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POWER SUPPLY HANDBOOK

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POWER SUPPLY HANDBOOK

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The circuit diagrams included in this handbook are intended to illustrate typical Power Supply applications and are not intended as constructional information. Although reasonable care has been taken in their preparation to insure their technical accuracy, no responsibility is assumed by Kepco, Inc., for any consequence of their use.

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PREFACE

The material in this handbook has its origin in the experience of Kepco engineers in the design and adaptation of regulated d-c power supplies for the divers requirements of industry. It never ceases to be a source of wonderment to us how varied are the possible applications for power supplies – particularly those applications having little to do with *traditional* power supply tasks.

In the Kepco Power Supply Handbook, we have attempted to set forth certain basic notions of power supply behavior in the hope that equipment users, faced with an unusual applications problem, may be inspired to its solution by the ideas that they find herein.

This volume supercedes an earlier Kepco publication, the *"Notes on Systems Applications of Regulated Power Supplies,"* from which much of Chapters Five and Seven has been derived. Chapters Three and Eight represent a new approach to the problem of special applications, in which the well established rules for operational amplifier manipulations are applied to the analysis of power supply behavior. Having drawn the analogy between regulated power supplies and d-c operational amplifiers, complex systems can be assembled using multiple power supply arrays, yet be explained in simple straightforward terms. The language of closed-loop control system engineering is heavily used for this purpose.

The circuit ideas are practical ones; in most cases they are derived from solutions to actual field problems involving the use of regulated power supplies in unorthodox ways. Many of the circuits have been proven and tested by the Kepco Applications Group – in working models and feasibility demonstrations. Others, such as the temperature controllers (Chapter Eight), are presently at work in our environmental laboratory.

Grateful acknowledgement is due to the contributions of Dr. Kenneth Kupferberg, Director of Engineering at Kepco, Inc., whose prolific ideas have contributed greatly at every stage in the development of this handbook. The Engineering Department at Kepco, Inc., under his direction, is responsible for the design and development of all the equipment described herein.

CONTENTS

LIST OF SYMBOLS

BRIDGE NOTATION: Voltage Bridge		BRIDGE NOTATION: Current Bridge		OPERATIONAL NOTATION		SYSTEMS NOTATION	
E_r	Reference voltage	E_r	Reference voltage	E_i	Input voltage	E_i	Input voltage
R_r	Reference resistor	R_r	Reference resistor	R_1	Summing resistor	R_c	Coupling resistor
R_{vc}	Voltage control	R_{cc}	Current control	R_f	Feedback resistor	R_f	Gain control
I_b	Bridge current	I_b	Bridge current	I_b	Loop current	I_b	Control current
$G = \frac{R_{vc}}{R_r}$	Closed-loop gain	$G = \frac{R_{cc}}{R_r}$	Closed-loop gain	$G = \frac{R_f}{R_1}$	Closed-loop gain	$G = \frac{R_f}{R_1}$	Transfer gain

A	Open-loop gain	E_o	Output or compliance voltage	e	A-c component of voltage	M	(10^{+6}) meg
ϵ	Error voltage	I_o	Load or output current	i	A-c component of current	h	Henry
E_u	Unregulated d-c	E_{cc}	Current control voltage	t	Time	f	Farad or frequency
E_p	Pass voltage	C_o	Output capacitance	T	Period	cps	Cycle per second
R_s	Current sensing resistance	C_f	Feedback capacitance	τ	Time constant	M	Meter
E_c	Command input	E_f	Feedback voltage	μ	(10^{-6}) micro	Flux-O-Tran®	Flux oscillating transformer
E_b	Feedback input	a-c	Alternating current	m	(10^{-3}) milli		
R_l	Load resistor	d-c	Direct current	k	(10^{+3}) kilo		
VIX®	Voltage/current crossover	SCR	Silicon controlled rectifier				

1 INTRODUCTION TO THE REGULATED POWER SUPPLY

Power supplies encompass many different devices including such prime sources as batteries and generators, as well as electronic converters, whose task it is to convert the output of a prime source to a useful form. We will concern ourselves herein with converters designed for the purpose of changing a-c utility-line electrical power to d-c. There are other kinds of converters, of course: d-c to a-c (inverters), a-c to a-c (frequency changers) and d-c d-c. For our purposes, however, the term "power supplies" will refer to *a-c – d-c* types.

1.1 BASIC RECTIFIER CIRCUITS

The simplest kind of a-c – d-c power supply is a simple rectifier, arranged in any of a large variety of half-wave, full-wave, bridge, and multi-phase circuits (with or without transformer). The rectifier forms the heart of such a power supply. It may be a thermionic vacuum tube or semiconductor diode, but whatever the form, its task is to convert an alternating current to a direct one. Table 1.1 summarizes the characteristics of the more common rectifier circuits.

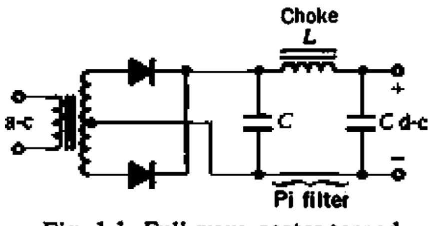
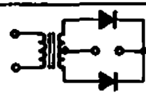
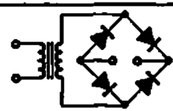
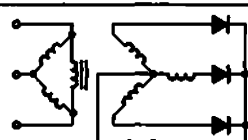
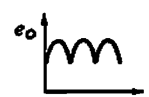
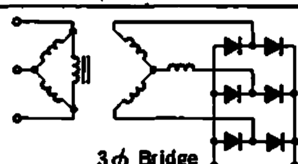
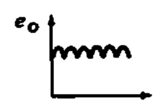


Fig. 1.1 Full wave center tapped

The basic transformer-rectifier circuit can be used directly or, more often, in conjunction with filters to supply a load. A common filter is the pi circuit shown in Fig. 1.1, in which capacitors and

Table 1.1 Simple Rectifier Circuits

Schematic	1 Cycle Output Waveform	Average d-c Volts Output	RMS Volts Output	Peak Volts Output	Peak Reverse Rectifier Voltage	Percent Ripple
 1ϕ Half wave		1	1.57	3.14	1.41	121%
 1ϕ Full wave ct.		1	1.11	1.57	2.82	48%
 1ϕ Full wave bridge		1	1.11	1.57	1.41	48%
 3ϕ Star (wye)		1	1.02	1.21	2.45	18.3%
 3ϕ Bridge		1	1.00	1.05	2.45	4.2%

inductors are used to smooth the output. The circuit shown is capable of producing rather well filtered d-c voltage, and is widely used in electronic equipment of all descriptions. In many critical applications, however, and in most laboratory situations, there is need for more flexible and precise power supplies.

The typical laboratory power supply will usually incorporate some means for controlling or adjusting its output, and must exhibit high immunity to a wide variety of external influences, as for example, loading, variation in its primary supply, temperature, time, etc.

1.2 VOLTAGE REGULATION

Voltage regulation is what most engineers envision as the desirable characteristic of a power supply. This means that they expect to be able to specify or define or adjust the *voltage* output of the power supply independently of all other parameters. How independently? Well, that's where regulation comes in. In ideal form, the voltage-regulated power supply becomes a "voltage source" with no ripple, zero internal resistance, and a precisely definable terminal voltage that is constant for all time. It can be plotted as a straight line, such that no matter what the load current, the terminal voltage remains constant. Since a real device is bounded, the practical power supply approximates a voltage source only to the limit imposed by a finite maximum current.

