A FOUR-QUADRANT MAGNET TRIM POWER SUPPLY

R. J. YAREMA

I. INTRODUCTION

A four-quadrant power supply is a power supply which can supply a bipolar output voltage and source or sink current with either polarity output. Figure 1 shows the four quadrants in which the power supply can operate. In quadrants I and III the power supply provides energy to the load; in quadrants II and IV the power supply absorbs energy from the load.

One application of a four quadrant supply is to adjust (increase or decrease) the current in one or more elements of a series connection of magnets as shown in Figure 2. It should be noted that with this type of supply, the correction can be done under DC conditions or plus or minus ramped conditions in the main bus. Furthermore, the approach works with conventional or superconducting magnets. It may be noted in Figure 2 that the trim supply also tries to drive the remaining magnets in the series connection. However, as long as the main supply is current regulated, the main supply current cannot change and the trim power supply only affects the current through the element to which the supply is connected.

Different approaches may be taken to the power supply design for different voltage and current requirements. For low voltages (10V) and low current applications, a transistorized supply may be considered up to the point where the transistor banks become excessively large. When the supply is required to absorb energy from

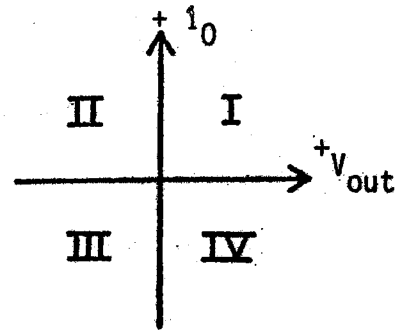


Figure 1 - Operating Regions of a Four-Quadrant Power Supply

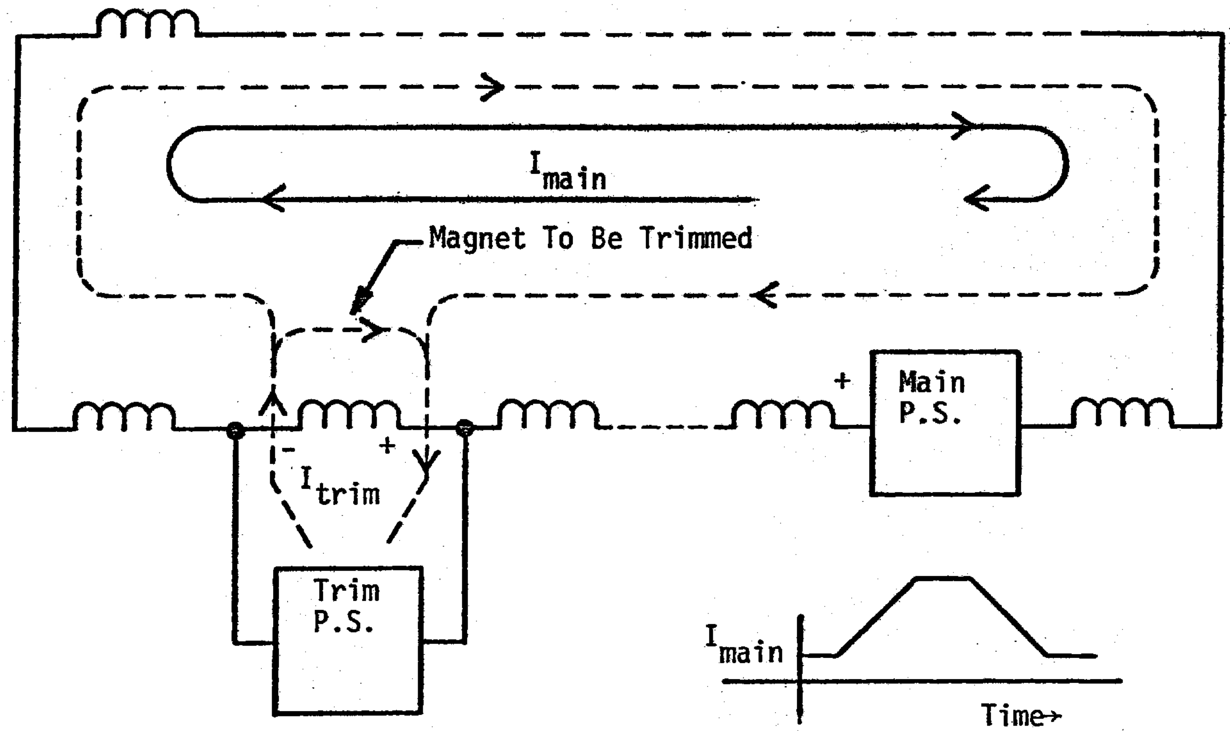


Figure 2 - Series Connection of Magnets with Trim Supply (Operating in Quadrant IV)

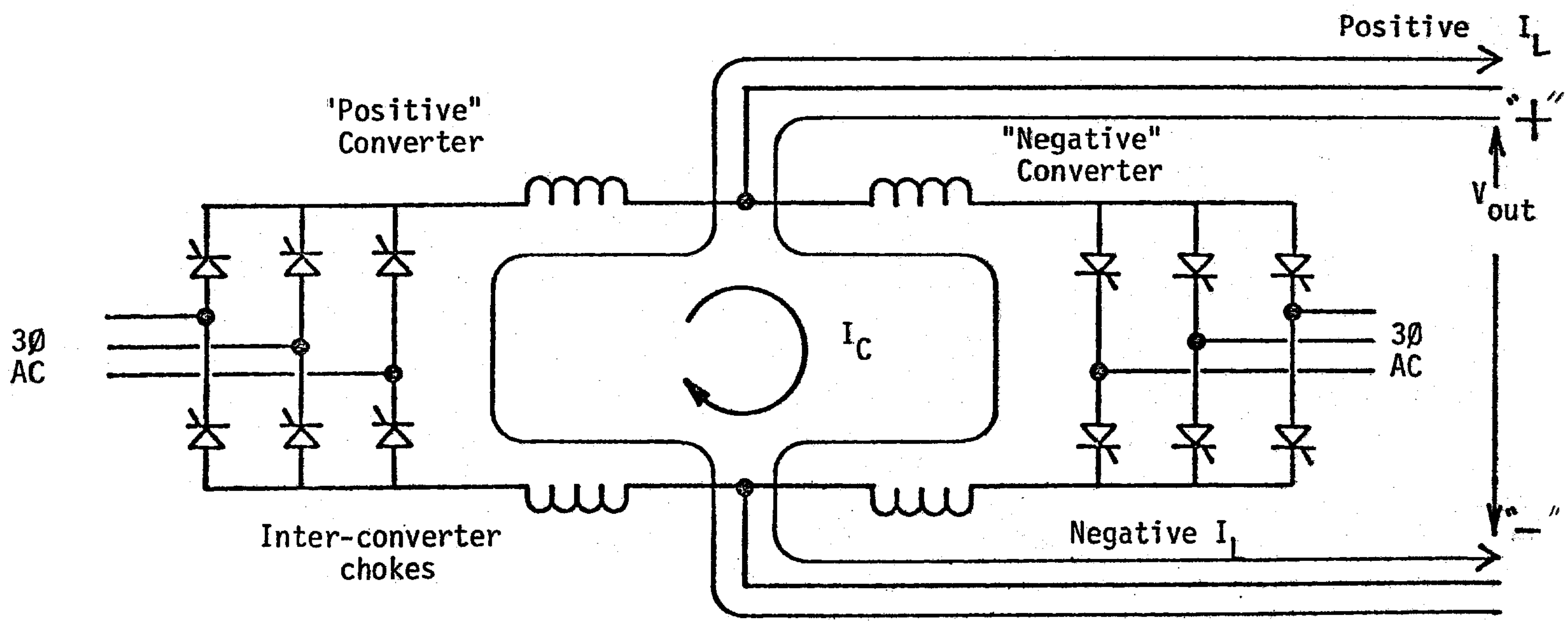


Figure 3 - Simplified General Dual Converter

the load, the transistor banks dissipate the power. For higher voltages (>100V) and higher power applications, SCR-controlled rectifiers and inverters may be more practical. When energy is to be absorbed by the supply, the SCR's can operate in the invert mode and feed power back into the AC mains rather than dissipate the power within the supply. In very large power applications, the SCR approach may provide some energy savings.

The application considered in this paper is for a $\pm$200V, $\pm$20A power supply which floats +2KV above ground, has moderate bandwidth and low output ripple. Because of the voltage and power requirements, the SCR approach was chosen. This paper will discuss the general and practical design considerations of an SCR-controlled four quadrant power supply to meet those requirements. In the literature this type of supply is often referred to as a dual converter (D/C). Historically, dual converters have been used for reversible DC motor drives and recently they have received some attention for accelerator applications.

II. DUAL CONVERTER PRINCIPLES

A dual converter is comprised of two SCR bridge converters. Each bridge is capable of operating either as a rectifier or as an inverter. Figure 3 shows a generalized dual converter. For purposes of discussion one converter is called the positive converter and the other, the negative converter, with the output polarity marked accordingly. (Since the converters are identical, the identification of the converters and output is arbitrary). The positive converter is capable of operating in quadrants I and II and only of supporting positive load current. The negative converter is capable of operating in quadrants III and IV and only of supporting negative load current. Thus by combining the outputs of the two converters, power supply operation in four quadrants is obtained.

<u>Operating Modes:</u> The dual converter of Figure 3 is commonly operated in two different modes.

1. The circulating current-free mode.
2. The continuous circulating current mode.

With the circulating current-free mode only one SCR converter operates at a time with the gating pulses to the other converter suppressed or blocked. When a load current reversal is required, a logic system detects the load current passing through zero, blocks the converter which was conducting and then after an appropriate delay, unblocks the other converter. In the circulating current-free mode the inter-converter chokes, shown in Figure 3 are not required. With the circulating current mode, both the positive and negative converters are operated simultaneously. The firing angles of the two converters are controlled in such a way that the average voltage at the output of each converter is essentially the same. Equal average output voltages is accomplished by setting the positive and negative converter firing angles so that $\alpha_p + \alpha_n = 180^o$ (see Appendix A). Thus at all times one converter is acting as a rectifier and the other as an inverter. Although the average converter voltages are the same, instantaneous voltage differences exist which can drive large AC currents between the converters. To limit the AC current which would flow, chokes (impedances) are placed between the converters. In the continuous circulating current mode, enough average current is allowed to flow between the converters to keep the instantaneous circulating current always greater than zero. Thus, both converters are kept in conduction at all times, even when the output current is zero. The main important advantage is that the power circuit has a natural freedom for current to flow in either direction at any time. A load current reversal is an inherently smooth process, not dependent on control systems to detect the load current and block or unblock the power circuits. The load current does not have glitches, and the control system does not have dead-bands, as is the usual case with the circulating current free operation. The price that one pays for the advantages of the circulating current mode is inter-converter chokes and an extra regulator to keep the circulating current under control.

The circulating current mode of operation was chosen for the magnet trim power supply because it was felt it would provide substantially better performance for transitions through zero output current.

<u>Power Transformer Connections:</u> The three-phase AC power connections in Figure 3 to a dual converter usually takes one of two forms:

1. The AC inputs to the positive and negative converters may be fed from separate secondary windings of a common transformer.
2. The AC inputs may come from a common transformer secondary

winding.

Operation of the dual converter for these two approaches is similar in many ways and yet there are important differences. For the two approaches, the SCR timing signals, the individual converter output voltage waveforms, and the total output voltage waveforms are identical. An important difference however, is that with a common secondary winding, inter-converter chokes are necessary in both the upper and lower inter-converter connections. The reason is that in the normal SCR firing sequence, SCR's in the positive and negative converters are on at the same time which would otherwise short out the transformer's secondary and result in excessive transformer currents. With independent secondary windings the transformer shorting problem does not exist since there are no common AC line connections between the converters. As a result, with independent secondaries, inter-converter chokes are only required in either the upper or lower inter-converter connections.

Another more subtle difference that occurs between the single and independent secondary cases is the magnitude of the ripple current which exists between the converters. Waveforms for operation of a dual converter with independent secondary windings is shown in Appendix A, Figures A1 - A6 with various firing angles (α_p = 15^o, 30^o, 45^o, 75^o, and 90^o) and resultant output voltages. Similar waveforms for operation with a common secondary are shown in Figures A7 - A12. In all waveforms, phase overlap has been neglected. The converter voltage waveforms V_{12}, V_{34}, and $(V_{12} + V_{34})/2$ are the same in each case as already mentioned. The circulating ripple current which flows is directly related to the volt-time integral which is applied to the inter-converter chokes as shown in equation 1.

$$i_{ripple} = \frac{1}{L}\int v_L \cdot dt = \frac{1}{L}\text{ (volt-time integral)} \qquad (1)$$

A comparison of the peak to peak circulating current ripple for the two types of transformer windings can be obtained simply by evaluating the maximum positive volt time integral that is applied to the choke. To make the comparison easier, assume that equivalent chokes are in the upper and lower inter-converter connections for both cases. Appendix Figures A1 - A6 show the total voltage applied to the inter-converter chokes with independent secondary windings. The voltage across the upper or lower chokes is 1/2 the total voltage. Appendix Figures A7 - A12 show the voltage across the upper and lower chokes when a common secondary winding is used. Notice that with a common secondary the upper and lower choke voltages are out of phase at all firing angles except $\alpha_p = 90^\circ$. If the upper and lower choke voltages were added together the sum would be the same as the total choke voltage obtained with the independent secondary connections. To obtain the volt-time integral for each case at different firing angles a graphical integration has been done for the upper choke voltage in Figures A1 - A12. The results of these calculations are shown in Table I where the constant K = $(x \frac{\text{msec}}{\text{sq}}$ $(y \frac{\text{volts}}{\text{sq}})$. The important fact to notice is that over the firing range the maximum volt time integral for a common secondary (14K) is 3 1/2 times larger than the maximum volt time integral with independent secondaries (4K). It follows that, if equivalent designs were made so that the circulating current

FIRING ANGLE		UPPER CHOKE VOLT-TIME INTEGRAL	
α_p	α_n	Common Secondary	Independent Secondaries
15	165	1K	.5K
30	150	4K	2K
45	135	7.5K	.75K
60	120	14K	3.5K
75	105	7.5K	.9K
90	90	4K	4K

TABLE I

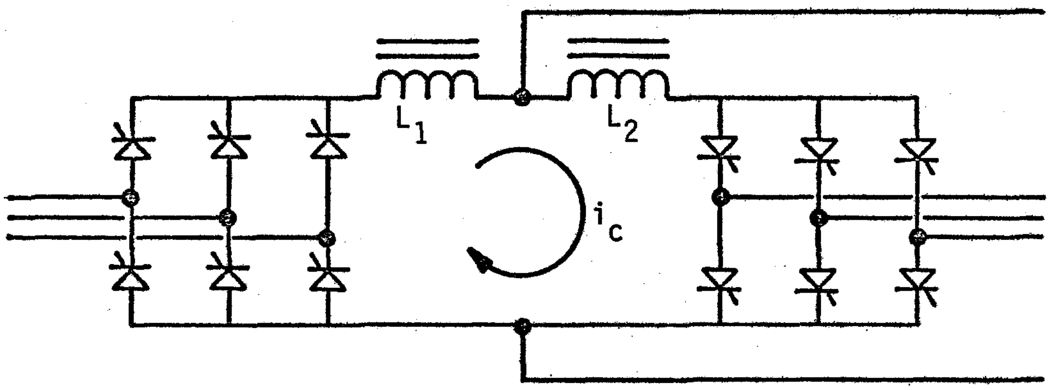
Choke Volt-Time Integral Versus Firing Angle

ripple was the same, the total inter-converter inductance with a common secondary connection would have to be 3 1/2 times larger than if independent secondaries were used.

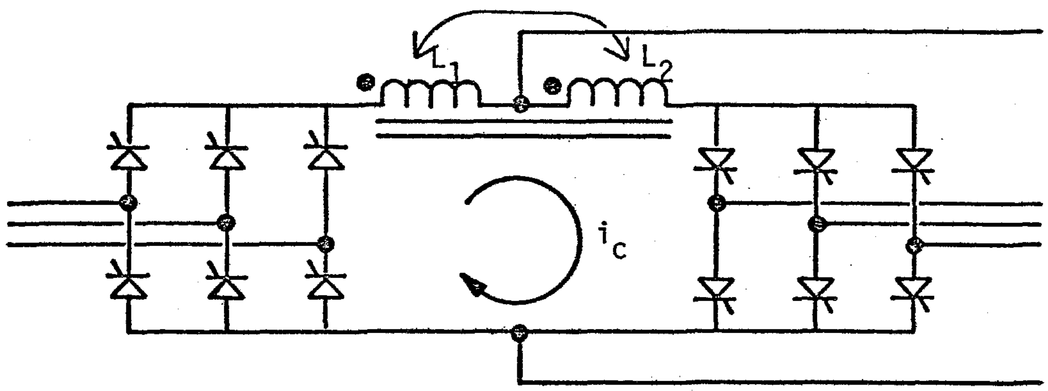
The independent secondary design uses smaller inter-converter chokes than the common secondary design, however, the transformer with two secondaries is larger and more costly. To make a better comparison of the approaches, three phase transformers and inter-converter chokes for a 4KW dual converter were designed for each approach. For each case the transformers and chokes was designed for an approximately 70° C temperature rise and equivalent window area fill. The results were that the material cost (cores and copper wire) and the weight of the common secondary design was about 40% larger than the independent secondary design. The actual cost comparison may vary somewhat since the labor for a two secondary transformer will be higher than for single secondary but then the labor for two chokes must be higher than for one and these costs tend to offset each other. The one fact that should not change, however, is that the weight of the two secondary design is considerably lighter than the alternate design. Other factors such as uses of standard transformers and chokes may influence the design for small quantity designs but for large quantities, the independent secondary design appears to be the best choice at the power level considered. Thus, because of the weight and cost advantages, the independent secondary design approach was chosen for the four quadrant magnet trim power supply. Since chokes are not required in both the upper and lower connections, an equivalent choke in only the upper inter-converter connection is used as shown in Figure 4.

Inter-Converter Choke: Several choices exist for the design and connection of an inter-converter choke as shown in Figure 4. Separate chokes may be used as in Figure 4a or a single choke with dual windings (dual choke) tightly coupled, and mmf's aiding or opposing as shown in Figures 4b and 4c. Each circuit performs differently and should be understood to make a reasonable design choice.

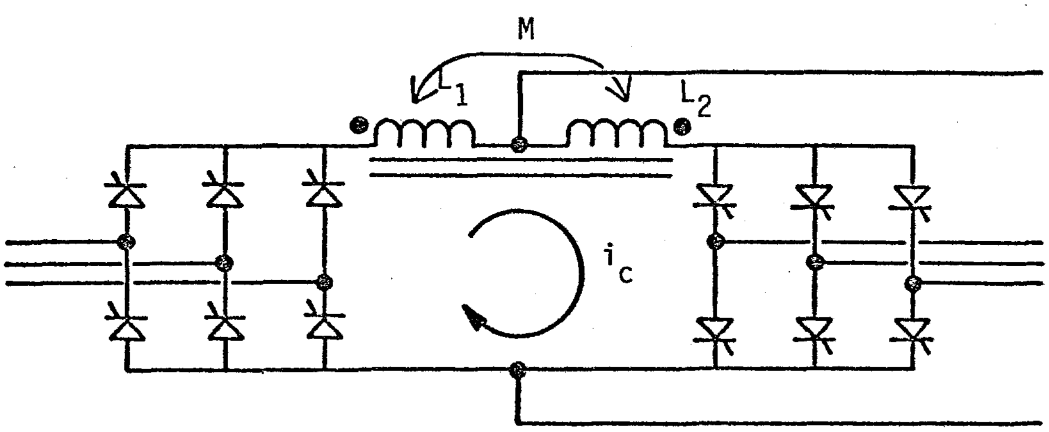
Operation with separate chokes between the converters is analyzed in Appendix B. Chokes L_1 and L_2 must each be sized (for ampere-turns) to handle the full load current plus circulating current. The output inductance of the dual converter is shown to be L/2. This inductance can be used as part of a following filter to reduce output voltage ripple. The appendix also shows that a ripple current in the dual



a). Separate Chokes



b). Coupled Choke (with mmf aiding, coupling coefficient $K_m = 1$)



c). Coupled Choke (with mmf opposing, coupling coefficient $K_m = 1$)

Figure 4 - Dual Converter with Different Inter-Converter Choke Connections (In all Cases $L_1 = L_2 = L$).

converter output tends to induce a circulating current. A circulating current regulator is necessary to hold the average circulating current constant. The other portion of the circulating ripple current which is important is that due to the positive and negative converter instantaneous voltage differences. Obviously the total impedance to interconverter voltage differences is proportional to 2L.

