the topology given in Fig. 20(b). The transfer function for L2 = 2L1 is given in Fig. 29. We see that if L2 is large a load independent region forms, very much like the LCC topology. This large a value for L2 would be unusual, however, unless it was deliberately designed in. For more normal values the effect of leakage inductance is to introduce a small load dependance in f_0. It is interesting that an inductor in series with the load produces the same effect as a capacitor in series with L1.

The examples given represent reasonable converter

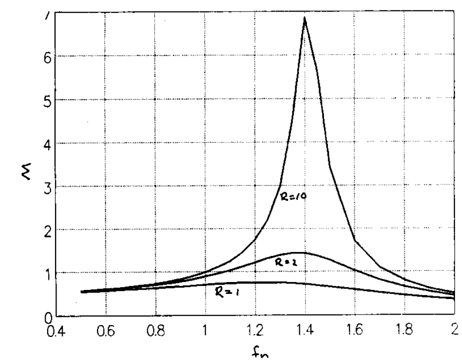


Fig. 28. Fig. 20(c), $k = 1$, $R = 1, 2, 10\ \Omega$.

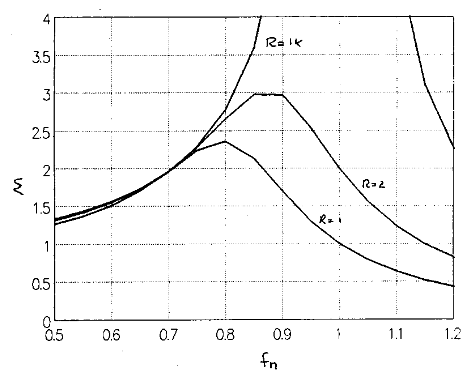


Fig. 29. Fig. 20(b), $k = 2$, $R = 1, 2, 1000\ \Omega$.

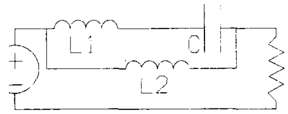


Fig. 30. Another LLC topology.

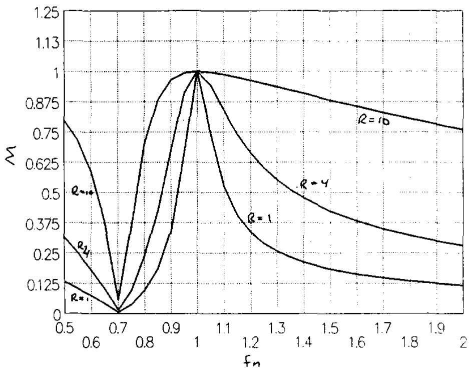


Fig. 31. Fig. 30, $k = 1$, $R = 1, 4, 10\ \Omega$.

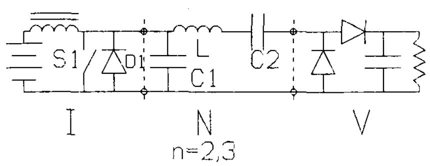


Fig. 32. Class E converter.

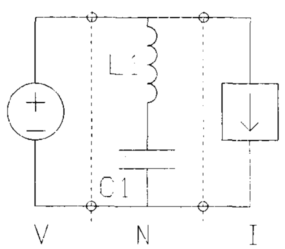


Fig. 33. Resonant commutation topology.

possibilities. Some three-element networks produce bizarre transfer functions. An example of this is the topology shown in Fig. 30, with its transfer function given in Fig. 31. It is not obvious what this topology would be useful for even though it is a functional power converter.

IX. Other Topologies

The class E topology [21] shown in Fig. 32 is an example of a three-element, multiresonant topology. It is called multiresonant because during the switching cycle C1 is periodically removed from the circuit (when S1 or D1 are conducting) so that $n = 2$ or 3 depending on the point in the switching cycle. This causes f_0 to have two different values during the switching sequence, hence the circuit is multiple or multiresonant. A quasi-resonant topology would be one where $n = 0$ or 1 during part of the sequence and 2 during other portions. Note that in this circuit there is a switch directly across a capacitor that violates the earlier stated requirement. It is acceptable in this case because the switch only closes when the voltage across the capacitor is zero.

Earlier, the definition of "resonant" was reserved for only those topologies that transferred energy at the fundamental of fs. Many of the topologies in Table II will have sinusoidal currents flowing but do not meet this restrictive condition. An example illustrating this point is the two-element topology shown in Fig. 33. This is nothing more than a PWM buck converter with L1 and C1 added to provide commutation currents for the source switches. This commutation scheme is well known to thyristor users. In this example N clearly does not act as a filter and the source voltage harmonics do contribute significantly to the output. A number of the three-element topologies are similar to this example and others act as high-pass filters so that circuit does not qualify as "resonant," but there will still be sinusoidal currents flowing. We have a nomenclature problem. What do we call this class of circuits?