1.3 CURRENT REGULATION

It is also possible to produce power supplies with output characteristics that are the dual of voltage regulation. These are current regulators in which the object is to be able to precisely define or adjust the *current* output of the power supply in an independent manner. Ideally, the current source would have an infinite internal impedance, and would be able to produce an infinite voltage if open circuited. A practical current regulator only approximates the current source characteristics, being, of course, bounded by a maximum open circuit voltage. A tabulation of the comparative specifications for voltage and current regulation is given in Fig. 1.2.

1.4 PRACTICAL POWER SUPPLY

Any practical power supply will lie somewhere between the ideal voltage and ideal current source characteristics.

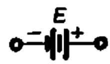
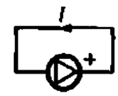
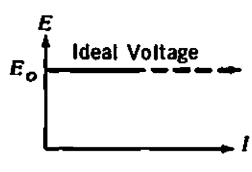
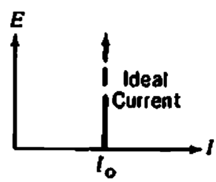
Characteristic	Voltage Source 	Current Source 
Controlled Parameter	Voltage	Current
Load	Current	Voltage
Idle	Open Circuit (Zero Current)	Short Circuit (Zero Volts)
Overload	Short Circuit	Open Circuit
Internal Source Resistance	Zero Ohms	Infinite Ohms
E, I Terminal Characteristic		

Fig. 1.2

Since they are dual, the presence of voltage source characteristics *degrade* the regulation of a current regulator, while, conversely, current source characteristics degrade the performance of a voltage regulated power supply. (See Fig. 1.3.)

A power supply is described as either a voltage regulator or a current regulator by virtue of its approximation of either of the ideal characteristics.

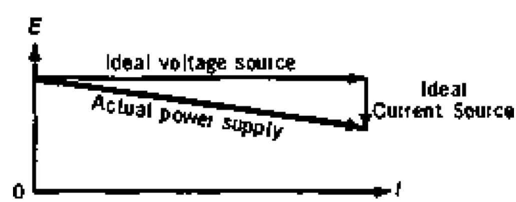


Fig. 1.3 An actual power supply output *E, I* characteristic constructed from an ideal voltage source and ideal current source. By the steepness of its *E/I* slope, the power supply can be made to resemble a voltage or a current generator. The slope shown resembles a voltage regulator, but a relatively poor one.

A power supply designed for voltage regulation will lie much closer to the ideal (horizontal) voltage source characteristic, while a supply designed for current regulation will have a characteristic that lies close to the (vertical) current source characteristic.

Power supplies are usually designed to resemble either ideal voltage or ideal current sources – at least over part of their range. More sophisticated power supplies may be designed to closely approximate both kinds of sources over the appropriate parts of their output range as shown by Fig. 1.4. This kind of power supply, called an automatic crossover supply, will be discussed in some detail in Chap. 4.

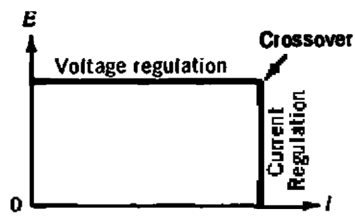


Fig. 1.4 Combined voltage and current regulator in an automatic crossover power supply.

The simple power supply of Fig. 1.1 can be made more versatile by adding a variable auto-transformer, as seen in Fig. 1.5. Such a device provides a set of adjustable transformer taps with a sliding contact (brush) that permits the output voltage to be varied conveniently by the user.

This simple circuit is still something less than a laboratory grade voltage regulator power supply. For one thing, its internal resistance is quite a bit different from zero. There are resistance contributions from the transformer, the rectifier, the choke, and indeed, even the wiring, all of which can add up to quite a significant internal resistance.

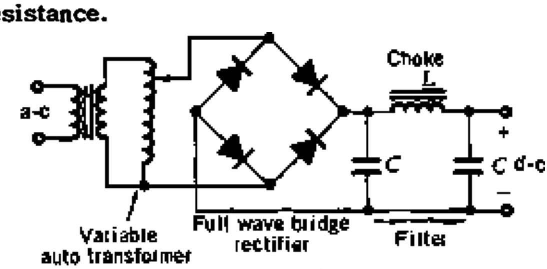


Fig. 1.5

When a load current is drawn from such a power supply, the voltage drop across the sum of the internal resistances subtract from the available terminal voltage. If the variable auto-transformer is not re-adjusted, the output voltage characteristic versus load can be depicted in Fig. 1.6.

Fig. 1.6

The amount of the change in output voltage ΔE is expressed as the load regulation of the supply. It may be expressed directly in volts or as a percentage change in output.

All materials change character slightly as a function of temperature. Copper resistance, for example, changes 0.39% per °C. If copper wire is used to make the power supply in Fig. 1.3, then the amount of series internal resistance will change somewhat as a function of temperature, giving rise to a temperature coefficient of output voltage (under load).

Should the primary a-c line voltage vary, its changes are passed through the transformer and rectifier to appear as an equal percentage variation in the d-c output. Line regulation is usually expressed as the amount of output d-c change that a given a-c input change will produce. If both are expressed as a percentage, then the effect of the transformer turns ratio can be ignored. The line regulation can also be expressed as the ability of a circuit to attenuate input line variations.

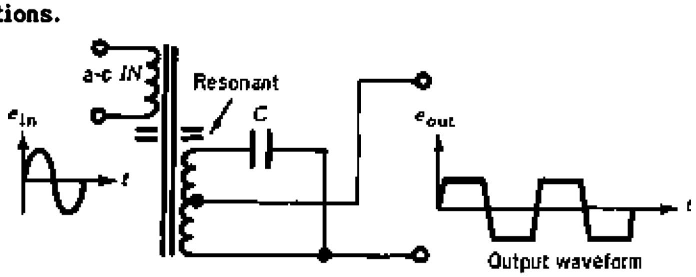
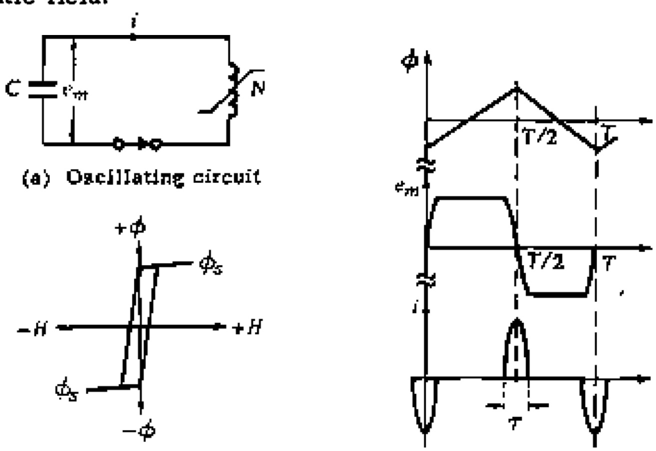


Fig. 1.7 Kepco Flux-O-Tran

A power supply possessing significant advantage over the simple circuits described above results from the use of a "regulating" transformer in place of the ordinary step-up/down power transformer.

One such transformer is the flux-oscillator resonant transformer, called Flux-O-Tran®. (See Fig. 1.7.)

The Flux-O-Tran consists of a primary and secondary wound separately, and coupled by a magnetic structure. As shown, a special magnetic shunt separates them. The secondary is connected to a capacitor and is made resonant at the line frequency, e.g., at 60 cps, with fairly high Q. When excited by a 60 cps line source, the resonant circuit builds up enough flux to saturate the iron core of the transformer on each alternate half cycle. The energy is interchanged back and forth between the magnetic field and the capacitor's electrostatic field.