Operation of a dual converter with a dual choke (aiding mmf's) is somewhat different. (See Appendix C). Under steady state operation the total impedance between the converters is proportional to 4L ($L_T = 2L + 2M = 4L$), due to the aiding coupling between the chokes. Thus, for the same total inter-converter impedance, the circuit in Figure 4b would use smaller inductances than that in 4a. Furthermore, with separate chokes, each choke must be sized to handle $i_L + i_C$ or for both chokes $2(i_L + i_C)$. With a dual choke, load current can only flow in one side at a time. Therefore the dual choke is only sized for $2i_C + i_L$. One compensating factor is that the separate choke design has a "built-in" series output filter choke of L/2, rated for i_L, which the dual choke does not have. This may or may not be of some advantage depending on the particular design. Normally the dual choke design still has the advantage.

While dual chokes are not normally commercially available, they are not hard to design and build once the requirements are established. Thus, even though the dual choke may require a special design, the extra effort should result in a lighter, smaller, and less costly dual converter package. It should be noted that operation of the dual converter with dual chokes is similar to that with separate chokes in one way. With load ripple current, both induce a circulating current between the converters. With a properly designed circulating current regulator, however, there should be no problem.

One may try to connect the dual choke with opposing mmf's as shown in Figure 4c to cancel the circulating current mmf components and reduce the dual choke size. This should not be done, however, since the total inter-converter inductance becomes zero ($L_T = 2L - 2M = 0$) due to the coupling between the chokes. The resultant high circulating ripple current results in a very unsatisfactory design. The connection in Figure 4c therefore is considered no further.

The magnet trim power supply uses the inter-converter choke connection shown in Figure 4b because of its cost and weight advantage. The choke was designed and especially built for this application. If commercially available chokes are to be used, the design in Figure 4a would be most likely choice.

Output Filter: A dual converter with or without circulating current, inherently has high output voltage ripple. The converter output voltage with circulating current is shown for several different operating points in Figures A1 - A6. At several points, $\alpha p = 45^\circ$ and $\alpha p = 75^\circ$, the output voltage ripple is very large. The voltage ripple causes unwanted current ripple in the magnet load. To reduce the ripple to an acceptable level an output filter is added to the dual converter as shown in Figure 5. The filter is designed to have an 80 Hz

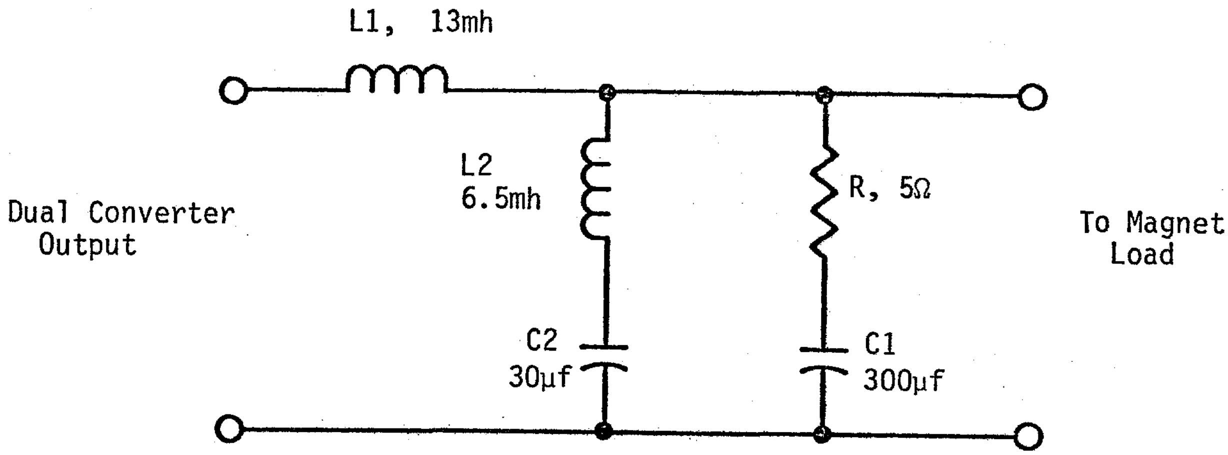


Figure 5 - Dual Converter Output Filter

bandwidth and provide large attenuation of the 360 Hz output ripple. Since a coupled inter-converter choke is used, all of the inductance represented by L1 is added in series with the D/C output. Inductor L2 and capacitor C2 form a tuned circuit at 360 Hz. The combination of L1, L2, and C2 provides a minimum output ripple attenuation of -35db at 360 Hz. Components L1, C1, and R form a well-damped low pass filter with a cut off frequency of 80 Hz. The purpose of the LCR filters is to attenuate the harmonic components above 360 Hz in the dual converter output. The two part filter described does not provide filtering for ripple frequencies below about 100 Hz. Attenuation of lower frequency ripple is discussed in a later section.

III. Dual Converter Operation

Several aspects of dual converter operation will be discussed: the derivation of the SCR gate timing signals, the measurement of circulating current, operation of the circulating current regulator, and operation of the load current regulator.

SCR Gate Timing Signals - Successful operation of the dual converter requires that $\alpha p + \alpha n = 180^0$ for all output voltages as already mentioned. To generate the positive and negative converter SCR timing signals, different but very similar firing circuits are required by the positive and negative converters. Figure 6a shows a simplified diagram of the firing circuits and power SCR's in the D/C. Figure 6b shows in more detail how the SCR timing signals for two SCR's in the positive converter and two SCR's in the negative converter are derived from one phase of the 3Ø reference and an analog control signal. The 3Ø reference input comes from three low voltage transformer windings connected so that they are in phase with the secondary windings of the 3Ø power transformer. In Figure 6b the AC input reference, V_{ab}, is integrated, and inverted and integrated to provide $\int V_{ab}$ and $\int V_{ba}$. The main purpose of the integrators is to remove noise from the reference signals. For the positive converters, $\int V_{ab}$ and $\int V_{ba}$ feed the positive inputs of a pair of comparators and the firing angles control signal is fed to the negative comparator inputs.[1] The comparator outputs drive one-shots which set the width of the SCR gate pulse at about 100^c. Each one-shot drives a high frequency gated oscillator which drives an SCR gate hard-fire circuit in the positive converter. For the negative converter, $\int V_{ab}$ and $\int V_{ba}$ feed the negative inputs of another pair of converters and the negative converter firing angle control signal is fed to the positive comparator inputs. The comparators drive one-shots and gated oscillators as already described. In Figure 6b the timing signals for SCR's 3, 6, 3', and 6' are generated. Timing signals for the

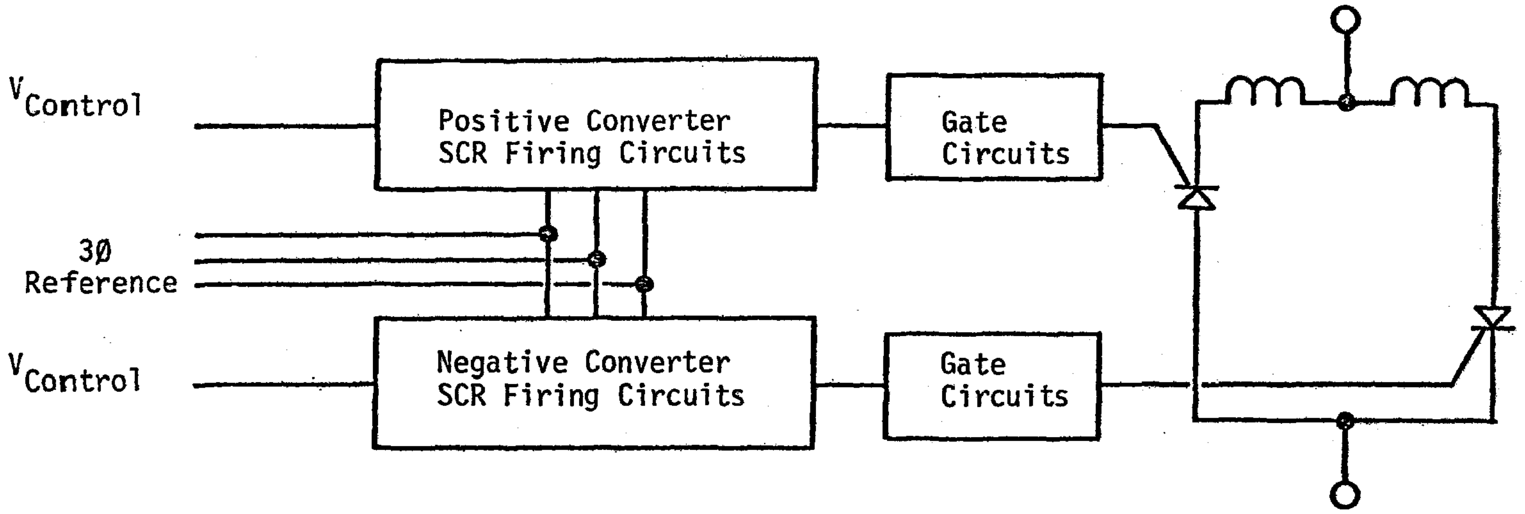


Figure 6a - Simplified Dual Converter SCR Control

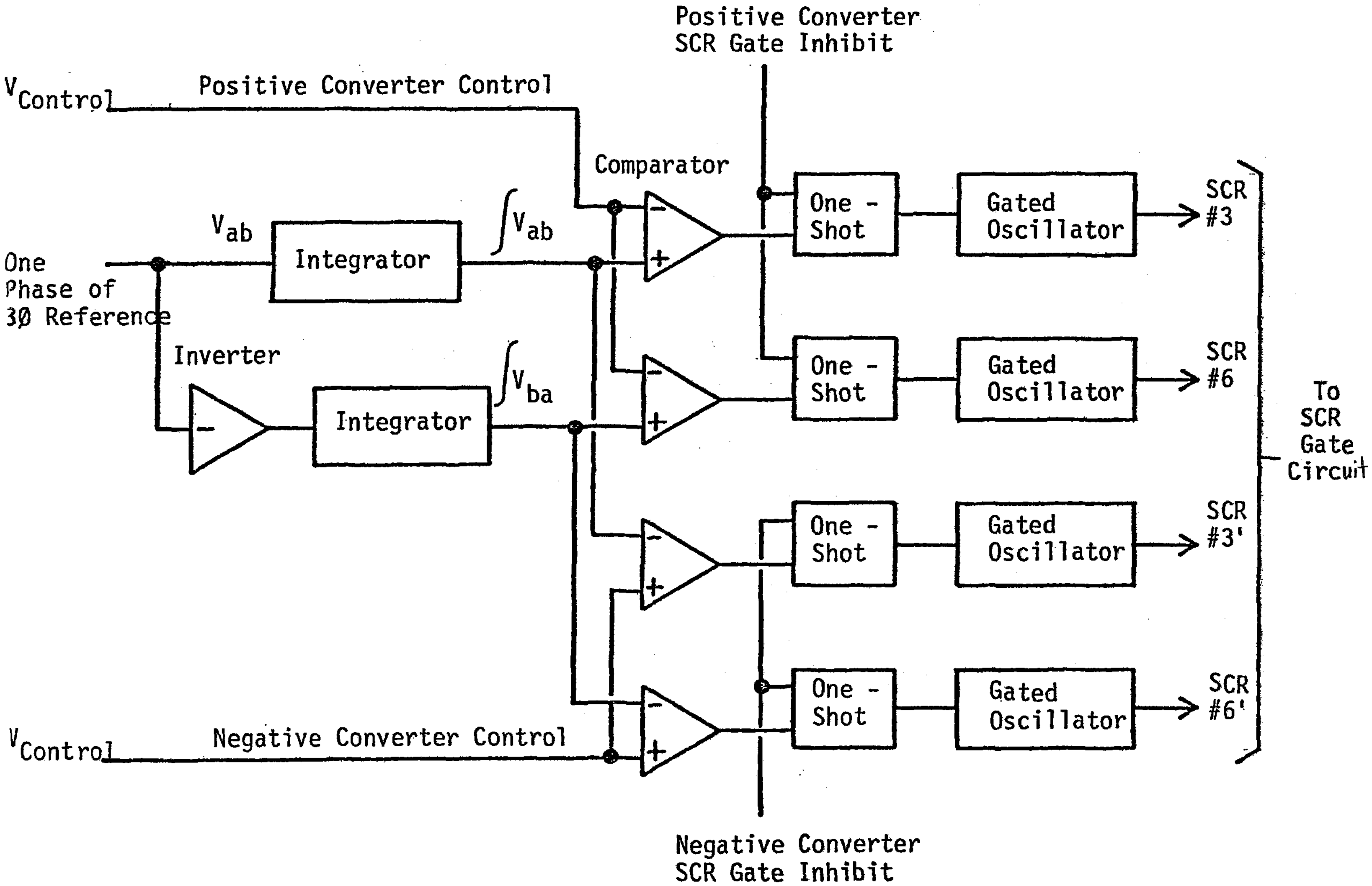


Figure 6b - Block Diagram of Typical SCR Firing Circuits remaining SCR's are derived in a similar way from V_{bc} and V_{ca}. Figure 7 identifies the position of each SCR and shows by means of the comparator outputs how all of the timing signals are generated. The time interval denoted by a circled number on the waveforms indicates the time an SCR gate signal is present for that particular SCR. With ideal components and balanced circuits, $\alpha p + \alpha n = 180^\circ$ whenever V_{pos} (control) = V_{neg} (control) as seen in Figure 7. In a practical application, V_{pos} is not exactly equal to V_{neg}, since some difference between the signals is needed to control circulating current. Thus the V_{pos} and V_{neg} inputs are separate.

One disadvantage of this firing circuit arrangement is that the integrated reference signals must be well balanced to minimize subharmonic ripple. Balance is accomplished by trimming the gain of each integrator circuit. Subharmonic ripple control will be discussed more later.

<u>Circulating Current Measurement</u> - The presense of circulating current (Icc) is important since it provides a freedom for load current to flow in either direction at any time which thus allows load current to reverse smoothly. Two important factors influence circulating current: load current ripple and inter-converter ripple. If due to the above factors, Icc becomes equal to at zero some instant, the output voltage waveforms given in Appendix A are no longer valid and smooth current reversal is unlikely. Therefore, I_{cc} is regulated so that zero values of I_{cc} do not occur particularly at low load current values.

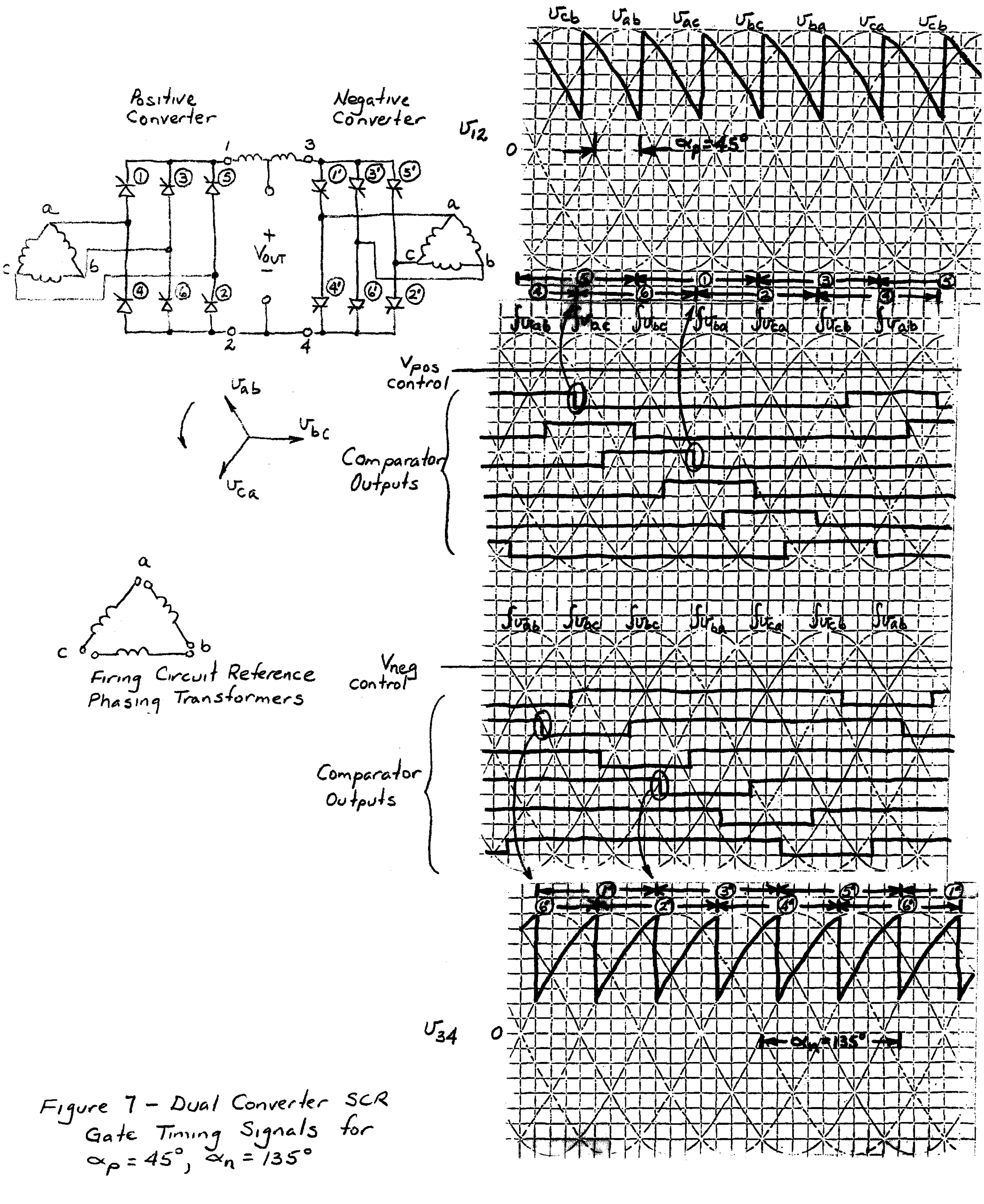


Figure 7 – Dual Converter SCR Gate Timing Signals for $\alpha_p = 45°$, $\alpha_n = 135°$

In order to regulate I_{cc}, a scheme such as that in Figure 8 is needed to first measure I_{cc}. Current sensor A measures the sum of the