X. Conclusions

A procedure for an orderly examination of multielement resonant topologies has been demonstrated. With this procedure a large number of new three-element topologies have been exposed. Many of the new topologies have interesting and potentially useful properties.

References

[1] J. Chen and R. Bonert, "Load independent ac/dc power supply for higher frequencies with sinewave output," *IEEE Trans. Industry Applications*, vol. IA-19, no. 2, pp. 223-227, Mar./Apr. 1983.

[2] R. Severns, "Generalized topologies for converters with reactive energy storage," *IEEE IAS Conf. Rec.*, Oct. 1989, pp. 1147-1151.

[3] R. Liu and C. Q. Lee, "Topological constraints on basic PWM converters," *IEEE PESC Conf. Rec.*, Apr. 1988, pp. 236-244.

[4] G. Seguier, *Power Electronic Converters: AC/DC Conversion*. New York: McGraw-Hill, 1986.

[5] R. Severns, "Techniques for designing new types of converters," *PCI Conf. Rec.*, Munich, Germany, Sept. 1979.

[6] —, "Generalized converter topologies with nonpulsating input and output currents," *High Frequency Power Conversion (HFPC) Conf. Rec.*, Santa Clara, CA, May 1990, pp. 223-234.

[7] R. Steigerwald, "A comparison of half-bridge resonant converter topologies," *IEEE IAS Conf. Rec.*, City, Oct. 1987, pp. 135-144.

[8] V. Vorperian and S. Ćuk, "A complete dc analysis of the series resonant converter," *IEEE PESC Conf. Rec.*, City, June 1982, pp. 85-100.

[9] C. Q. Lee, I. Batarseh, and R. Liu, "Design of the capacitive coupled LCC-type parallel resonant converter," *IEEE IECON '88 Conf. Rec.*, City, Month, pp. 672-677.

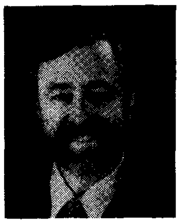
[10] A. K. S. Bhat and S. Dewan, "Analysis and design of a high frequency converter using LCC-type commutation," *IEEE IAS Conf. Rec.*, Oct. 1986.

[11] A. K. S. Bhat, "Fixed frequency PWM series-parallel resonant converter," *IEEE IAS Conf. Rec.*, Oct. 1989, San Diego, CA, pp. 1115-1121.

[12] I. Batarseh, R. Liu, Lee, and A. K. Upadhyay, "150W, 140 kHz multi output, LCC-type parallel resonant converter," *IEEE APEC Conf. Rec.*, Baltimore, MD, 1989, pp. 221-230.

[13] K. Ngo, R. Steigerwald, T. Yerman, and M. Kuo, "A high-density resonant power supply using high-voltage IC," *HFPC Conf. Rec.*, San Diego, CA, Apr. 1987, pp. 244-257.

[14] D. V. Jones, "A new resonant converter topology," *HFPC Conf. Rec.*, City, Apr. 1987, pp. ?-?.
[15] A. K. S. Bhat, "A resonant converter suitable for 650V dc bus operation," *IEEE APEC Conf. Rec.*, 1989, pp. 231-239.
[16] Fischer, Korman *et al.*, "Performance of low loss synchronous rectifiers in a series-parallel resonant dc-dc converter," *IEEE APEC Conf. Rec.*, 1989.
[17] Seidel, "A high power factor tuned class D converter," *IEEE PESC Conf. Rec.*, 1988, pp. 1038-1042.
[18] Carsten, "A hybrid series-parallel resonant converter for high frequencies and power levels," *HFPC Conf. Rec.*, 1987.
[19] Schmidtner, "A new high frequency resonant converter topology," *HFPC Conf. Rec.*, 1988, pp. 390-403.
[20] Tabisz and Lee, "Zero-voltage-switching multi-resonant technique—A novel approach to improve performance of high frequency quasi-resonant converters," *IEEE PESC Conf. Rec.*, 1988, pp. 9-16.
[21] Sokal and Sokal, "Class E—A new class of high efficiency tuned single ended switching power amplifiers," *IEEE J. Solid State Circuits*, vol. SC-10, no. 3, pp. ?-?, June 1975.



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