(a) Oscillating circuit

(b) Magnetization characteristic

(c) Circuit waveforms

Fig. 1.8

The operation of Kepco's Flux-O-Tran may be understood by reference to the equivalent circuit of Fig. 1.8a. It consists of a capacitor connected to an N-turn winding whose core is made of material with a nearly rectangular magnetization characteristic (Fig. 1.8b). To begin a cycle, assume that the capacitor C has been charged to potential e_m by a current i which has left the core in its negative saturation region ($-\phi_s$ in Fig. 1.8b). The voltage e_m across the coil begins to change the flux from $-\phi_s$ to $+\phi_s$. Only a negligible magnetizing current is required to produce this change, as implied by the steep slope of the magnetizing curve. The capacitor's voltage e_m, therefore, remains substantially constant. The flux, changing linearly with time, reaches $+\phi_s$ in a time $T/2$ (Fig. 1.8c), whereupon the core saturates. In its saturated state, the residual inductance becomes small, since the winding N behaves like an air core inductor. The residual inductance is L_r.

The capacitor now rapidly discharges through L_r, building up a magnetic field around L_r. When C is fully discharged, this field begins to collapse, inducing a charging current in the capacitor which tends to charge it in the opposite direction. This starts the flux changing back from $+\phi_s$ to $-\phi_s$.

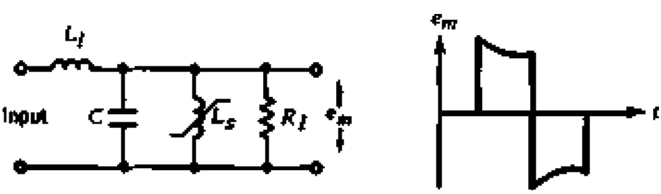
The time for the change of polarity across C is

$$\tau = \pi \ L_r C$$

Since L_r is very small, the transistion time is extremely short. The period of the complete cycle is

$$T = \frac{4N\phi_s}{e_m}$$

Without excitation, oscillations would slowly decay due to losses in the circuit. In operation, however, energy is supplied by exciting the circuit from an a-c source. Figure 1.8d shows an equivalent circuit for a line-excited flux-oscillator. In practice, of course, a separate primary winding is used to excite the oscillator.



(d) Equivalent circuit for Flux-O-Tran

(e) Output waveform from Flux-O-Tran

Fig. 1.8

In Fig. 1.8d, a linear inductor L_l is shown coupling the oscillator to an a-c power line. A resistor R_l is connected in parallel across the oscillating circuit to represent the load (including the losses of L_s). The line frequency must be close to the natural resonance of the oscillator ($1/T$), as given above. With the operating frequency fixed by the excitation frequency, the capacitor voltage e_m can be expressed in terms of that frequency as

$$e_m = 4N\phi_s f$$

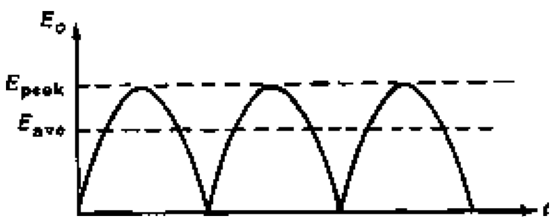
For a constant frequency, the capacitor voltage is constant and is essentially independent of the input amplitude, provided that the input is large enough to sustain oscillation. The voltage is no longer rectangular, but assumes the shape shown in Fig. 1.8e. The impedance of the linear inductor L_l must be sufficiently high so as to provide isolation from the line, but not so large that it prevents the transfer of the requisite output power. Additionally, L_l is chosen so that it forms a series resonant circuit with C to provide sufficient voltage across C to be sure of saturating L_s for a wide range of input voltages.

Output from the transformer may be obtained by means of a tap (as shown in Fig. 1.7) or by means of a separate output winding.

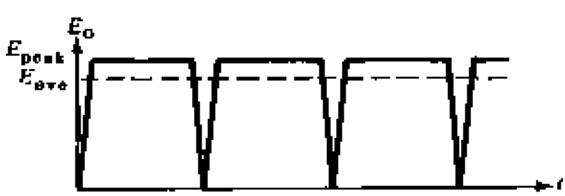
The advantages of this circuit are several:

(1) The peak-to-peak output voltage (and thus the d-c from a rectifier-filter) is primarily dependent only on the magnetic structure – not the magnitude of the line voltage. Thus, the flux-oscillator provides a good degree of line regulation.

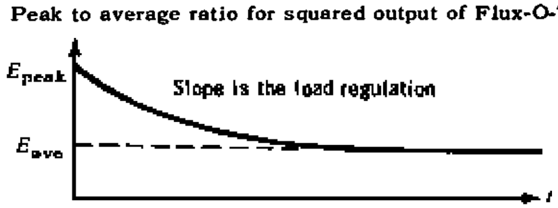
(2) The output waveform is squared, enhancing rectifier utilization. Also, the peak-to-average voltage ratio is reduced, which improves the load regulation.



(a) Peak to average ratio for ordinary full wave rectified sinusoid



(b) Peak to average ratio for squared output of Flux-O-Tran

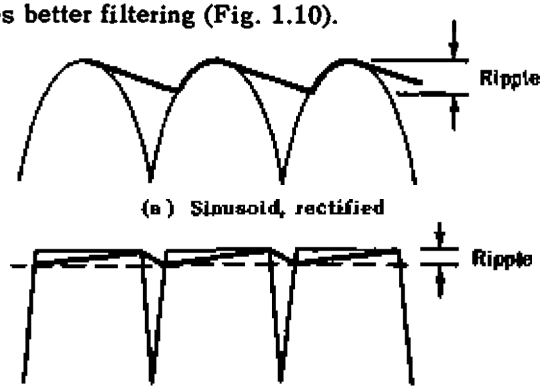


(c)

Fig. 1.9 Reduced peak to average ratio flattens load regulation slope.

Because the load regulation to a large degree reflects the peak charging of the filter capacitors at no load, reducing the peak-to-average ratio significantly improves the load regulation. The capaci-

tors also discharge for a shorter period with the squared flux-oscillator waveform, the ripple currents are reduced, and so a given filter capacitor provides better filtering (Fig. 1.10).



(b) Squarewave, rectified

Fig. 1.10

The transformer is immune to short circuits. Whenever the output of a Flux-O-Tran is short circuited, the equivalent circuit modifies to that of a linear inductance in series with the output current, which limits the output to a safe maximum (Fig. 1.11).

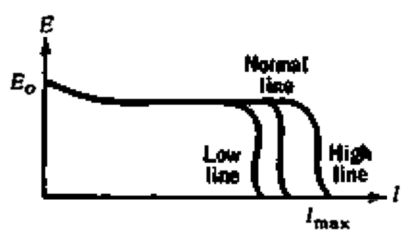


Fig. 1.11 Output overload curves for Flux-O-Tran as a function of line voltage.

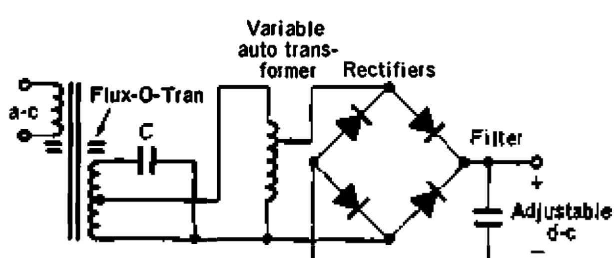


Fig. 1.12 Adjustable line regulated power supply using the Flux-O-Tran power transformer, Kopco PR design.