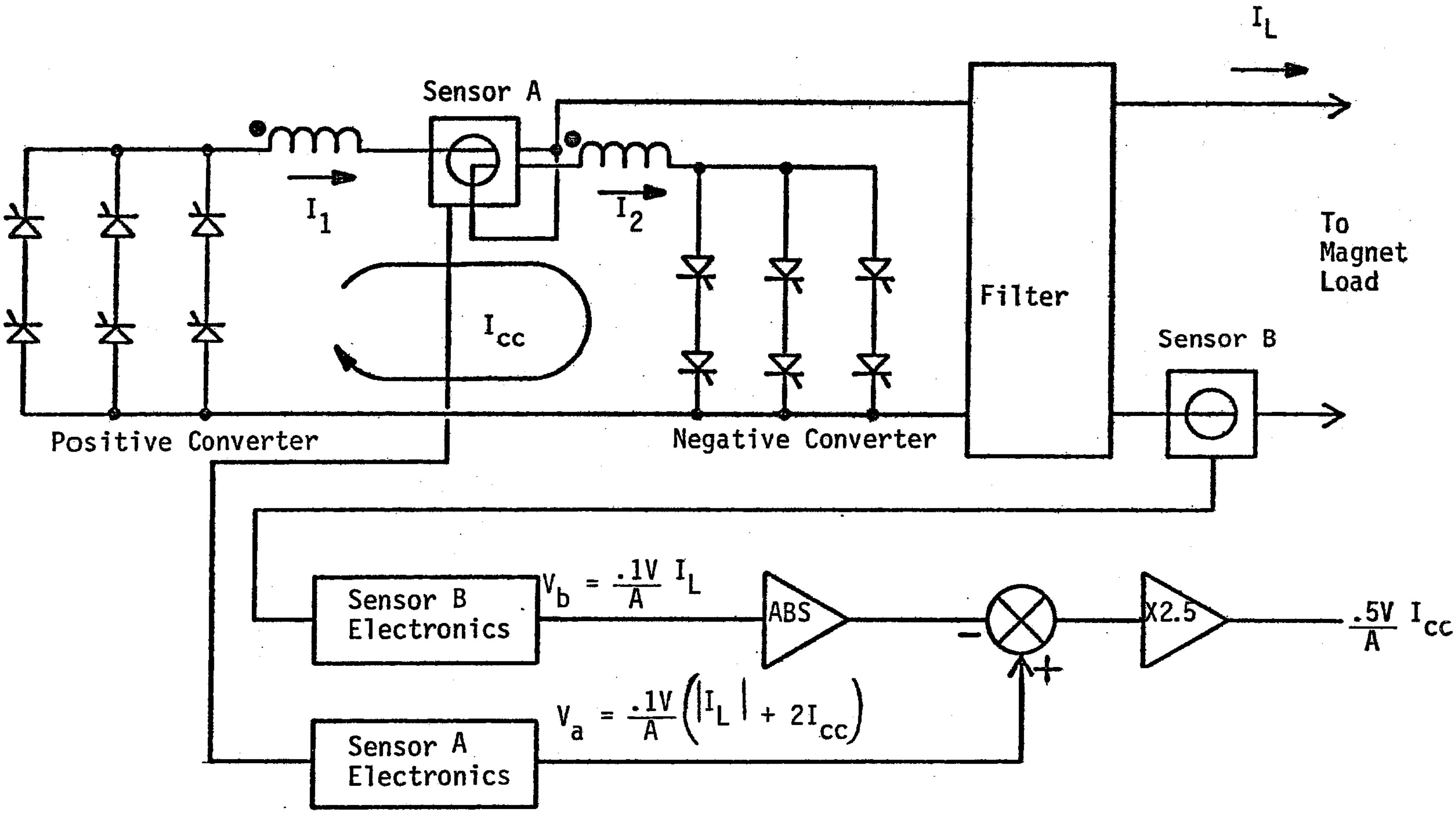


Figure 8 - Circulating Current Measurement Simplified Block Diagram

current I_1 in the positive converter, and I_2 in the negative converter. Bipolar current sensor B measures the load current, I_2, which can be positive or negative. When I_L is positive, $I_1 = I_c$ and $I_2 = I_c + I_L$. In either case, $V_a = \frac{.1V}{A}(I_L + 2I_c)$ and $V_b = \frac{.1V}{A} I_L$. The signal from the load current sensor is fed to an absolute value circuit to obtain $|V_b|$ and then the circulating current is obtained by taking $V_a - |V_b|$. The result is multiplied by the scale factor 2.5 to give a voltage signal equal to $\frac{.5V}{A} I_{cc}$. In deriving I_{CC} the higher frequency shunt effects of the output filter can be neglected in most cases.

Current sensors A and B should be isolated from the bus. Hall-effects current sensors and DCCT's are well suited to this purpose. Alternately, shunts and optical isolation amplifiers could be used to measure I_1, I_2, and I_L. For the present application, Halltron (Hall-effect) current sensors were used to provide isolated signals with relatively low output ripple.

An alternate arrangement for measuring circulating current is shown in Figure 9. Two diodes are inserted to shunt load current around current sensor A, leaving only circulating current to pass through the sensor. One diode conducts for positive load current and the other diode for negative load current. The arrangement permits a direct measurement of I_{CC} at the expense of adding load carrying diodes and additional dual converter loss. This arrangement is attractive for low current outputs. It should be noted that no cross-over problems have been observed for current passing through zero due to change of one conducting diode to the other.

Circulating Current Regulator - Separate regulator loops are used for the circulating current and load current although some components are common to both regulators. Figure 10a shows the circulating current regulator. In the present design the average value of I_{CC} is set at about 2-3 amps or 10% of the maximum output current. The I_{CC} monitor provides a feedback signal equal to $\frac{.5V}{A}$ I_{CC} which is compared to a variable circulating current reference voltage, V_{CC}. The

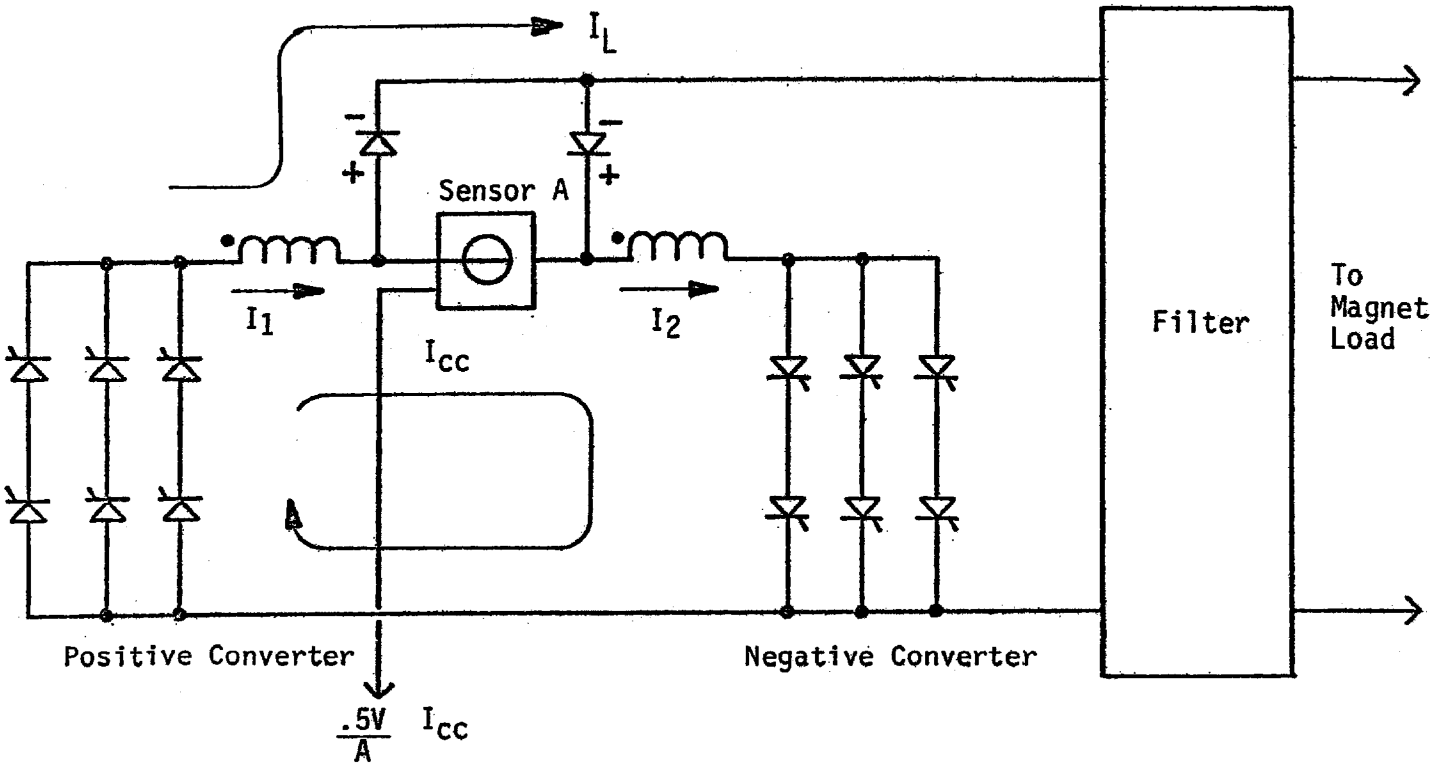


Figure 9a - Alternate Circulating Current Measurement-Positive Load Current

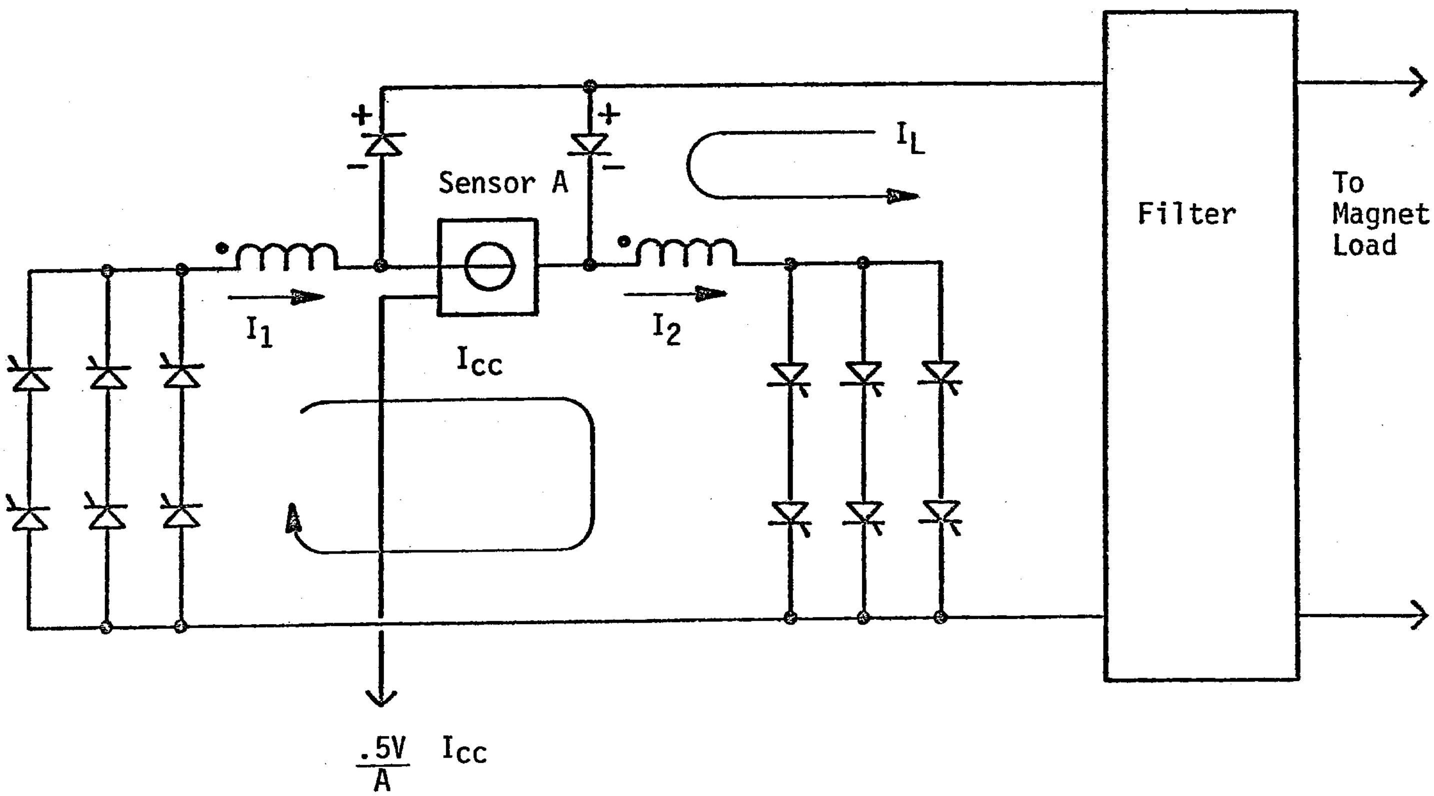


Figure 9B - Alternate Circulating Current Measurement-Negative Load Current difference signal is amplified by the circulating current amplifier, A1 to give the error signal V_e. Then V_e is inverted to form $\overline{V}_e$. The noninverted error signal is summed with the load current regulator signal and fed to the positive converter SCR firing circuits. The inverted error signal is summed with the load current regulator signal and fed to the negative converter SCR firing circuits. As a result, the converter control signals are made slightly different in a symmetric way and $\alpha p + \alpha n \neq 180^{\circ}$. (The error signal could be added to only the positive or negative control signals but it results in a non -symmetric output voltage range). The effect of V_e is to increase slightly the output voltage of the positive converter and to decrease slightly the output voltage of the negative converter, resulting in an average difference in the voltage, V_c, between the converters. Whenever, the average positive converter output voltage is larger than that of the negative converter an average current, I_{cc}, equal to V_c(ave) divided by the choke resistance flows between the converters. The purpose of the circulating current regulator circuit is to control this current.

Figure 10b shows a simplified diagram of the I_{cc} regulator. The regulator transfer function is given by equation 2.

$$\frac{I_{cc}}{V_{cc}} = \frac{G(s)}{1+G(s)H(s)} \qquad (2)$$

Where G(s) = (A1) (A2) (A3) and H(s) = A4. A Bode diagram is shown in Figure 11 which shows the closed loop response and stability of the I_{cc} regulator. Amplifier A1 has a gain of .6 and a rolloff at 35 Hz to reduce the effect of ripple in the feedback circuit. The transfer characteristic for Vc/Ve which is represented by A2, was measured and