The addition of the Flux-O-Tran power transformer to the variable transformer circuit gives fairly good voltage regulation characteristics with a minimum of circuit complexity, and hence, unsurpassed reliability. The Kepco PR Series power supplies are of this type, as shown by Fig. 1.12.

The Flux-O-Tran is particularly useful in fixed output power supply situations such as the Kepco PRM design of fixed voltage modular supplies. (Fig. 1.13).

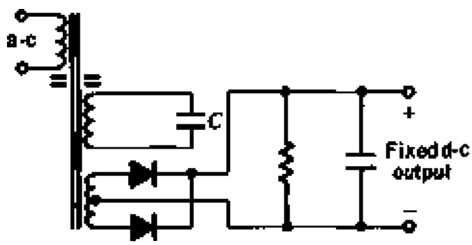
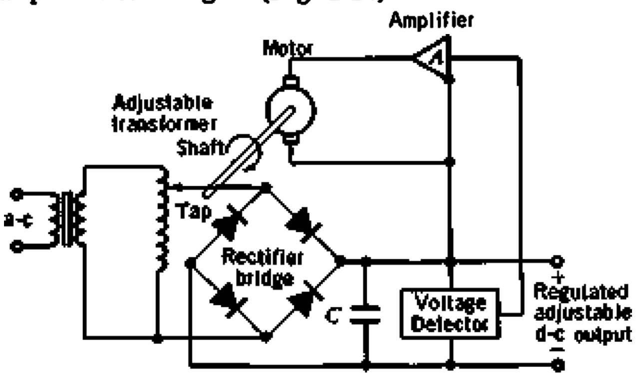


Fig. 1.13 Kepco PRM design, fixed output, line regulated module, typical schematic.

1.5 FEEDBACK REGULATORS

The most important step that can be taken to improve regulation performance is the introduction of feedback. This involves active control, employing an output sampling device, reference source, a comparison device and an amplifier whose gain acts upon the error to correct the output.

A classic regulator consists of a motor coupled to a variable auto-transformer to physically position its output tap in accordance with an amplified error signal (Fig. 1.14).



1.14 Feedback regulator system

A major part of the art of regulated power supply engineering is the design of better feedback control systems employing electronic control means to detect, amplify, and respond to the derived error. The heart of such a system is the device or devices which act upon the rectified d-c, actually performing the regulation. In addition to motor driven variable transformers – saturable reactors, thyratrons, silicon controlled rectifiers, vacuum tubes and transistors are all used in regulator circuits. Many of these circuits are described in following chapters.

2 THE CONCEPT OF THE BRIDGE REGULATOR

2.1 ELEMENTARY DIFFERENTIAL CIRCUIT

The heart of a Kepco regulated d-c power supply is its comparison bridge circuit. The comparison bridge is an arrangement of reference voltage and proportioning resistors which provides feedback around a high gain, stable d-c amplifier. An understanding of this control mechanism can be reached in a variety of ways, using several distinct approaches, some of which are explored in this and other chapters.

A useful way of approaching the bridge is through the programming concept. In this manner, the closed-loop can be examined as a way of exercising command control over the output voltage of a regulator. Consider the simple voltage comparison circuit shown by Fig. 2.1. The voltmeter reads the difference between the two "batteries" $E_o - E_f$. If the two voltages are made very nearly equal, then

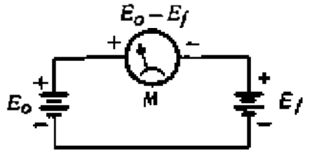


Fig. 2.1 Voltage comparator.

the meter may be switched to a more sensitive scale in order to obtain significant deflection of the indicator. If the differential $E_o - E_f$ is reduced still further, a point will be reached where it becomes necessary to amplify the difference in order to detect it. The amplifier in Fig. 2.2 produces a voltage equal to $A(E_o - E_f)$ where A is the gain of the d-c amplifier.

If, somehow, the output of the d-c amplifier could be made to control one of the two batteries, say E_o, so that E_o decreased when $A(E_o - E_f)$ increased, then a closed-loop feedback system will have

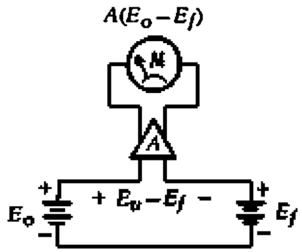


Fig. 2.2 Amplified voltage comparator.

been established, such that $E_o - E_f$ was minimized. One way to accomplish this in a practical sense is to insert a lossy element (like a resistor) in series with a source E_u, drawing some current I, as shown in Fig. 2.3.

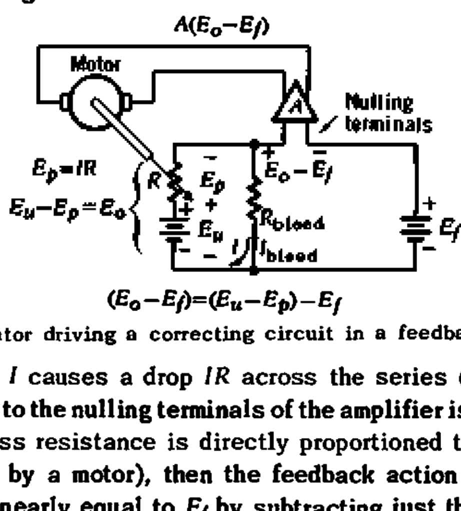


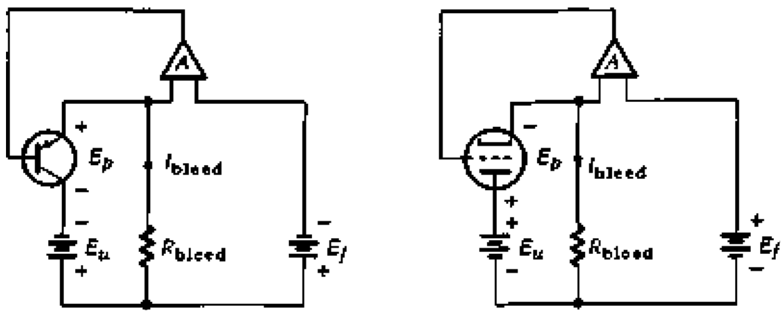
Fig. 2.3 Comparator driving a correcting circuit in a feedback arrangement.

The current I causes a drop IR across the series (pass) element, so that the input to the nulling terminals of the amplifier is $(E_u - IR) - E_f$. If the series pass resistance is directly proportioned to the amplifier output (perhaps by a motor), then the feedback action will cause E_o to become very nearly equal to E_f by subtracting just the right amount (IR) from E_u. The voltage drop across the passing resistance is called *the pass voltage* E_p.

The difference $E_u - E_p$ constitutes the output E_o of this rudimentary control system. It is clear that the amplifier's control over the pass resistance will cause E_o to equal E_f, less only the small differential $E_o - E_f$ necessary to actuate the amplifier. If the amplifier gain A is high, the difference is small, and E_o can be made to follow E_f quite closely.

2.2 PASS ELEMENTS

In a power supply, active elements, such as vacuum tubes and transistors, are used for the pass elements. The conduction of such elements (and thus their equivalent circuit resistance) can readily be controlled by adjusting their control element bias (Fig. 2.4).