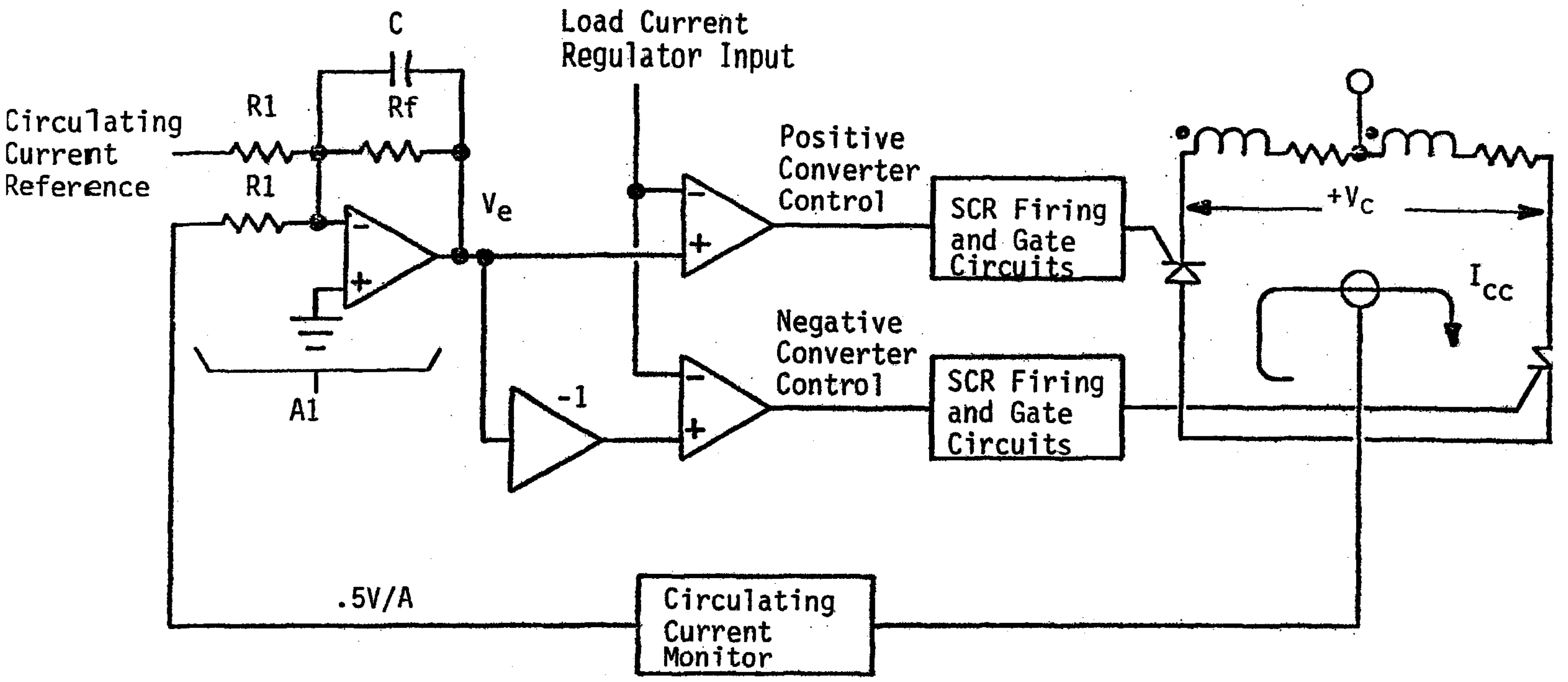


Figure 10a - Circulating Current Regulator Block Diagram

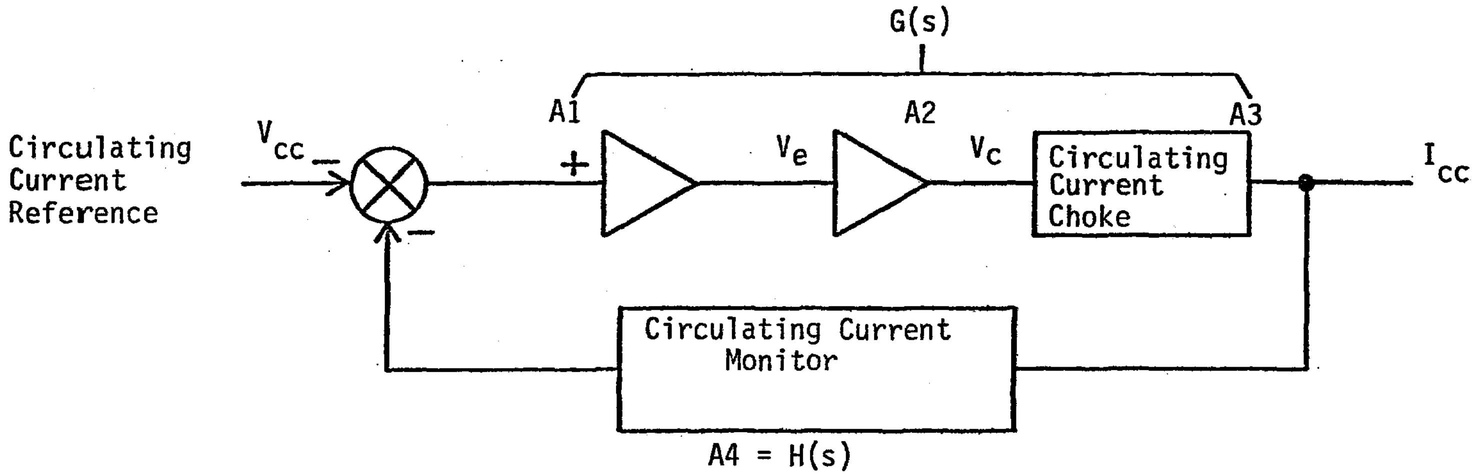


Figure 10b - Simplified Circulating Current Regulator

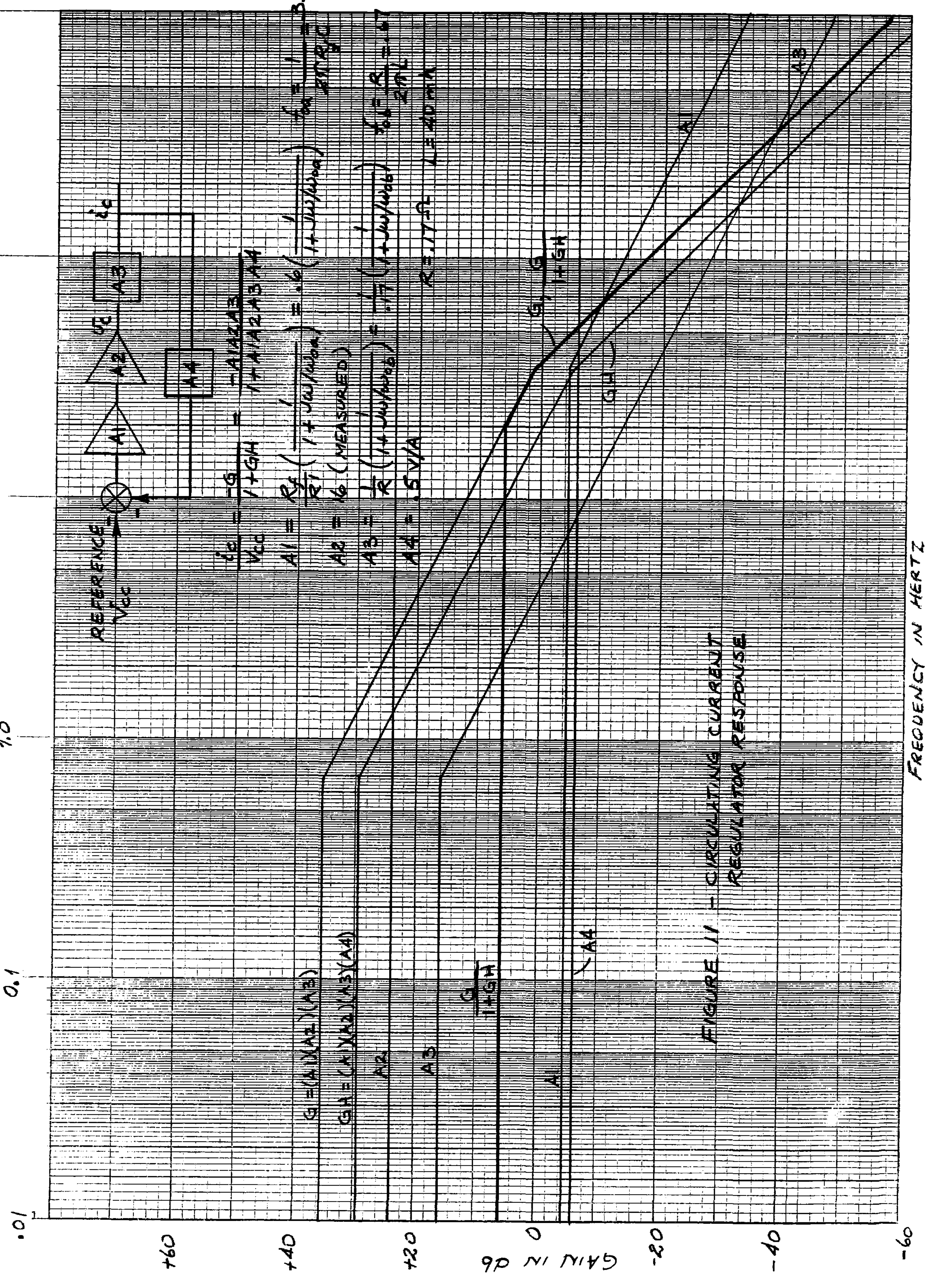


FIGURE 11 - CIRCULATING CURRENT REGULATOR RESPONSE

found to be about sixteen for the frequencies of interest. The transfer function for I_{cc}/V_c is simply determined from the total inductance of the inter-converter choke and the resistance of the choke and circuit wiring. Figure 11 shows that the unity gain cross-over point for G(s) H(s) is at 20 Hz with a slope of -20db/dec, thus insuring regulator stability. Finally, the closed-loop response shows that the regulator DC gain is +6db, giving 2A of average circulating current for each volt of reference voltage.

Load Current Regulator - The design of the load current regulator differs slightly for different magnet loads. However the same general approach can be used in each case. Figure 12a shows a block diagram of the dual converter load current regulator. In many ways, the regulator is similar to the circulating current regulator. The load current reference is compared to the load current monitor signal and the resultant signal amplified by the input amplifier. The gain of the input amplifier for either signal is given by equation 3. At lower frequencies the amplifier acts like an integrator with a gain

$$A1 = \frac{(1 + R_1 C_1 S)}{SR_1C_2} \qquad (3)$$

determined by $1/R_1 C_2$. A load compensation factor, $(1 + R_1 C_1 S)$, becomes important at higher frequencies. The purpose of this term is to compensate for the natural roll-off characteristic of a magnet load. The amplifier output is used to drive both the positive and negative converter firing circuits, thus controlling the dual converter output voltage. Attached to the converter output is the filter with a 360 Hz trap and an 80 Hz roll-off as already described. The filter output is connected to the magnet load which is basicly a

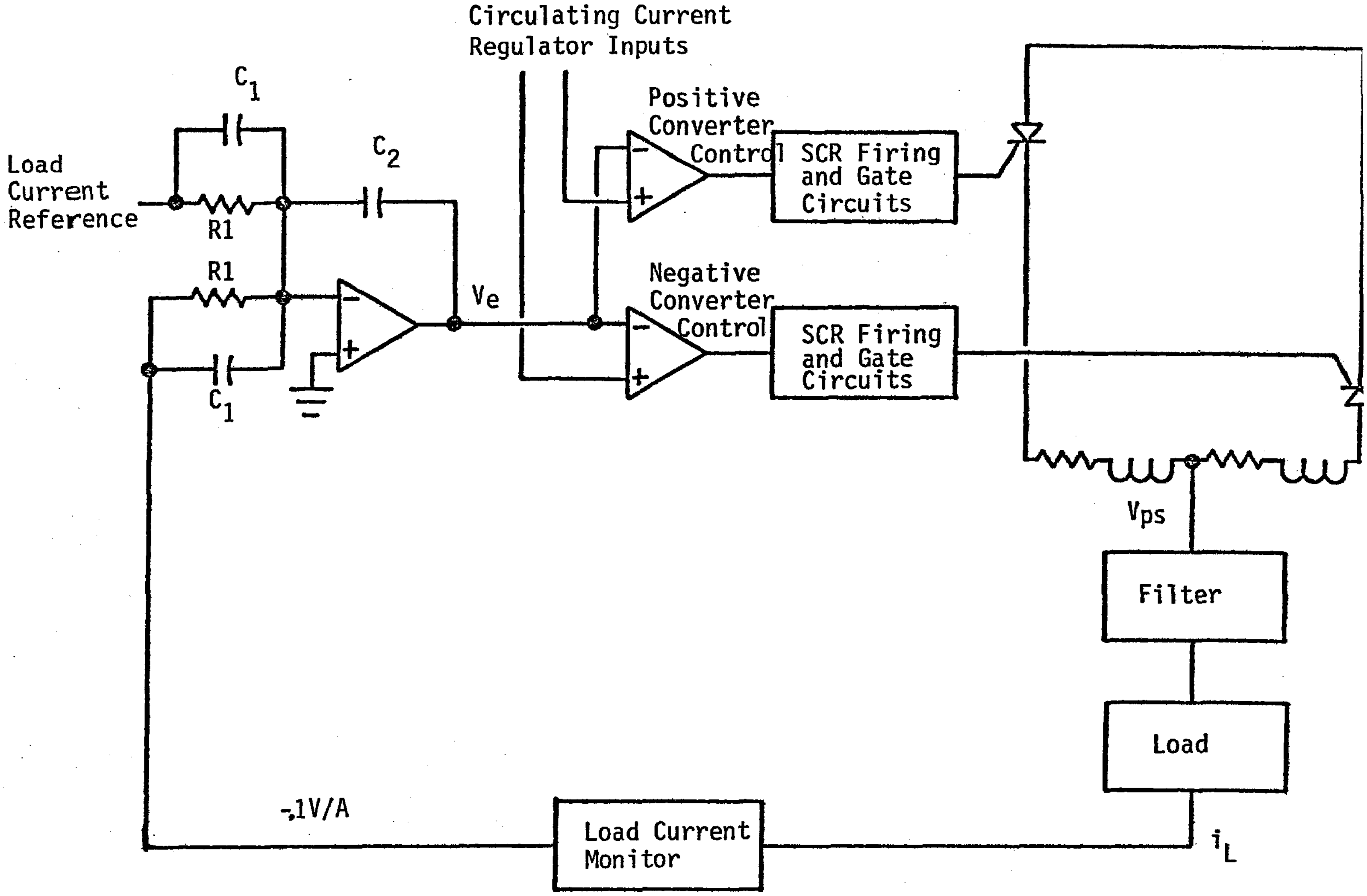


Figure 12a - Load Current Regulator Block Diagram

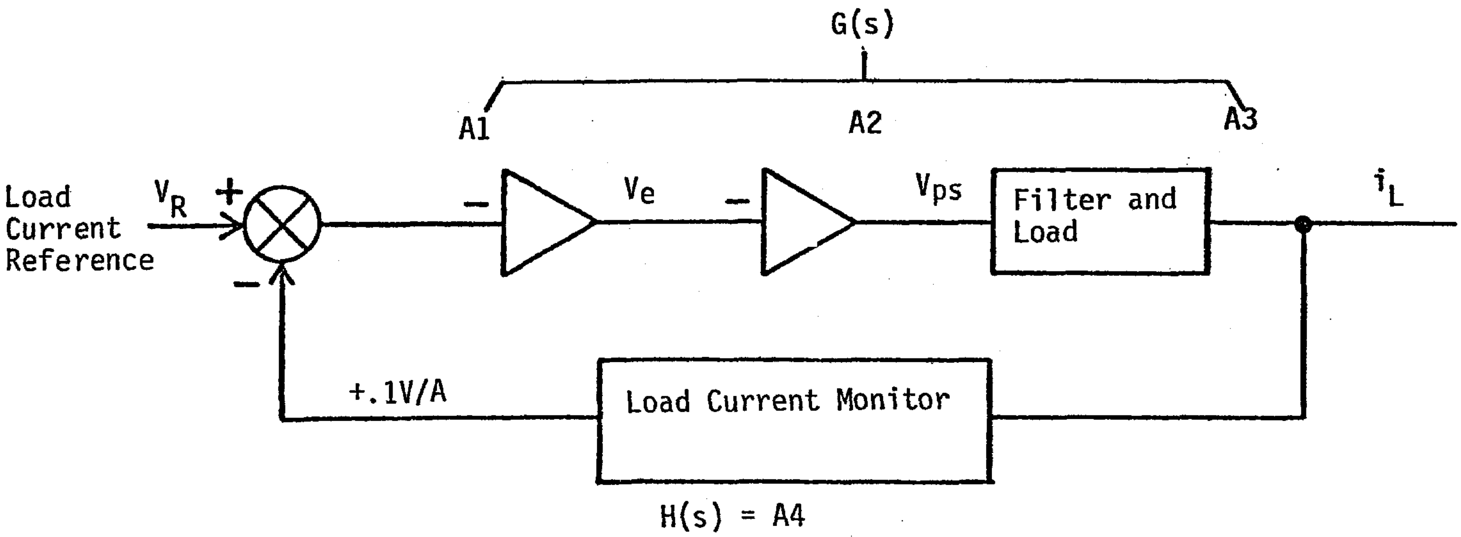


Figure 12b - Simplified Load Current Regulator series R, L load. Figure 12b is a simplified diagram of the load current regulator. The closed-loop transfer function for the load current regulator is given by equation 4 where G(s) = (A1) (A2) (A3) and H(s) = A4.

$$\frac{i_L}{V_R} = \frac{G(s)}{1+G(s)H(s)} \qquad (4)$$

A Bode diagram of the load current regulator varies for different loads. Since the actual superconducting load for the dual converter was not conveniently available for testing the converter, a differnt load was assembled in the lab. Figure 13 shows a Bode diagram of the load current regulator using the laboratory test load. Some assumptions were made in plotting Figure 13. First the output filter was assumed to have unity gain for the frequencies of interest and therefore the plot for A3 simply shows the natural roll-off of the series R-L load at 22 Hz. (Normally the load roll-off frequency is much lower). And secondly, the converter transfer characteristic, A2, is assumed to be constant at 25V/V. The current regulator is designed to have a bandwidth of 20 Hz. The input amplifier is designed with a leading compensation factor at 22 Hz to match the load characteristic and a gain such that GH=1 at 20 Hz. At 20 Hz the slope of GH is -20 db/dec and thus the load current regulator is stable. The closed-loop response, $\frac{G}{1 + GH}$, shows that the corner frequency is 20 Hz as set by the bandwidth requirement, and that the DC gain of the regulator is 10A/V. The assumptions which were made do not change these results.

Tests with the dual converter showed that the converter transient response to a small step function was quite good and that operation

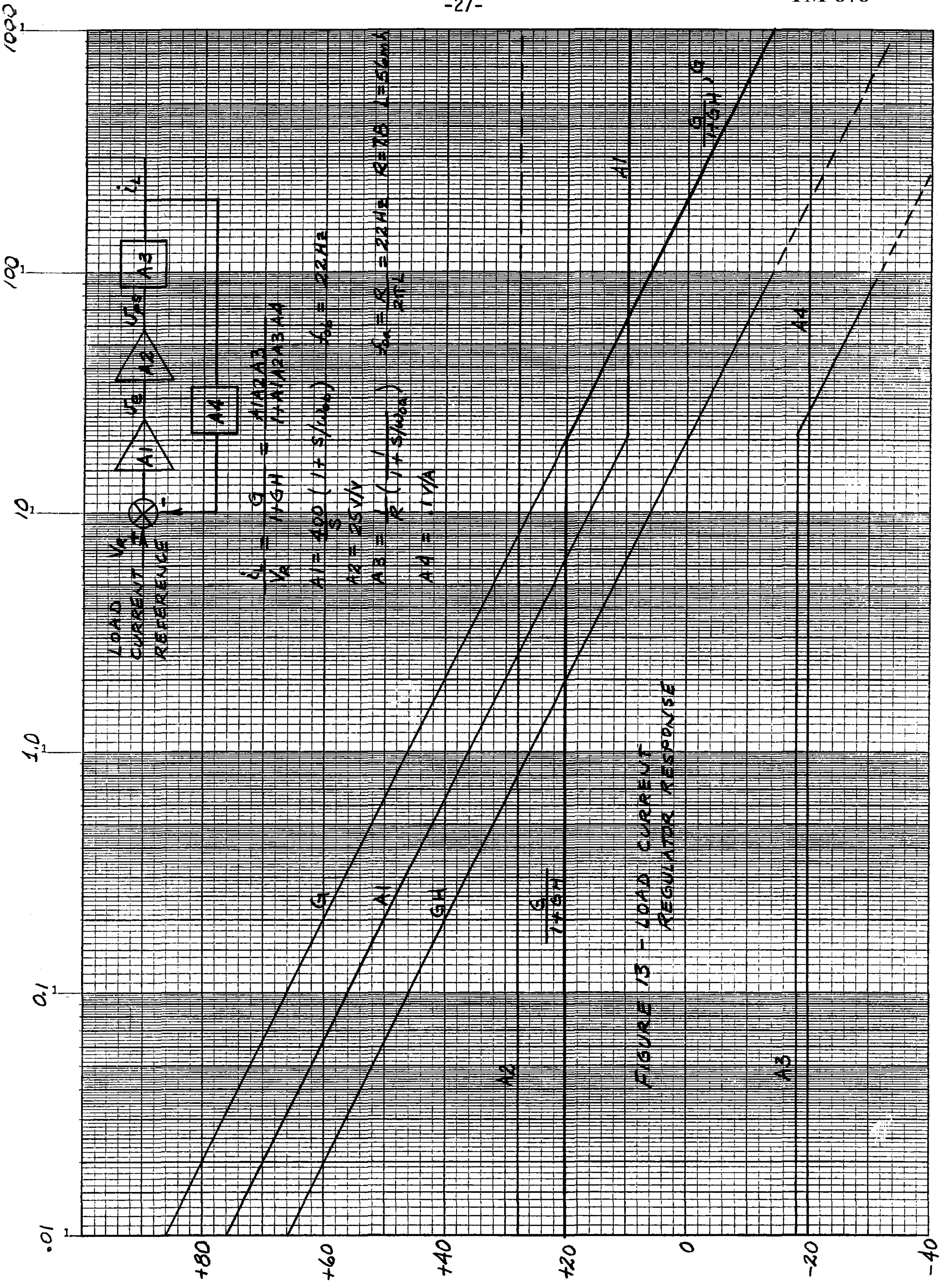


FIGURE 13 - LOAD CURRENT REGULATOR RESPONSE

with circulating current was easy. Looking closely at the load current ripple showed that the filter had adequately reduced the 360 Hz fundamental ripple. However several subharmonics had become predominant once the fundamental was attenuated. The load current contained 15 Hz ripple due to the Booster modulation of the 3Ø, 480V power line. Due to firing circuit imbalance, 60 Hz was present, and 120 Hz was present probably due to transformer imbalance. A Bode diagram can be drawn which shows that the regulator provides no ripple voltage attenuation above 20 Hz and very little at 15 Hz. To reduce the subharmonics to acceptable levels, three (3) parallel, tuned, voltage, feedback circuits were connected between Vps and Ve. The circuits were tuned to 15, 60, and 120 Hz, each with a Q = 30. (The tuned circuit feedback technique is described in much greater detail in Reference 2). The feedback circuits reduced the subharmonics to an acceptable level. Presence of the 60 and 120 Hz feedback circuits had little effort on the transient response of the dual converter. The 15 Hz circuit caused some ringing in the output for small step inputs. Perhaps a better solution for the 15 Hz ripple would be to widen the load current regulator bandwidth somewhat while at the same time moving the output filter's 80 Hz roll-off to a higher frequency to avoid stability problems.

IV SUMMARY

A four-quadrant power supply has been described which can be attached to one or more conventional or superconducting magnets in a magnet string and raise or lower the current in the magnets to which the supply is connected. General design considerations have been discussed and typical application to a +200V, +20A power supply described. Details of this design can be found in the following Fermilab drawings which are not included in this report. The drawings are meant to show a working but not necessarily optimum power supply design.

1. Dual Converter General Schematic, FNAL 2230.0-ED-46625.
2. Dual Converter Firing Circuit Schematic, FNAL 2230.0-ED-46626.
3. Dual Converter Current Monitor Electronics, FNAL 2230.0-ED-46627.
4. Dual Converter Circulating Current Regulator and Firing Circuit Drive Amplifiers, FNAL 2230.0-ED-46628.
5. Dual Converter Load Current Regulator and Subharmonic Ripple Filters, FNAL 2230.0-ED-46629.

Discussion of such things as isolation and power supply protection have been ommitted. Various approaches would be taken for different applications.

V REFERENCES

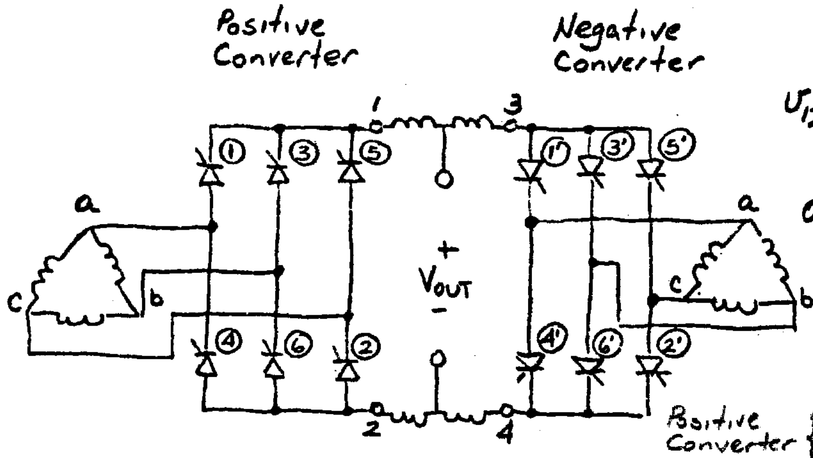
1. M. Berndt and C. Guracar, Private communication.

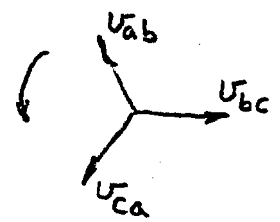
2. R. J. Yarema, "Subharmonic Ripple Reduction In SCR-Type Power Supplies", IEEE Particle Accelerator Conference, March 1979.