(a) PNP transistor pass element (b) Vacuum tube pass element

Fig. 2.4

E_u is the power supply's unregulated, or raw, d-c source of power. The tube or transistor is the series regulator driven by an amplifier whose output is derived by nulling against E_f, the reference. The difference $E_u - E_p = E_o$ is the output.

With either tube or transistor, control over the output E_o is exercised by causing the voltage drop across the pass element E_p to vary so that $E_u - E_p = E_o$, which is approximately equal to E_f.

If we redraw these circuits somewhat more symbolically (Fig. 2.5)

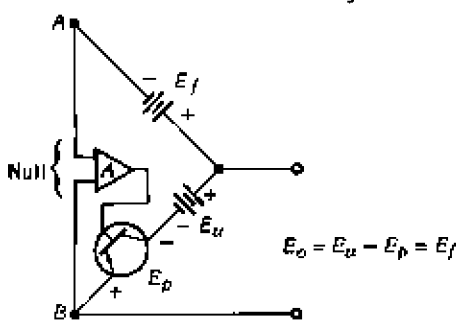


Fig. 2.5

using a transistor (although the vacuum tube is equally appropriate), the feedback action becomes more apparent. Here it is assumed that the pass element is properly biased and supplied with auxiliary sources, so that the voltage E_p can be developed without the need to show a bleeder current I. Since the dynamics of the corrective feedback are such as to make E_p independent of I, these assumptions introduce no significant error.

2.3 CIRCUIT EQUATIONS

If the amplifier gain A is assumed to include the amplification of the pass transistor, then the complete loop equation is

$$[(E_u - E_p) - E_f]A = E_p$$
$$AE_u - AE_f = E_p + AE_p$$
$$A(E_u - E_f) = E_p(1 + A)$$
$$E_p = \frac{A(E_u - E_f)}{1+A}$$

This will be recognized as the familiar feedback expression. If A is large, $A/(1+A)$ is close to unity, $E_p = E_u - E_f$. Thus, $E_u - E_p$ is $E_u - (E_u - E_f)$ or $E_u - E_p = E_f$, and E_o, which is $E_u - E_p$, equals E_f.

To program this circuit, a variable voltage source could be substituted for E_f. The nulling action of the amplifier causes the output E_o to track E_f exactly. Ideally, the nulling amplifier has a very high input impedance, so that no current flows through its terminals. On the other hand, the source E_u and its pass transistor can be made to deliver large amounts of power. Thus, the circuit functions as an impedance transformer, transforming a high input impedance at E_f to an identical voltage (E_o) at very low impedance. Since they have one terminal in common, any voltage $E_{in} = E_f$ is simply repeated as E_o without any change in reference point or polarity (Fig. 2.6).

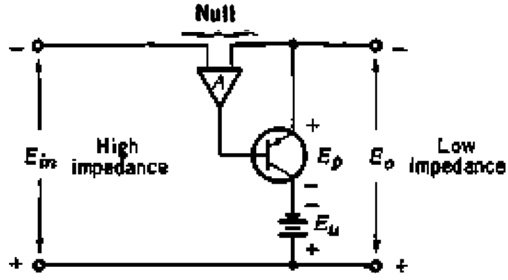


Fig. 2.6

The circuit has not yet been explicitly described as a regulator, but it should be apparent that no matter how E_u might vary, E_p will always oppose it in such a way that their sum equals E_f. If E_f is a constant, then E_o is regulated against variations in E_u (such as might be caused by line voltage changes or loading).

2.4 PRACTICAL PROGRAMMING MEANS

The problem now is to devise a *practical* programming method, a way of causing effective variation in E_f. One way to do this would be to take advantage of the fact that the input impedance of the circuit in Fig. 2.5 is very high and would not significantly load a potentiometer divider (Fig. 2.7).

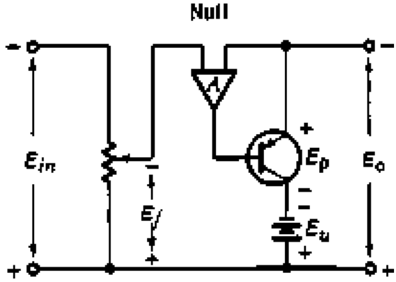
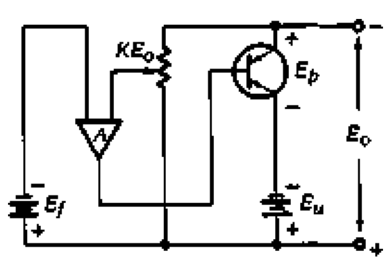


Fig. 2.7 Potentiometer adjustment of the reference voltage.

This is a perfectly legitimate way of programming; it is, in fact, a circuit of some value in the system application of power supplies.

The major disadvantage, and it is a serious one, is that the input voltage E_{in} must be equal to, or larger than the output voltage E_o. If the circuit were being used for a power supply, the reference would have to be larger than the design output voltage.

Another possibility, one which gets around the problem of needing a large reference, is to use a divider across the output (Fig. 2.8).



K is the potentiometer divider ratio.

Fig. 2.8 Potentiometer sampling of the output voltage.

This circuit fails of practicality because the amplifier has lost its common connection with the emitter of the transistor that it is driving, and is no longer common to the output. In addition, the amplifier gain is divided by the potentiometer factor, *K*.

Another method by which the output may be controlled is to substitute a resistance and parallel current source for the command. With an adjustable resistance, the voltage E_f can be generated as shown in Fig. 2.9.

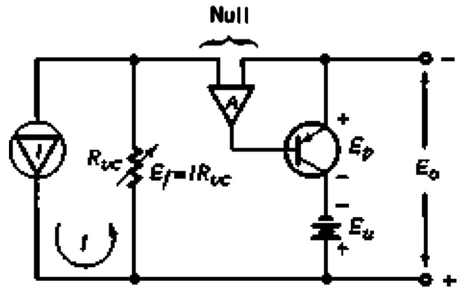


Fig. 2.9 Current source method of creating an adjustable reference.

The passage of I through R_{vc} generates a voltage drop (IR_{vc}) which then programs the circuit as though it were a voltage source E_f. This, of course, is only possible because the amplifier input draws little or no current, minimizing the loading on R_{vc}. The advantage to this circuit is that E_f (and thus E_o) can be varied simply by varying R_{vc}. R_{vc} then constitutes a voltage control. The voltage E_f could also be controlled by changing I.

2.5 SIMULATING A CURRENT SOURCE

The problem now is to simulate the current source. Recall that a series resistance circuit (Fig. 2.10) functions as a

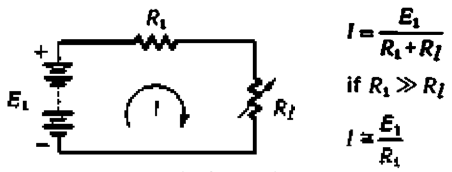


Fig. 2.10 Series resistance method of simulating a current source.

constant current generator if $R_1 \gg R_l$. In usual practice, this means very high values for R_1, and correspondingly high voltages needed for reasonably constant current. On the other hand, if R_1 can be made to approach zero by interposing an impedance transformer (Fig. 2.11), the current I becomes equal to E_1/R_1 and is constant so long as E_1 and R_1 are unchanged.