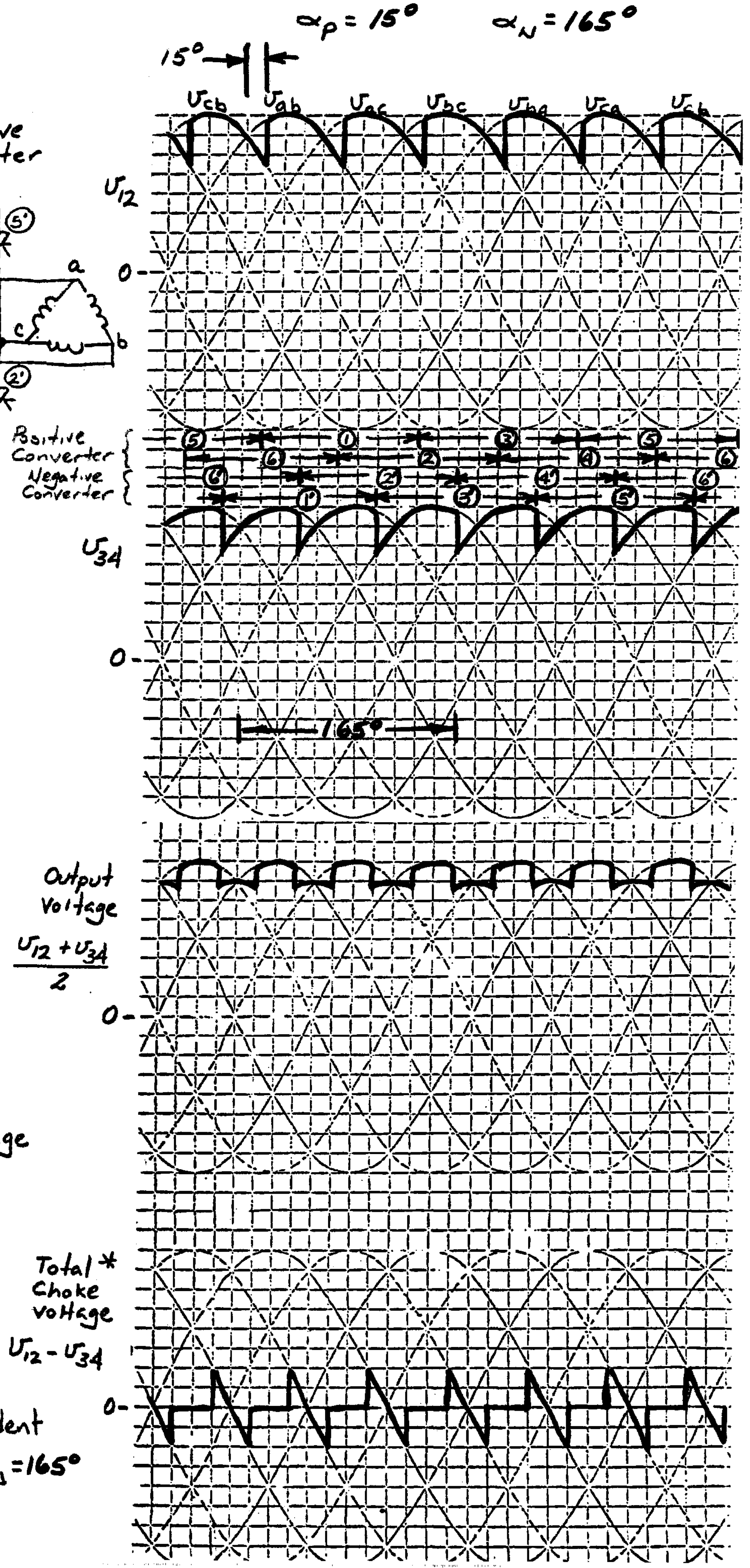
3. B. R. Pelly, Thyristor Phase-Controlled Converters and Cycloconverters, New York, Willy-Interscience, 1971, pp. 111-144.

Six Pulse Dual Bridge Converter

APPENDIX A



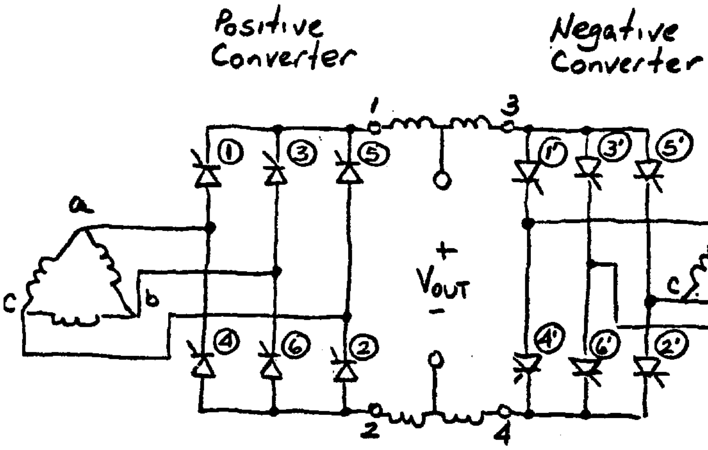


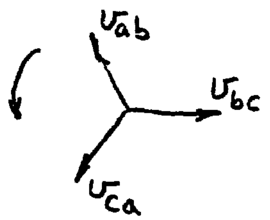


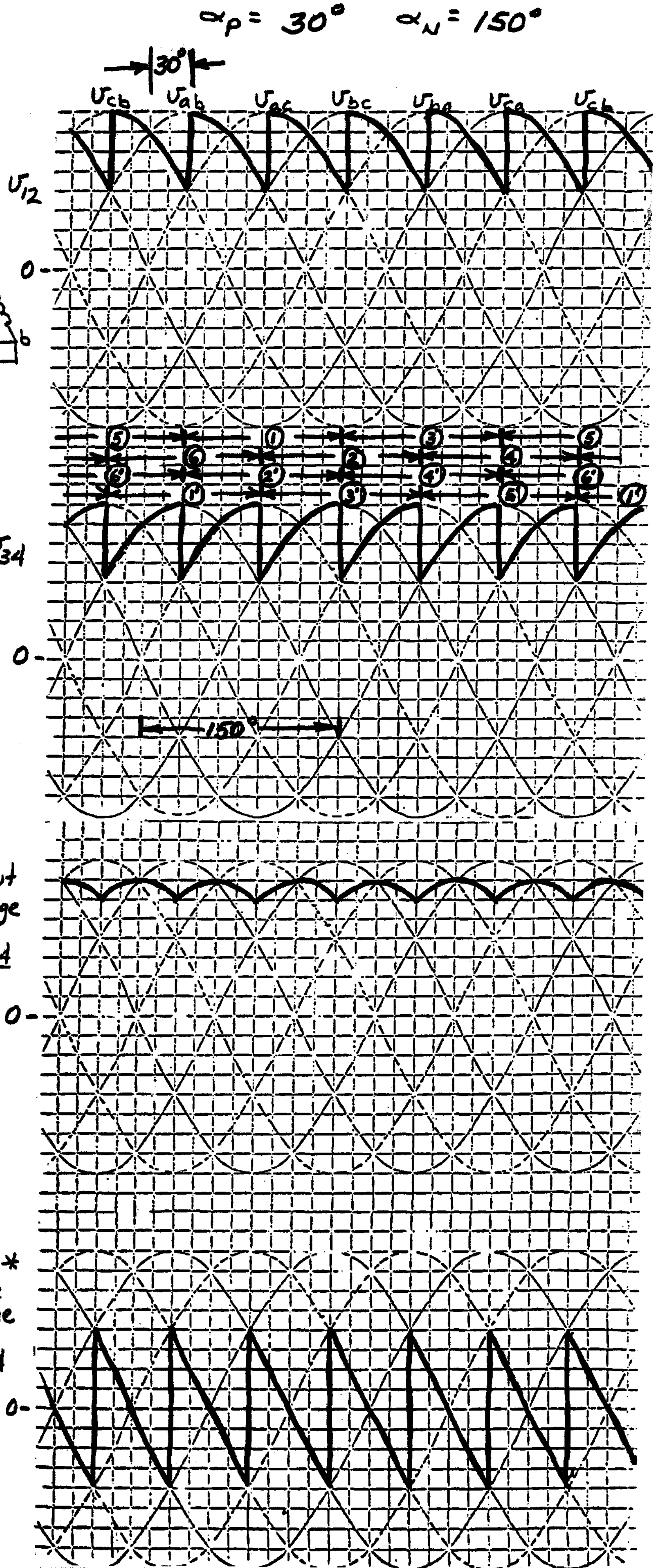
* Note: $v_{13} = v_{42} = \frac{1}{2}$ Total choke voltage

Figure A1 - Dual Converter, Independent Secondaries $\alpha_p = 15°$ $\alpha_N = 165°$

Six Pulse Dual Bridge Converter



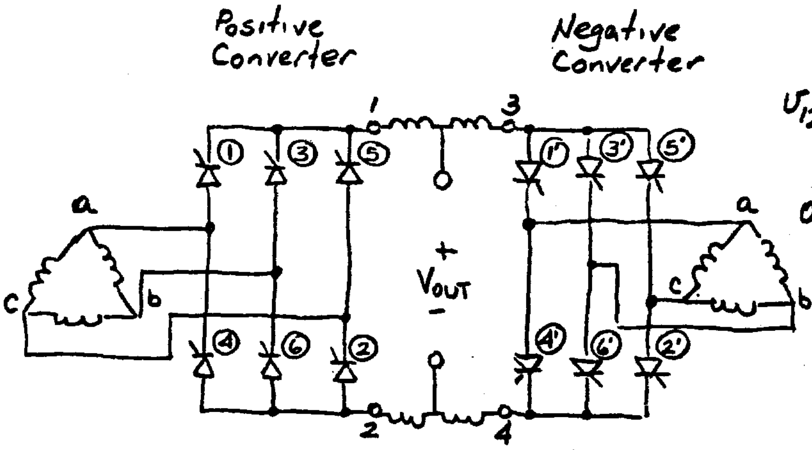


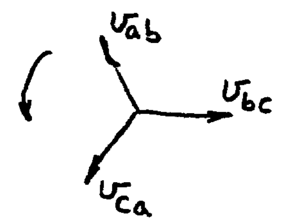


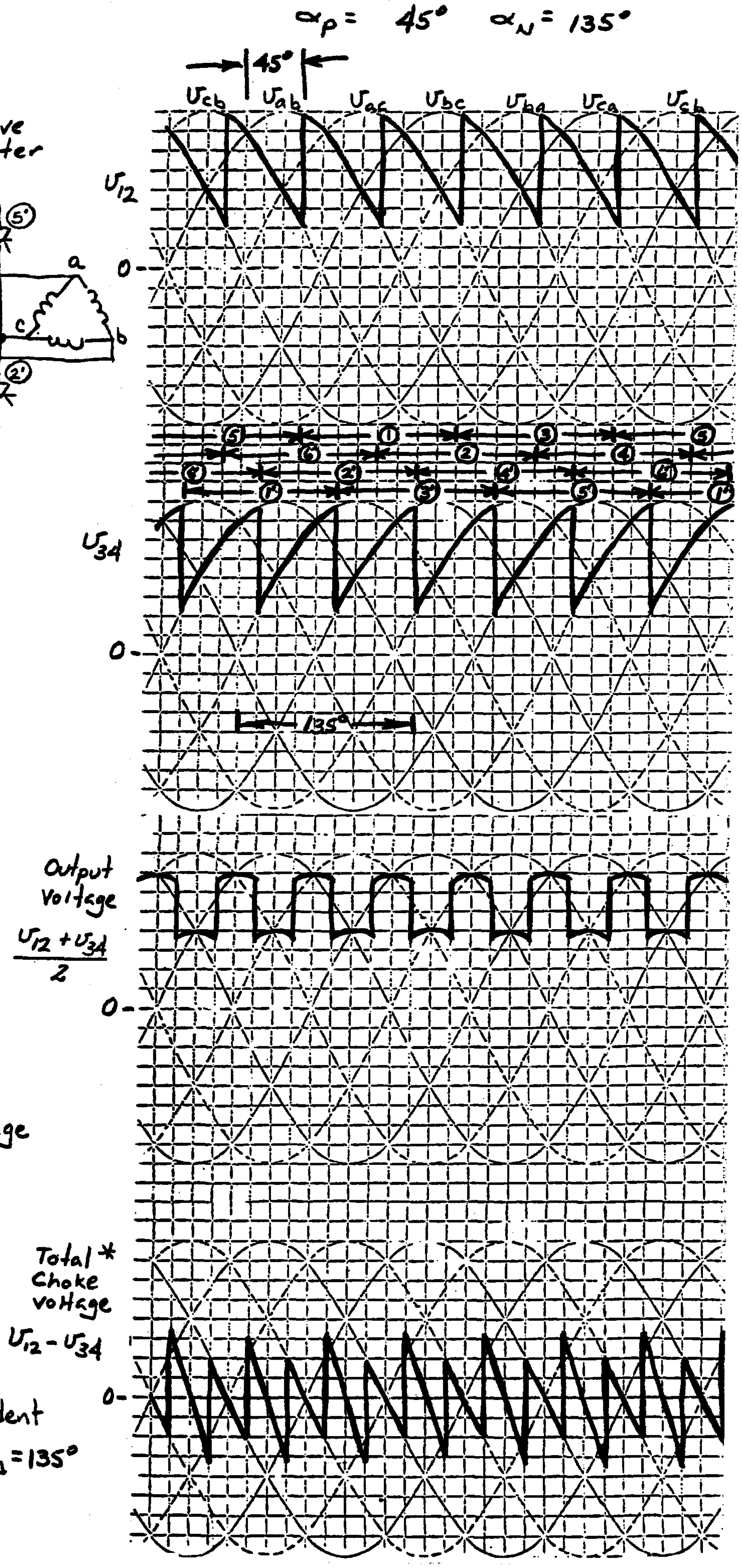
*Note: $U_{13} = U_{42} = \frac{1}{2}$ Total choke voltage

Figure A2 - Dual Converter, Independent Secondaries $\alpha_p = 30° \alpha_N = 150°$

Six Pulse Dual Bridge Converter



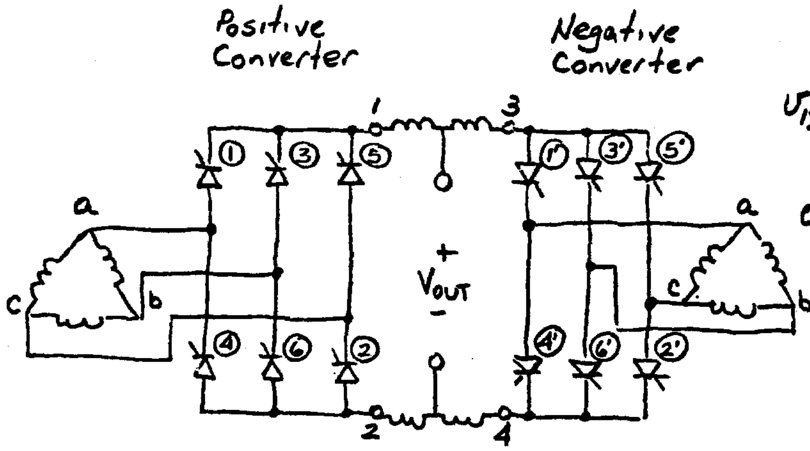


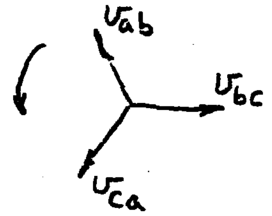


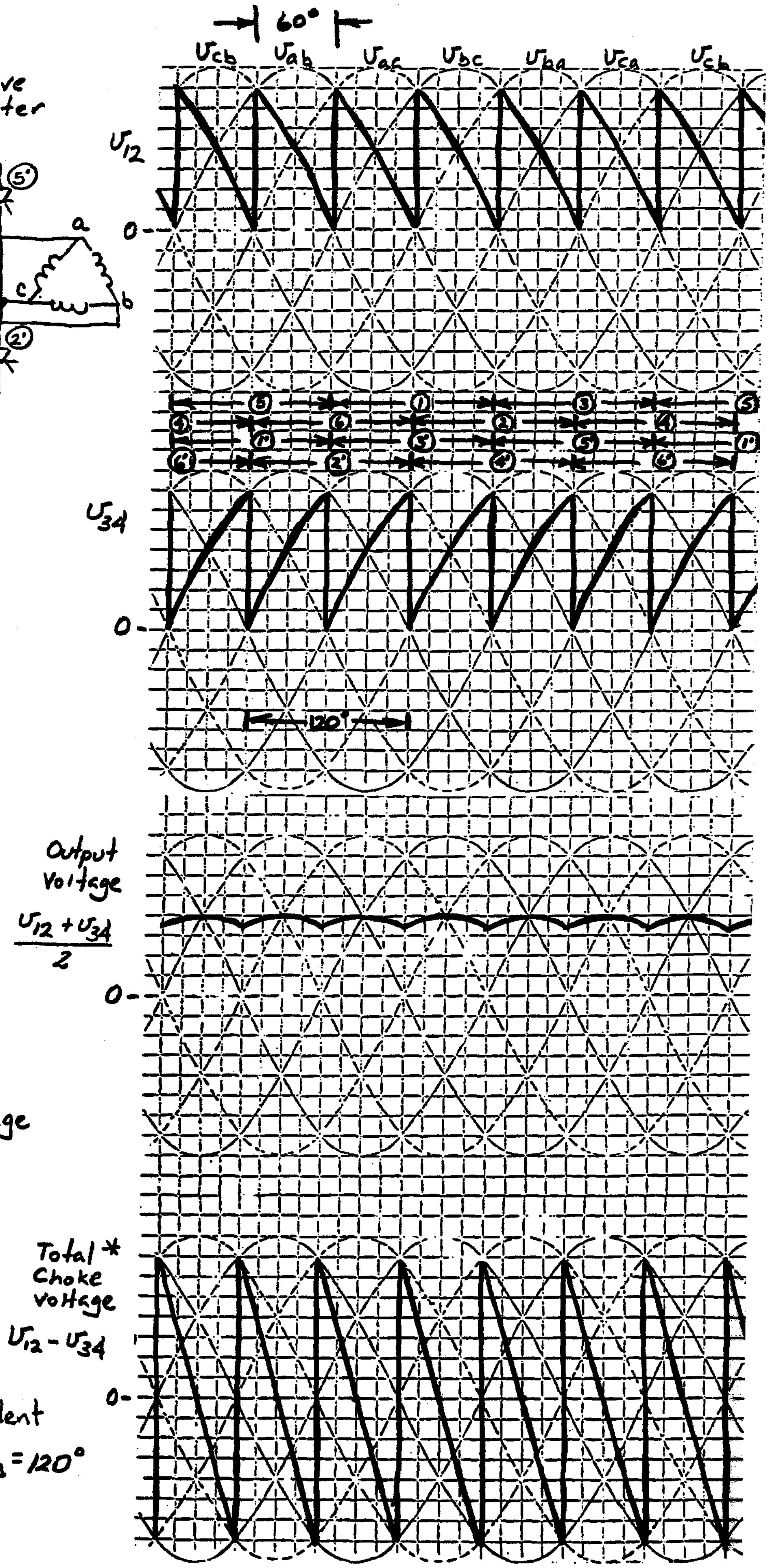
*Note: $v_{13} = v_{42} = \frac{1}{2}$ Total choke voltage

Figure A3 - Dual Converter, Independent Secondaries $\alpha_p = 45°$ $\alpha_N = 135°$

$\alpha_p = 60° \quad \alpha_N = 120°$



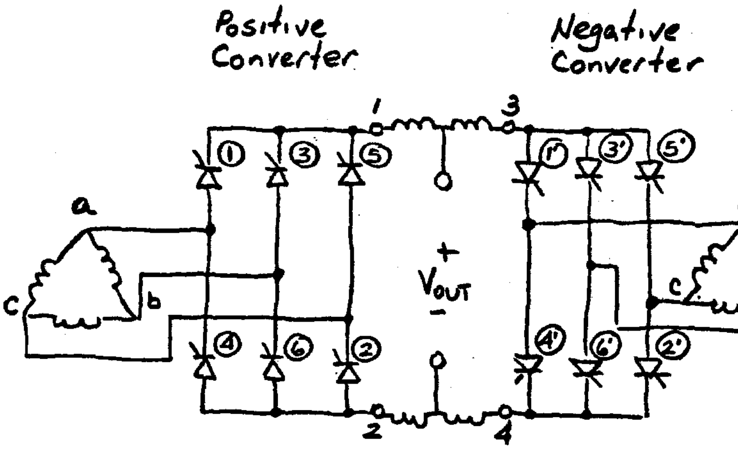


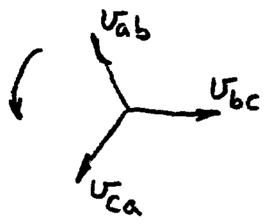


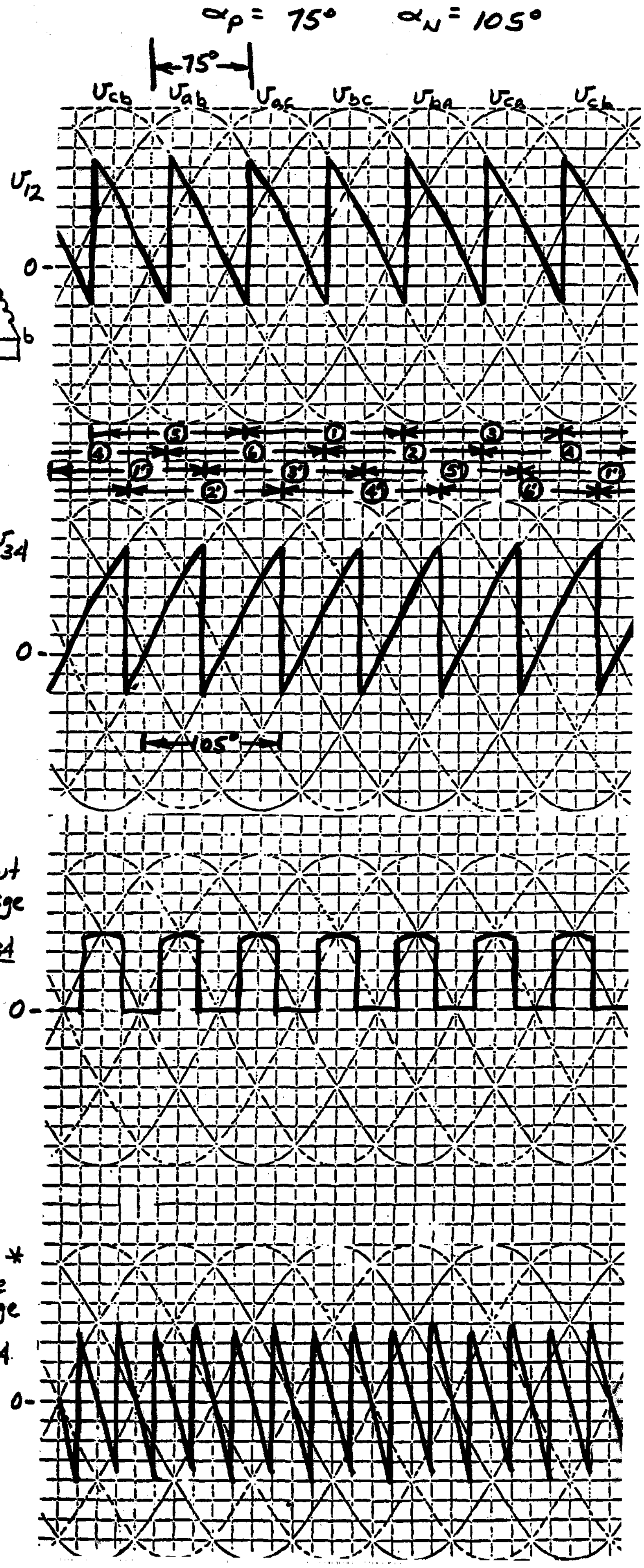
* Note: $v_{13} = v_{42} = \frac{1}{2}$ Total choke voltage

Figure A4 - Dual Converter, Independent Secondaries $\alpha_p = 60° \; \alpha_N = 120°$

Six Pulse Dual Bridge Converter



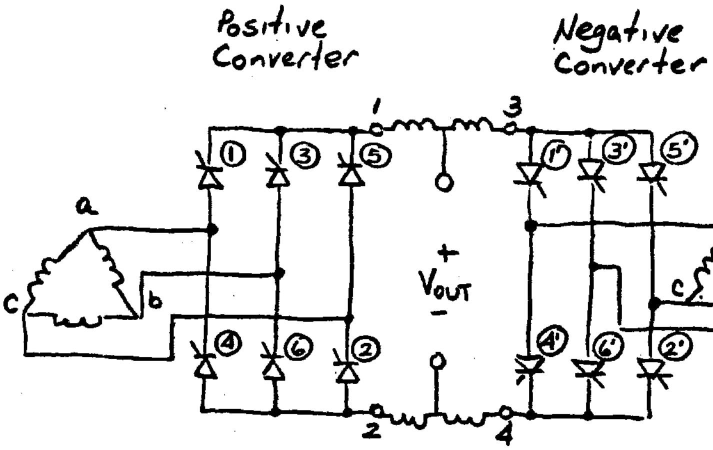


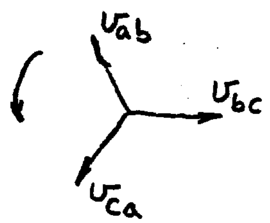


*Note: $v_{13} = v_{42} = \frac{1}{2}$ Total choke voltage

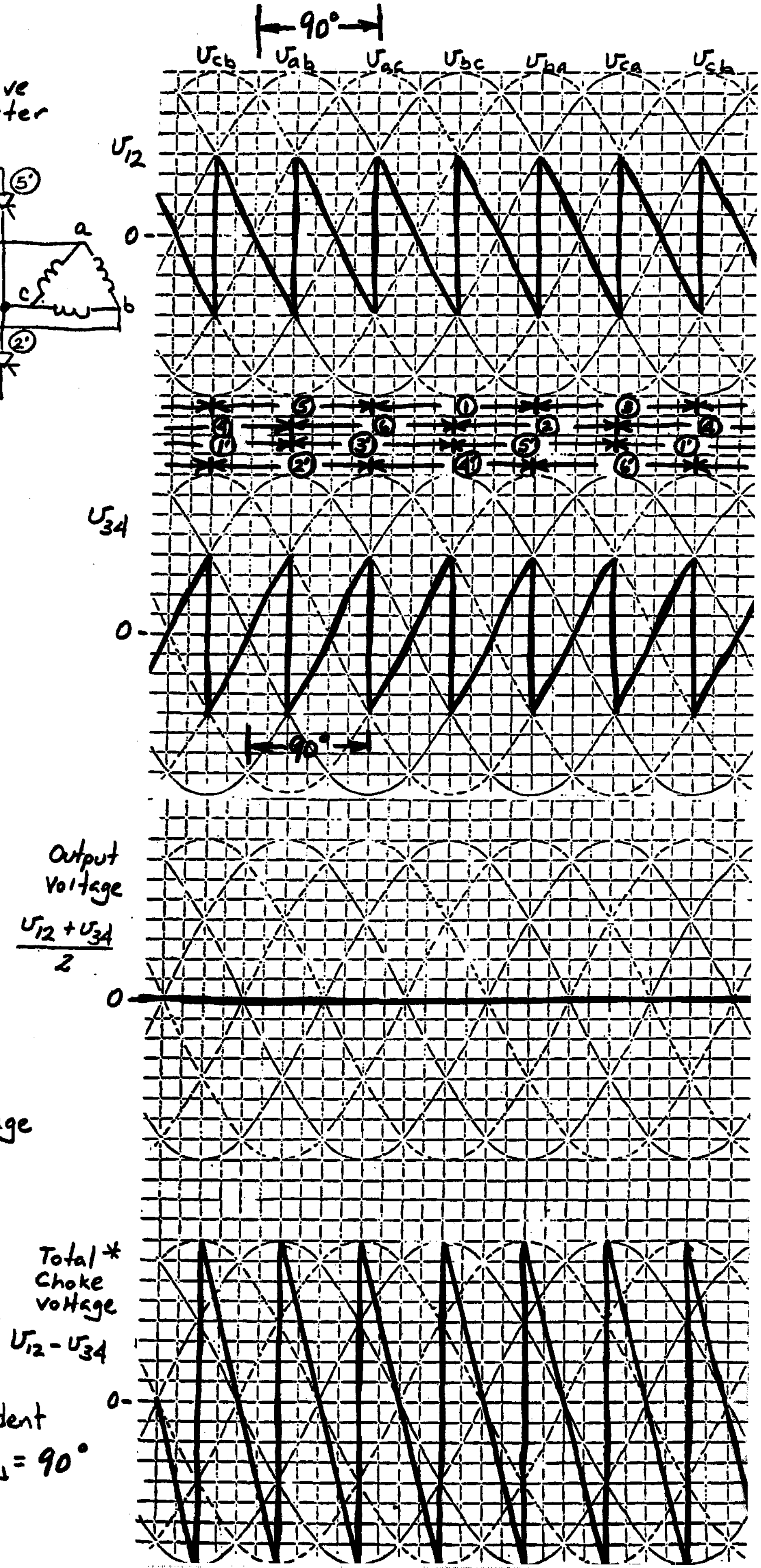
Figure A5 - Dual Converter, Independent Secondaries $\alpha_p = 75° \; \alpha_N = 105°$

Six Pulse Dual Bridge Converter





$\alpha_p = 90° \quad \alpha_N = 90°$

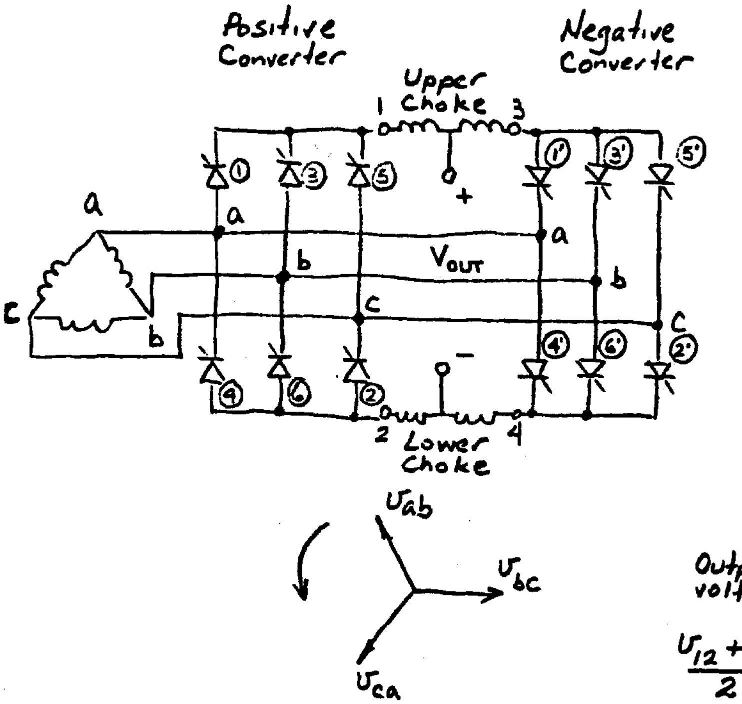


*Note: $v_{13} = v_{42} = \frac{1}{2}$ Total choke voltage

Figure A6 - Dual Converter, Independent Secondaries $\alpha_p = 90° \; \alpha_N = 90°$

Six Pulse Dual Bridge Converter

$\alpha_p = 15^\circ \quad \alpha_N = 165^\circ$



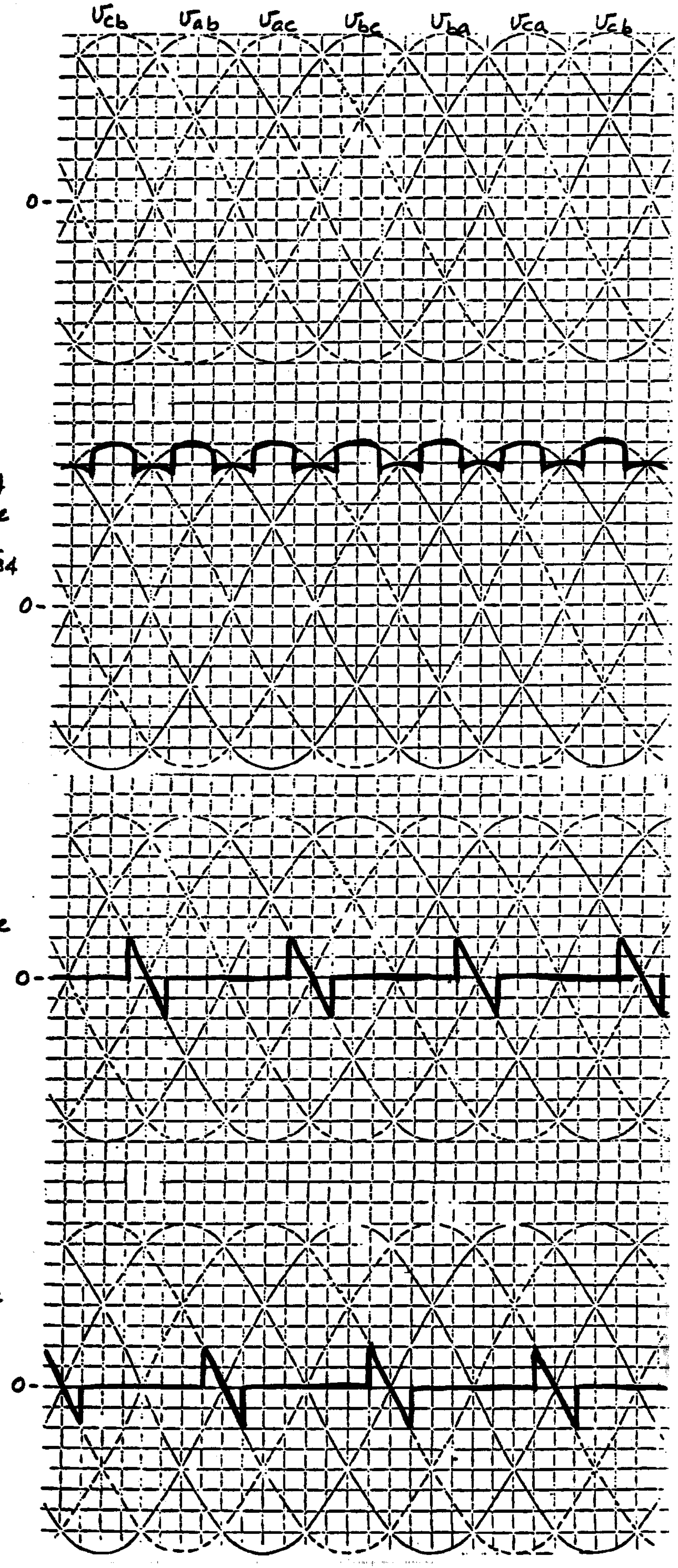
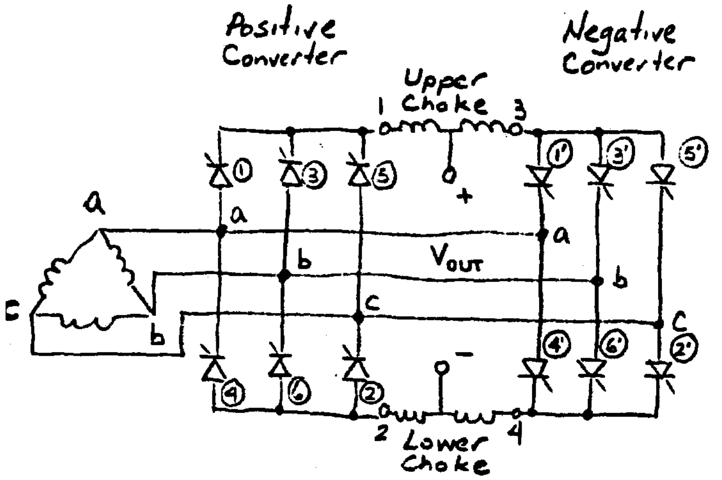
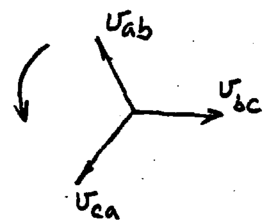


Figure A7 – Dual Converter, Single Secondary $\alpha_p = 15^\circ \alpha_N = 165^\circ$

Six Pulse Dual Bridge Converter

$\alpha_p = 30°$ $\alpha_N = 150°$





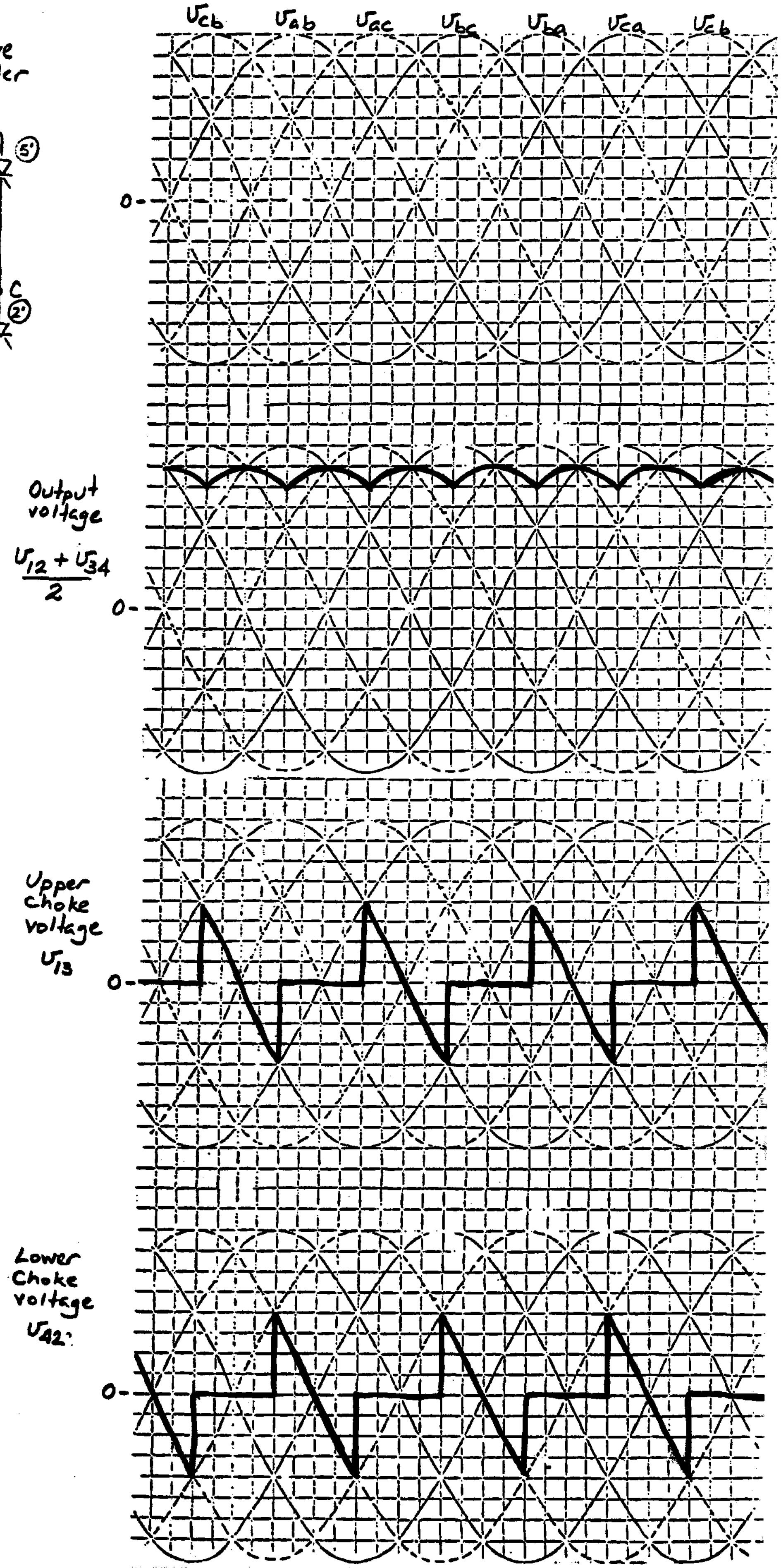
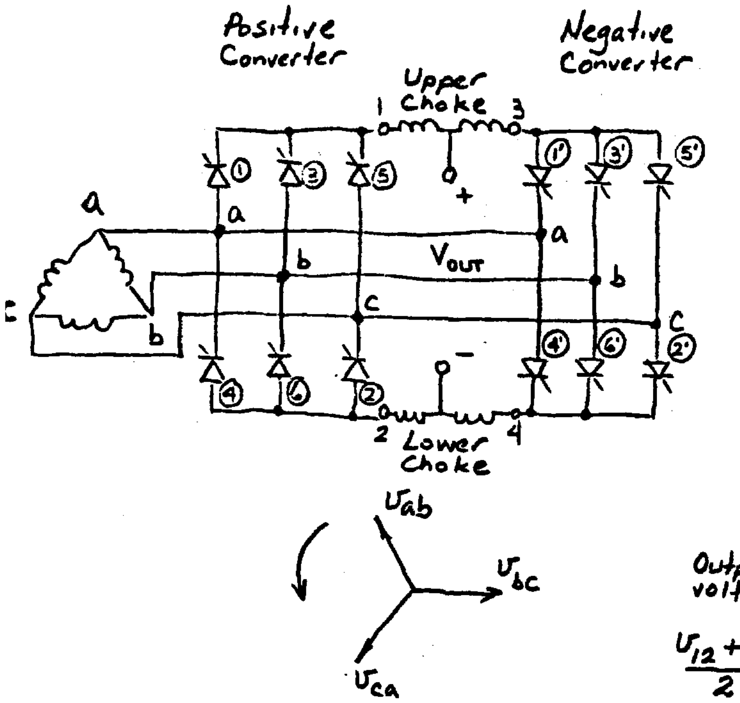