Referring back to the comparison circuit (Fig. 2.5), we recall that the amplifier's input terminals support a null (zero volts). This is

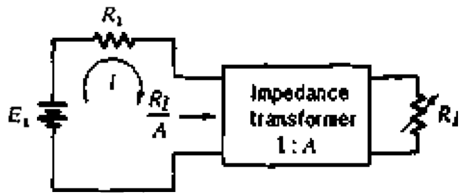
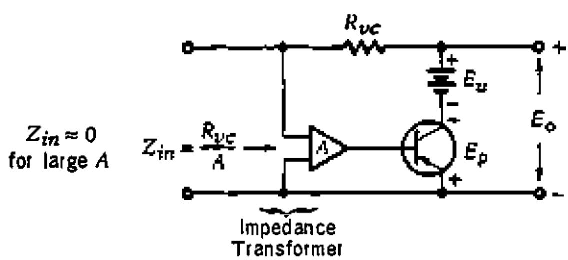


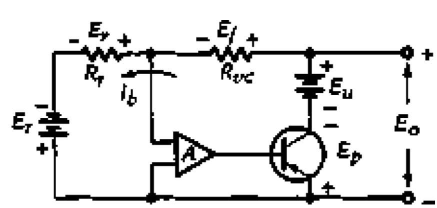
Fig. 2.11 For A large, R_l/A is small compared to R_1, when R_l/A approaches zero, $I = E_1/R_1$ and is constant despite variations in R_l.

electrically equivalent to a short circuit of zero ohms, *except that no current actually flows* through the amplifier's terminals. The null is established by the bucking of one current against the other (Figs. 2.12a - b).

If our series resistance circuit is connected to the zero ohm (virtual ground) established by the amplifier's null, it becomes a current generator of $I_b = E_1/R_1$. I_b cannot go through the amplifier's terminals, so it goes through the resistance R_{vc}. Thus, the nulling circuit itself constitutes the impedance transformer that permits a simple series resistance circuit to appear as a nearly ideal current source.



(a)



(b) $I_b = E_r/R_r$, constant irrespective of R_{vc}. I_b passing through R_{vc} generates E_f, the secondary control voltage which governs E_o.

Fig. 2.12

2.6 THE KEPCO BRIDGE CIRCUIT

This arrangement of circuit elements (Figs. 2.12b and 2.13) is the Kepco bridge circuit* in which I_b equals the bridge current generated by E_r/R_r (redesignating E_1 and R_1 as E_r and R_r, respectively, to denote their function as the reference generator of the bridge). As I_b flows through the voltage control resistance R_{vc}, the control voltage ($E_f = I_b R_{vc}$) is generated. The nulling action of the amplifier causes a voltage drop across the pass element, so that $E_o = |I_b R_{vc}|$, establishing the critical null.

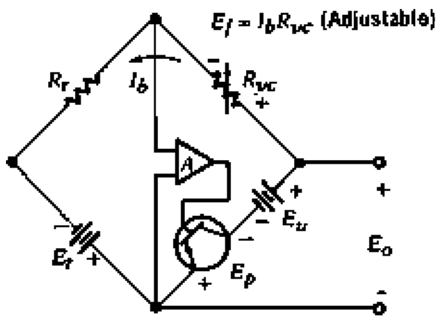


Fig. 2.13 The Kepco bridge.

In a power supply, the voltage E_r may be generated by a cascade zener circuit, perhaps driven by another regulator. In normal operation, the bridge current is constant, thus permitting the designer to choose the most nearly optimum operating point for his reference zener. Generally, the reference resistor R_r is chosen to develop a convenient bridge current. Any reasonable value will do, but for programmable power supplies, it is desirable to use the whole numbers, like 1 ma or 10 ma.

Because the nominal zener voltage can have wide tolerance limits (a 6.2V diode may actually be anywhere from 5.9 to 6.5 volts), some provision is usually made for adjusting a portion of R_r so that a precise bridge current I_b can be set.

The output voltage of the complete Kepco bridge is programmable by means of three possible inputs. Each exercises a proportional control over the output according to the formula $E_o = E_r \times 1/R_r \times R_{vc}$. E_r and R_{vc} have direct control; R_r exercises inversely proportional control (Fig. 2.14).

Normally, the reference elements E_r and R_r are kept constant while the output is adjusted using R_{vc}, the output voltage $E_o = |I_b R_{vc}|$,

 *Patent No. 3,028,538.

giving rise to a linear, direct form of programming which lends itself conveniently to precision control using a rheostat for programming. Because the current (I_b) through the control resistance (R_{vc}) is small, low power, precision controls are readily employed. The technique also lends itself to decade programming by means of a number of remote control and motor driven programs.

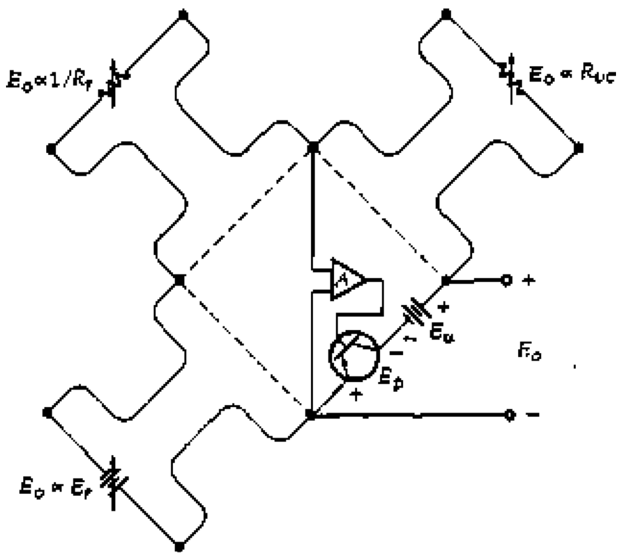


Fig. 2.14 Remote programs separately or together.

The output E_o is equally sensitive to variations in I_b, leading to several alternate programming techniques. Programming might be accomplished by variations in R_r as a means of controlling I_b. Normally, the variable portion of R_r is used only as a calibration control to set I_b to some precisely desired value. It can be expanded, however, to control E_o over narrow limits to superimpose a pattern over another program at R_{vc}. As a voltage control, R_r has the advantage that the voltage across its terminals is equal to E_r (approximately 6 volts). R_{vc}, on the other hand, always carries a voltage drop equal to E_o, which in high voltage power supplies may sometimes be inconvenient. Another advantage is that if the connections to an external R_r programmer are accidently broken, E_o collapses to zero (I_b becomes zero). While, by comparison, should the wires connecting R_{vc} to the bridge break, E_o approaches the maximum unregulated output, E_u.

On the other hand, control by means of R_r is inherently nonlinear since it is a reciprocal function. Further, the control must be exercised by making R_r larger and I_b smaller, in order to prevent the circuit from pulling an excessively large current from the zener reference diode (a current larger than the design value would cause the zener current to decrease and perhaps go out of the zener region). Even with this restriction, the current through the zener varies with R_r by an amount equal to the magnitude of the original bridge current. In a 10 ma bridge this can cause quite a large change in zener current – from the normal 7.5 ma design center to 17.5 ma. Naturally, this causes some change in the zener voltage which adds to the inherent errors of this kind of program. There may also be some longer term effects occasioned by the heating which results from the slightly increased dissipation in the reference diode.

The third control possibility involves substituting for E_r itself a voltage, E_i (E input), which serves to voltage program the power supply's E_o. The source used for E_i will probably have a voltage different from E_r, so that in order to maintain I_b, an appropriate coupling resistor R_1, replacing R_r, is used. In effect, the supply is programmed by replacing the entire command side of the bridge by an external source and coupling resistor. If it is desired to retain the built-in voltage control, it will usually be desirable to maintain the design value of I_b. Thus, the coupling resistor R_1 is chosen so that $R_1 = E_i$ (maximum)/I_b design. The input impedance seen by the source is R_1 and, of course, it must be capable of delivering I_b – which means that it must be able to sustain a load of R_1 ohms (Fig. 2.15).