Figure A8 - Dual Converter, Single Secondary $\alpha_p = 30° \alpha_N = 150°$

Six Pulse Dual Bridge Converter

$\alpha_p = 45°$ $\alpha_N = 135°$



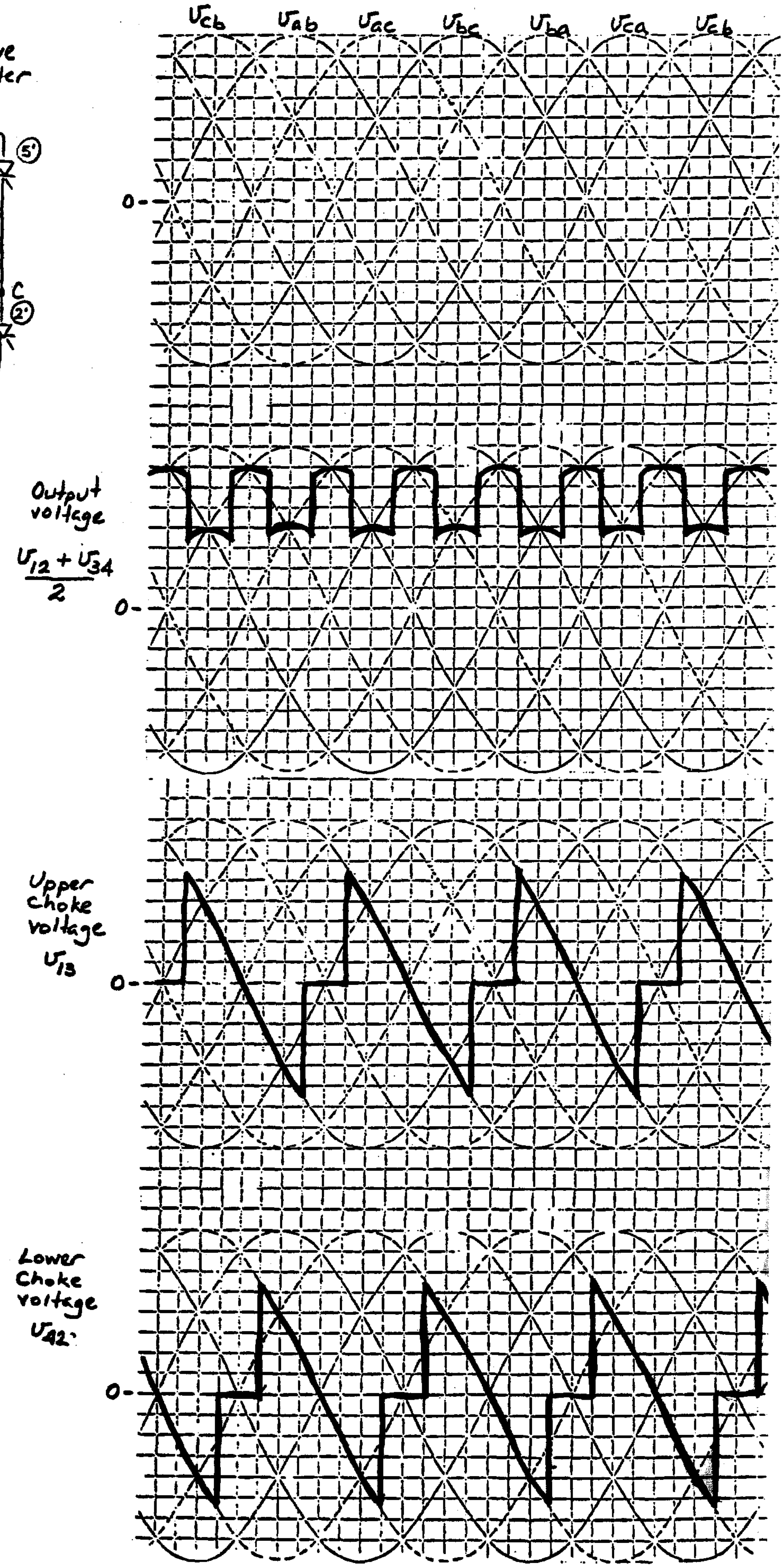
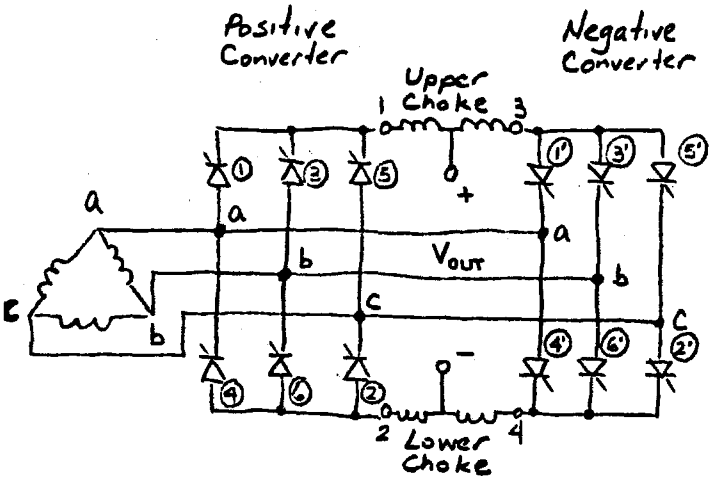
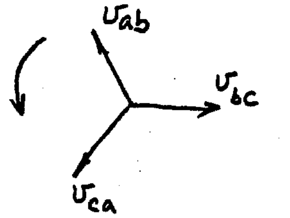


Figure A9 – Dual Converter, Single Secondary $\alpha_p = 45° \alpha_N = 135°$

Six Pulse Dual Bridge Converter

$\alpha_p = 60^\circ \quad \alpha_N = 120^\circ$





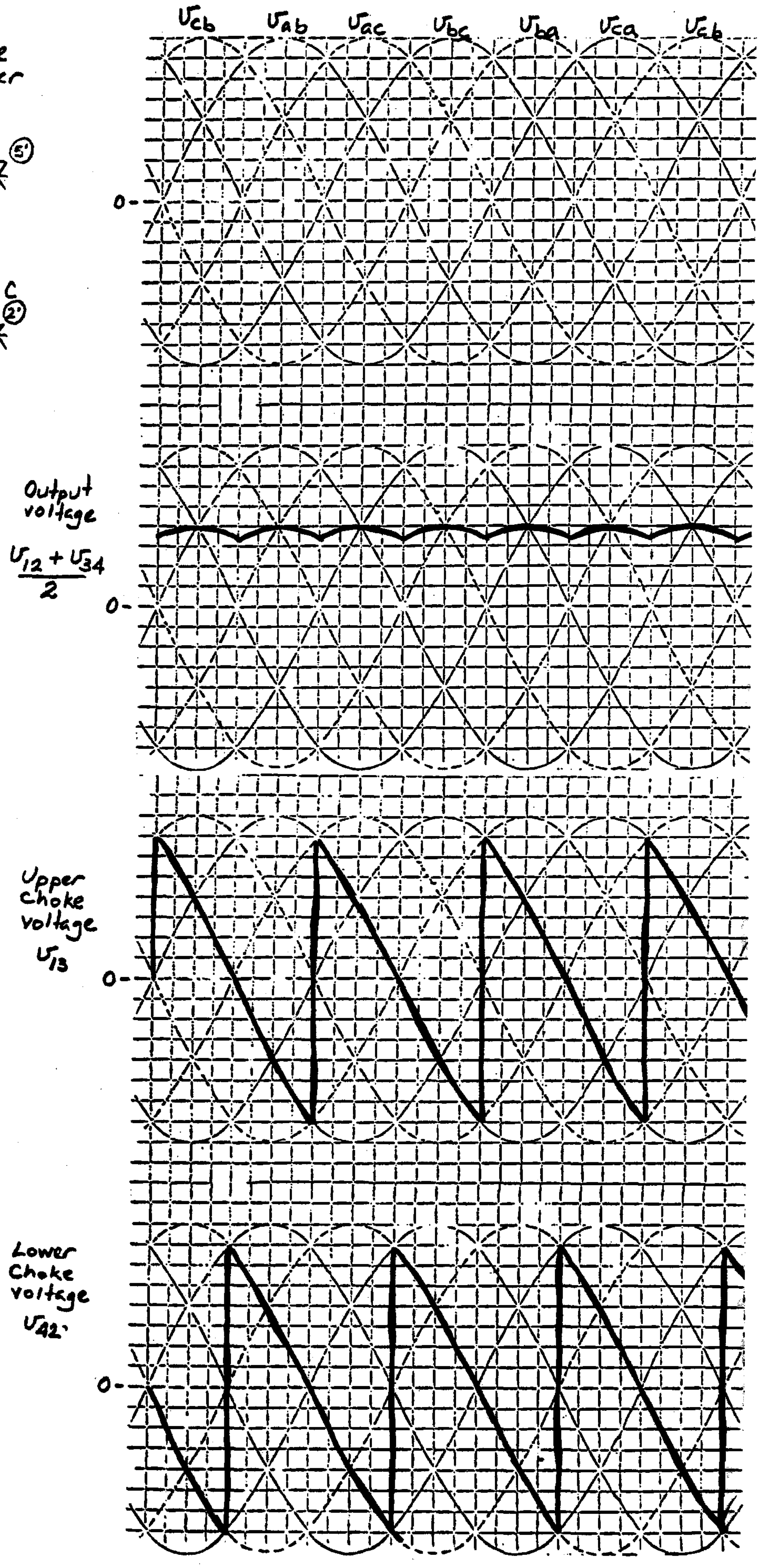
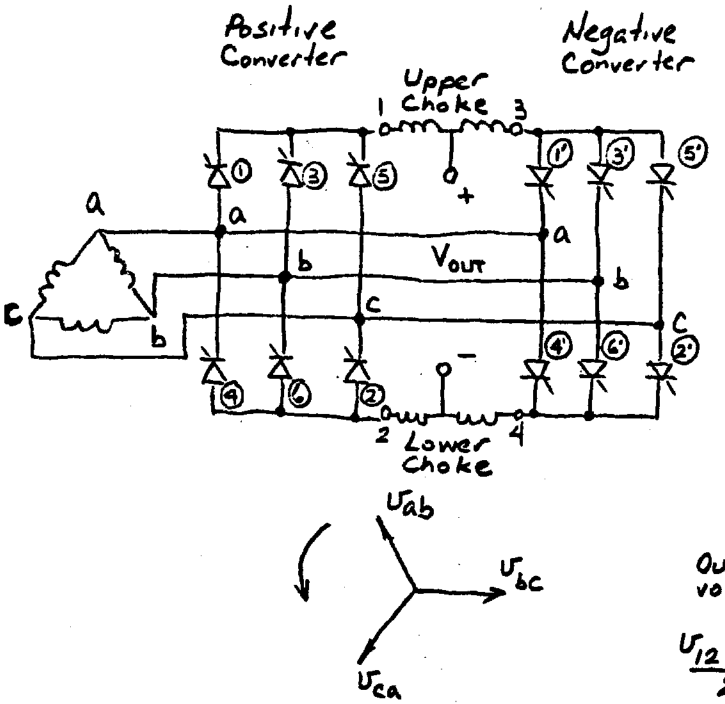


Figure A10 - Dual Converter, Single Secondary $\alpha_p = 60$ $\alpha_N = 120^\circ$

Six Pulse Dual Bridge Converter

$\alpha_p = 75°$ $\alpha_N = 105°$



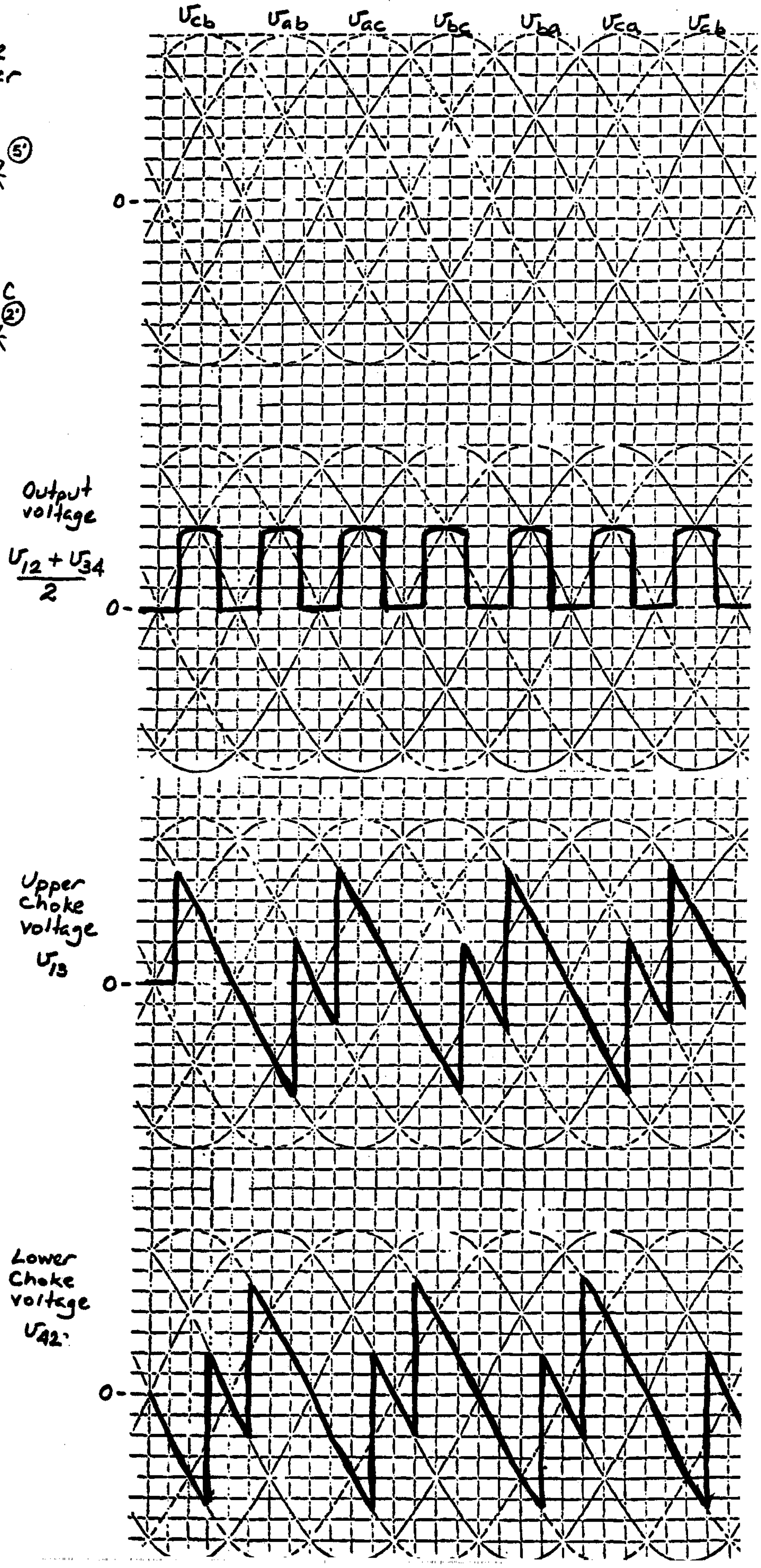
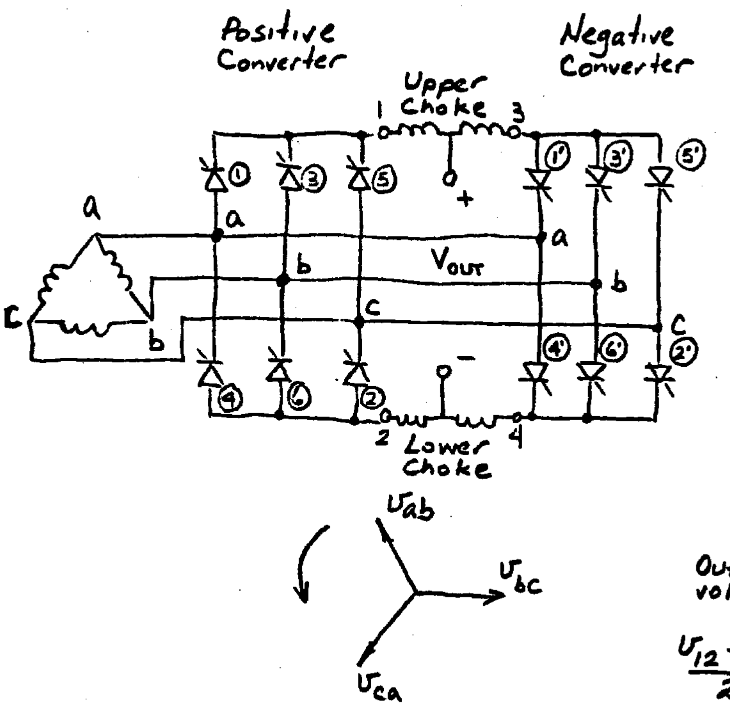