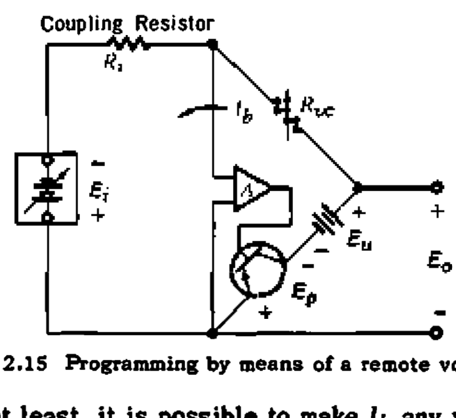


Fig. 2.15 Programming by means of a remote voltage.

In theory, at least, it is possible to make I_b any value whatever, so long as R_{vc} is chosen appropriately. This would seem to invite large value input resistances to reduce circuit loading. Unfortunately, stray leakage currents limit the effective minimum I_b (maximum R_1). There are some leakage cancellation techniques treated in Chap. 3 which are useful in this regard. However, if current loading is a real problem, one excellent solution is to employ a small intermediate power supply as an impedance transformer (Fig. 2.16).

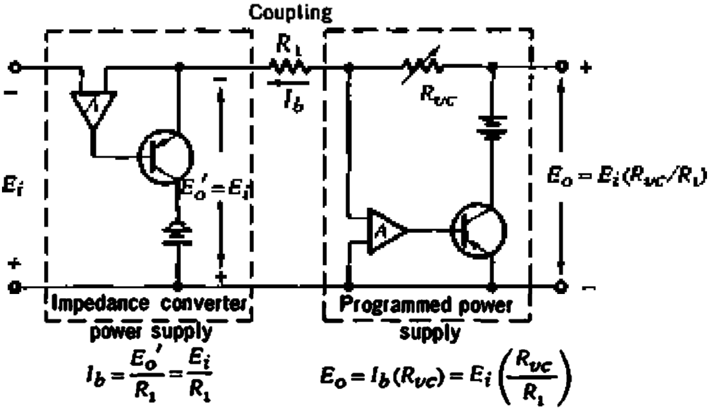


Fig. 2.16 Input impedance approaches infinity. Current drawn from E_i (input) is the leakage of the impedance converter's amplifier which can be cancelled by appropriate trimming.

Another restriction on the programming source E_i is that it must always be clamped, boosted or summed in such a way that it is a unidirectional source. One simple way is to use a d-c source in series with the program, so that the input E_i equals their sum. Another way is to use another input and coupling resistor to inject a d-c component of I_b. The original E_r and R_r may often be used for this. The

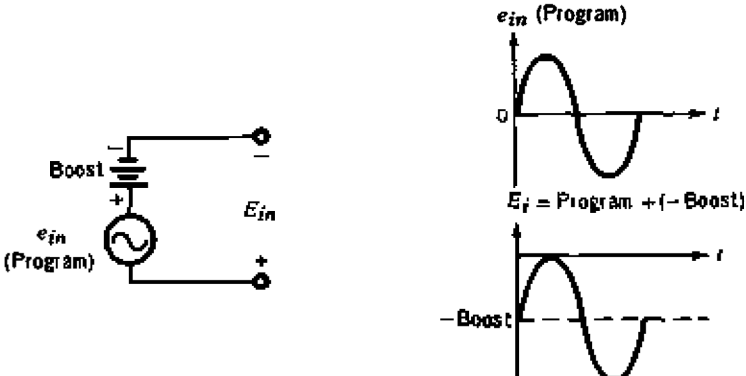


Fig. 2.17a Input signal constraints.

null at the amplifier's input terminals effectively decouples such circuits so that the total bridge current $I_b = E_i/R_1 + E_r/R_r$ + etc. The object is to make sure that the direction of I_b does not reverse for even a portion of the input cycle. If it did, it would be rectified by the unipolar nature of the pass elements.

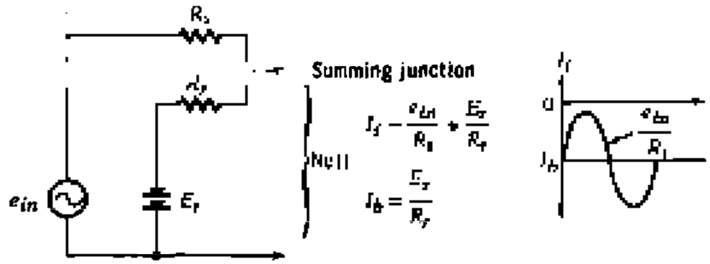


Fig. 2.17b Method of biasing the input.

3 OPERATIONAL ANALYSIS AND SYMBOLOGY

3.1 OPERATIONAL AMPLIFIERS

The 4-element Kepco bridge circuit has been developed in a way which leads naturally to an operational amplifier analogy. By convention, the bridge is usually drawn in the traditional diamond figure (Fig. 3.1).

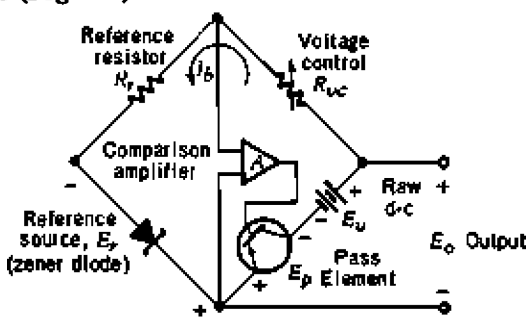


Fig. 3.1 Simplified power supply diagram drawn as a 4 arm bridge. Note: the diagram, like most in this handbook, is drawn for a PNP transistor pass element. For NPN pass transistors, or hybrid power supplies using a vacuum tube pass element, the polarities are reversed.

In order to see the analogy more clearly, it is possible to rearrange the bridge without actually changing the way the elements are connected (Fig. 3.2).

By substituting a symbolic input E_{in} for the zener reference E_r

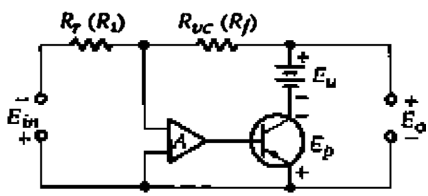


Fig. 3.2 $E_o = E_{in}(R_f/R_1)$; $R_f/R_1 = G$, **operational gain.**

at the input terminals, we have a simple voltage amplifier, whose output voltage E_o is just E_{in} multiplied by the factor G. Since G is R_{VC}/R_r, G is greater than unity whenever $R_{VC} > R_r$ and vice versa. The power supply, like an operational amplifier, is simply a high gain d-c amplifier in which degenerative feedback is arranged so that the operational gain is the ratio of two resistors.

Originally, the term "operational amplifier" was used in connection with analog computers to describe amplifiers that performed various mathematical *operations*. By the proper selection of feedback components, operational amplifier circuits could be used to add, subtract, average, integrate and differentiate. By properly altering the symbology, it is entirely possible to analyze the bridge controlled power supply in operational terms, thus opening the way to an easy understanding of power supply behavior, the effects of various control circuit elements, and the many possible control and feedback applications.

Actually, such a power supply combines the function of a high gain voltage amplifier with a high powered booster (the pass element, and its raw d-c supply). For purposes of analysis, it will sometimes be convenient to show these two parts separately, as in Fig. 3.2, or they can be combined into a single symbol when this facilitates the understanding of overall circuit performance.