Figure A11 - Dual Converter, Single Secondary $\alpha_p = 75° \alpha_N = 105°$

Six Pulse Dual Bridge Converter

$\alpha_P = 90°$ $\alpha_N = 90°$



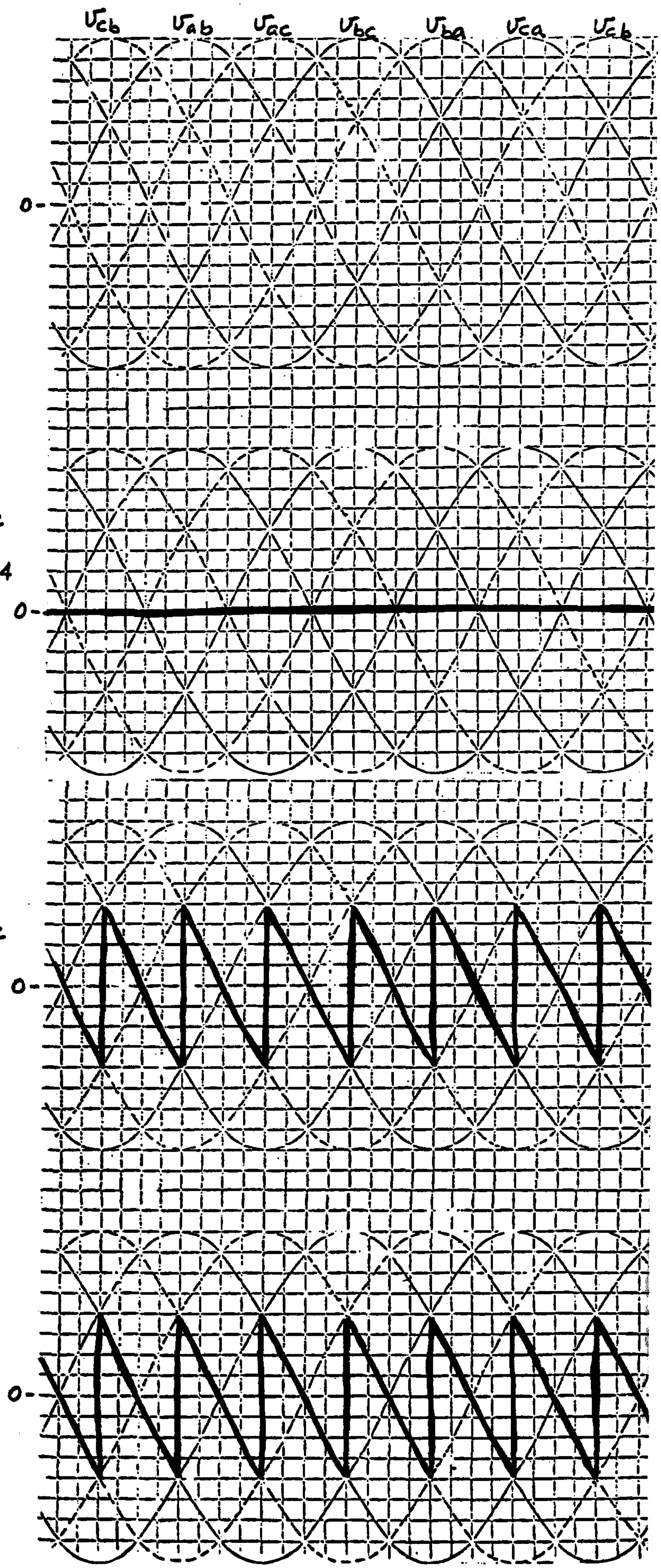
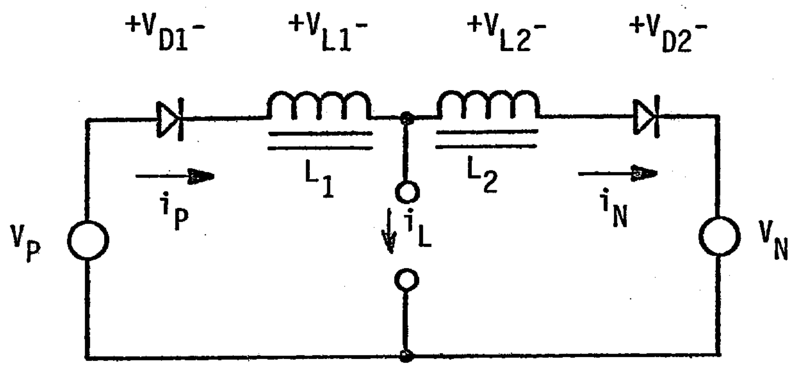


Figure A12 – Dual Converter, Single Secondary $\alpha_P = 90°$ $\alpha_N = 90°$

Appendix B

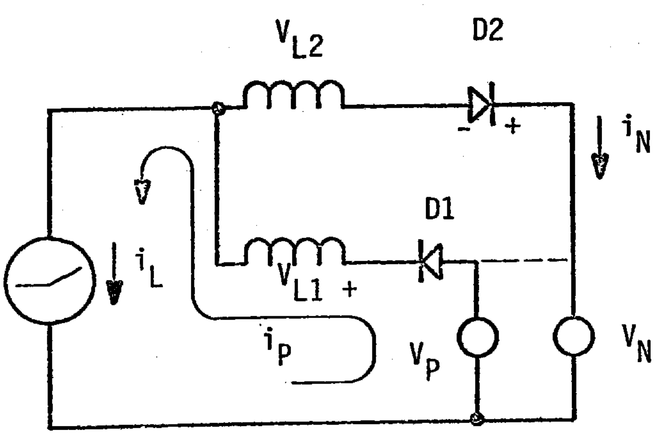
Dual Convecter Operation with Separate Inter-Converter Chokes

For purposes of analysis, the circuit in Figure 4c can be replaced with the following equivalent circuit wherein the converters are replaced by ripple-free sources and ideal diodes.



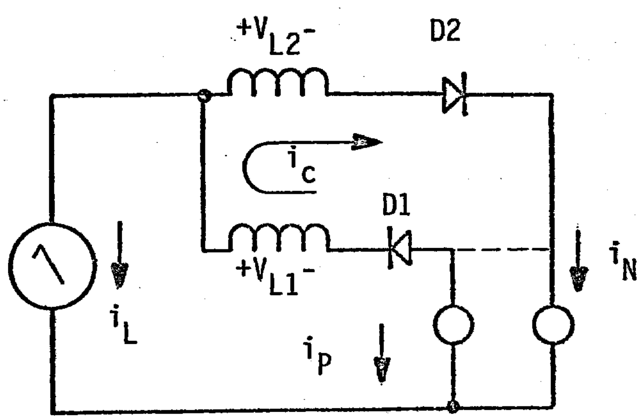
$L_1 = L_2 = L = 1H$
$V_P = V_N$
$N_1 = N_2 = N$
N_1 = Turns in L_1
N_2 = Turns in L_2

One way to visualize operation of this circuit is to assume an output current waveform as shown on the next page and evaluate the pertinent converter waveforms. In mode I, D1 is on, $v_{L1} = L_1(di_L/dt)$ and D2 is reversed biased by v_{L1}. In mode II, D2 turns-on and i_N flows when the voltage polarity across L_1 changes to allow i_L decrease.



MODE I

$V_{L1} = L_1(di_1/dt) = 5V$

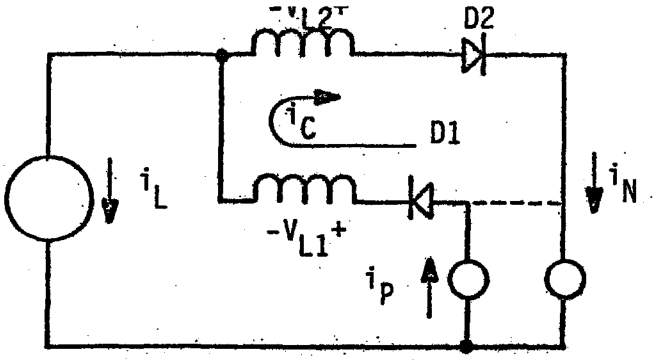


MODE II

$i_L = i_P - i_N$ $\quad pi_L = pi_P - pi_N$ where $p=d/dt$

$(V_{L1} - V_{L2})/1 = pi_L = -5A/sec$

$V_{L1} = -2.5V$, $V_{L2} = +2.5$

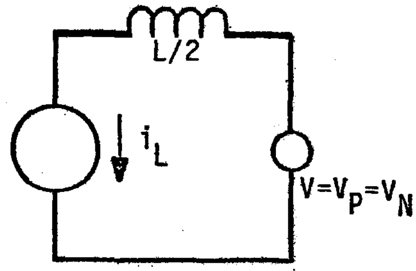


MODE III

$pi_L = pi_P - pi_N = +5A/sec$

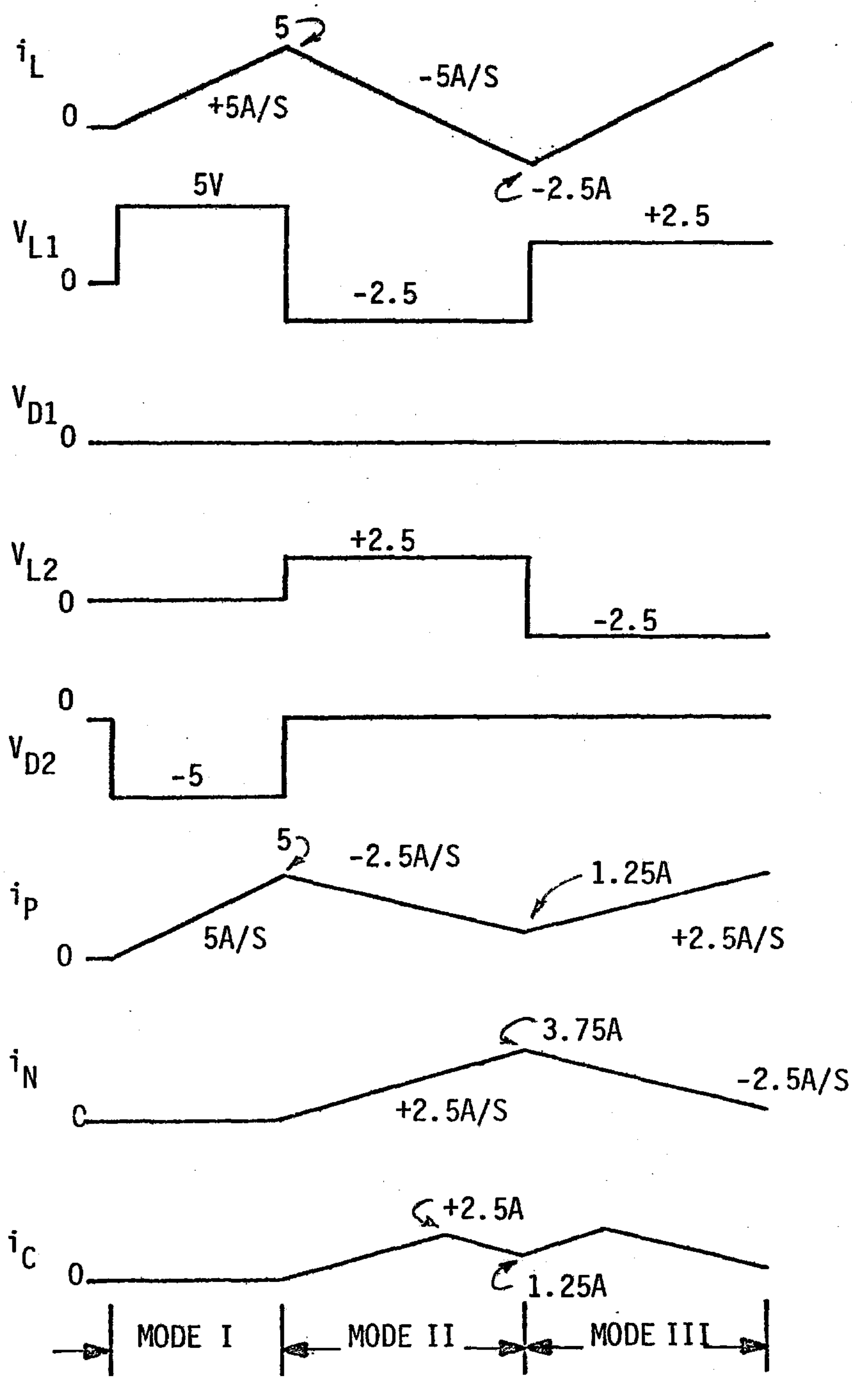
$(V_{L1} - V_{L2})/1 = +5$

$V_{L1} = +2.5V, \; V_{L2} = -2.5V$



STEADY-STATE EQUIVALENT CIRCUIT

WAVEFORMS

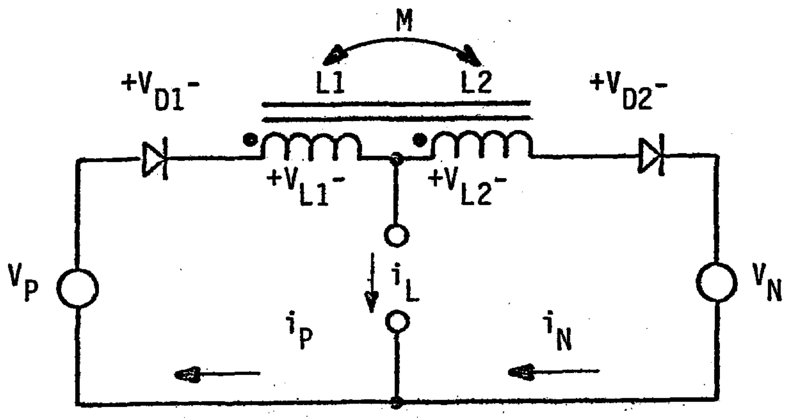


In mode III, load current increases and v_{L1} and v_{L2} change polarity from mode II. Current flowing in the converter not supplying the load can be considered circulating current. Modes II and III show that load current ripple tends to induce a circulating current in the dual converter with which the circulating current regulator must contend. Modes II and III also show that the chokes L_1 and L_2 appear to be in parallel and therefore the output impedance of this dual converter connection is L/2.

Appendix C

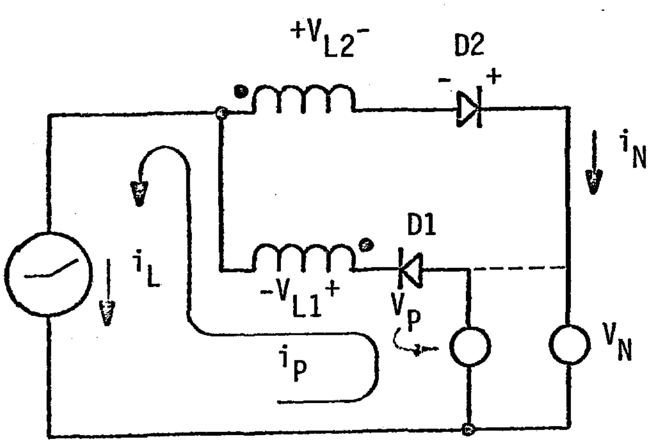
Dual Converter Operation with Dual Inter-Converter Chokes (mmf's Aiding)

The circuit in Figure 4b can be replaced with the equivalent circuit shown below wherein the converters are replaced by ripple-free sources and ideal diodes.

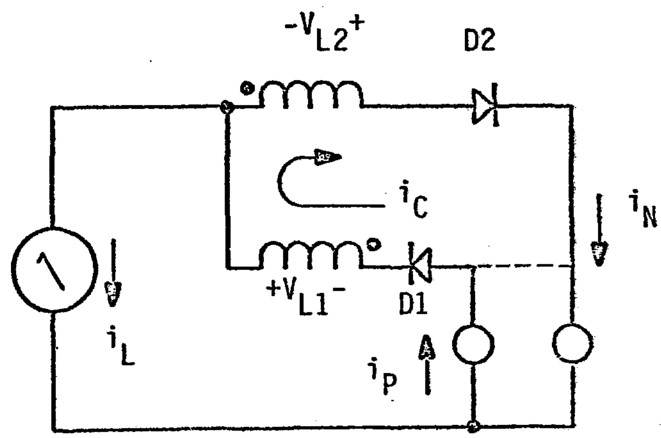


$L_1 = L_2 = L = 1H$
$V_P = V_N = V$
$K_m = 1$ (perfect coupling)

As in Appendix B, an output current waveform is assumed as shown on the next page, and the pertinent converter waveform are evaluated. In mode I, D1 conducts, $v_{L1} = L_1(di_L/dt)$, and D2 is reversed biased by $2v_{L1}$ due to the transformer coupling of L_1 and L_2. In mode II, the load current begins to decrease and the voltage across L_1 an L_2 attempts to reverse.



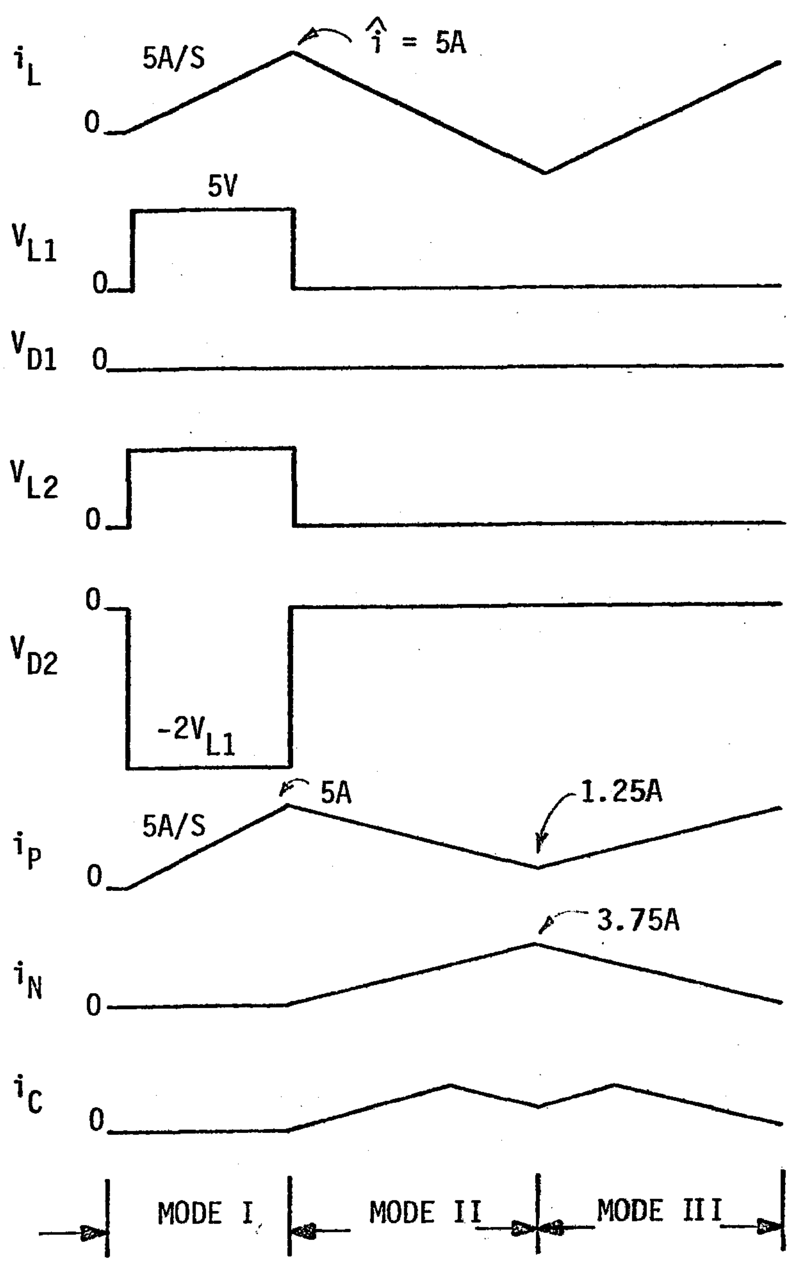
MODE I
$V_{L1} = L_1(di_L/dt) = 5V$



MODE II
$V_{L1} = V_{L2} = 0$

However since $v_p = v_n$, the diodes D_1 and D_2 are forced to conduct and the voltages across L_1 and L_2 is clamped to zero. It follows that the energy stored in the dual choke due to the peak load current is trapped as long as $v_p = v_n$. The total mmf $(N_1 i_p + N_2 i_n)$ of the dual choke is held constant, however, current is allowed to switch instantly between L1 and L_2 with zero voltage across L_1 and L_2. Thus, after the output current has reached a peak value, a decrease in the output current results in extra energy locked in the dual choke.

WAVEFORMS



The extra energy or mmf shows up as a circulating current between the converters as shown in the waveforms. In the absence of a circulating current or regulator there are losses within the converter which cause the circulating current or trapped energy to slowly decay. (The circulating current regulator will also tend to correct the induced circulating current). Neglecting these losses and the circulating current regulator, $Ni_p + Ni_N = N\hat{i}_L$, where $\hat{i}_L$ = peak load current. Since $i_L = i_p - i_N$ it follows that $i_p = (\hat{i}_L + i_o)/2$ and $i_N = (\hat{i}_L - i_o)/2$ as plotted in the waveforms. Operation in mode III is the same as mode II. Note the induced circulating current in modes II and III. From the forgoing discussion it follows that the output impedance of this dual-converter connection is zero once peak i_L has been reached.