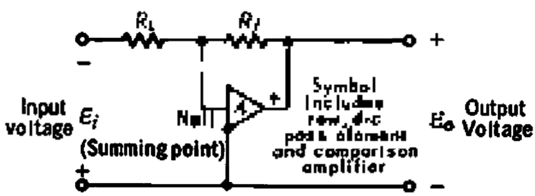


Fig. 3.3 Single ended, unipolar, inverting d-c amplifier with operational gain $G = R_f/R_1$.

By showing the polarities of E_i and E_o as in Fig. 3.3, the closed-loop gain G can be regarded as a positive number.

$$E_o = E_i \frac{R_f}{R_1} \text{ where } G = \frac{R_f}{R_1}, \qquad \begin{matrix} R_{VC} \to R_f \\ R_r \to R_1 \end{matrix}$$

ϵ is the error voltage, which is multiplied by A, the open-loop gain, to produce E_o. If A is sufficiently large, ϵ approaches zero giving rise to a null or virtual ground across the amplifier's input.

The "operational gain" $G = R_f/R_1$ is very nearly independent of A (provided $A \gg G$). The loop gain is the gain difference (in db) between open- and closed-loop gain, $(A - G)$. Actually, loop gain =

open-loop gain/closed-loop gain and is a measure of the feedback. As a power supply, it is the loop gain that provides the regulating action. Just as local feedback around a transistor can reduce circuit sensitivity to parameter changes and environmental conditions, feedback around the operational amplifier/power supply reduces sensitivity to open-loop parameter changes, load, line voltage, temperature and time. In many Kepco power supplies, the open-loop gain A ranges from 10^5 - 10^6 (100-120 db), so that if a minimum of 40 db feedback is to be retained (loop gain = 100,) the operational gain can be as much as 1,000-10,000. In practice, operational gain G will be in the range 0.1 to 1,000.

The highest operational gain used presently in Kepco power supplies is 400 (for Model ABC 2500M); the lowest is 0.3 (for Models CK 2-8M and ABC 2-1M). These figures are, of course, based upon the use of a 6V reference zener diode as the *input* for each.

The terms associated with power supply *programming* can be easily translated into d-c (operational) amplifier terminology. Both resistance and inverse resistance (conductance) programming are seen to be simply ways of varying the operational gain $G = R_f/R_1$ with a constant input $E_i = E_r$. This form of programming amounts to having an amplifier with a constant input and variable gain. On the other hand, the process of programming by means of a variable external voltage (voltage drive) is a matter of having a fixed gain amplifier with a variable input voltage.

The input impedance of the d-c (operational) amplifier is always equal to the input resistor R_1. Since the summing point is a virtual ground, the input impedance of the comparison circuit is zero and the resistor R_1 is the total input impedance. The output impedance is very low, being the open-loop impedance of the output circuit, divided by the loop gain. The input impedance of the amplifier can, in theory, be set to any desired value simply by selecting R_1. Provided the ratio R_f/R_1 is maintained constant, the gain is unaffected by changes made simultaneously in R_f and R_1. In practice, the maximum input resistance will depend on the amount of tolerable noise pick-up. Tens of megohms can be used for R_1 in special integrator circuits where the noise is swamped by an exceptionally long time constant. The minimum input resistance will depend on how much current the "input" source or reference is able to deliver. Two values common for power supplies are 100Ω/V and 1,000Ω/V, which correspond to 10 ma and 1 ma, respectively.

In a sense, the operational circuit is a current drive device, since its input impedance (at the null junction) is nearly zero. It does not matter one whit what the input voltage may be, so long as it can deliver the requisite current through whatever input resistor may be selected. Usually this current is picked first by determining a value for R_1 based on the expected input voltage. R_f is then selected to provide the desired gain for the wanted E_o.

It is important to realize that substantial differences exist between the operational amplifier intended for computational functions, and the d-c power supplies that we are describing. For one thing, the power supplies are unipolar, capable of varying only in one direction from zero. Another notable difference is that the power supply contains substantial filtering – to enable it to deliver high currents at low ripple – and to shape the phase-gain characteristic in such a way as to make the supply immune to the possible range of load phase angles. This filtering has the effect of producing a very low-frequency breakpoint, sharply limiting the frequency response of the device. For some power supplies, interchangeable "high speed" regulators are available which provide increased bandwidth by reducing the filtering. In other supplies, the regulator is easily changed for high speed operation (see Chap. 9).

On the other hand, the d-c power supply is self-contained; it requires no external d-c sources, it is capable of relatively huge amounts of output power, and is fully protected against all conditions of loading, such as short circuits. There is a growing class of circuit applications which can use these peculiar power supply/operational characteristics to considerable advantage.

3.2 OPERATIONAL CIRCUIT APPLICATIONS

Some fairly straightforward applications for the operationally programmed power supply include:

(A) Impedance Transformer: Voltage follower (Fig. 3.4). The impedance transformer can be used in a number of ways:

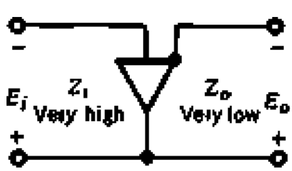


Fig. 3.4 D-c impedance transformer configuration.

1) To eliminate potentiometer loading (Fig. 3.5).

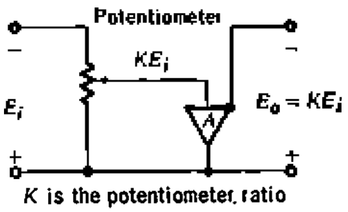
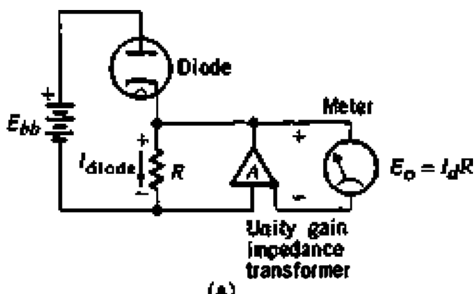


Fig. 3.5 Potentiometer follower, eliminates loading.

2) As a current sensing device, measuring, for example, the cathode current of a thermionic diode (Fig. 3.6a-b).



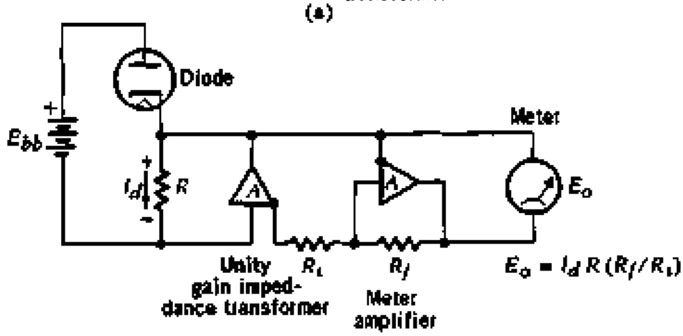
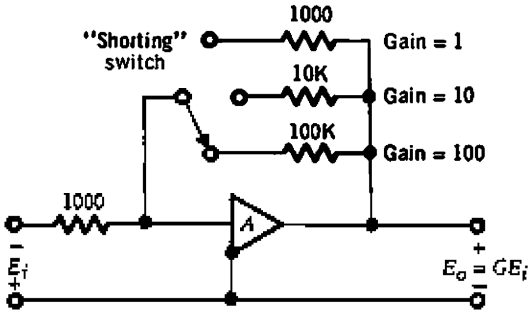
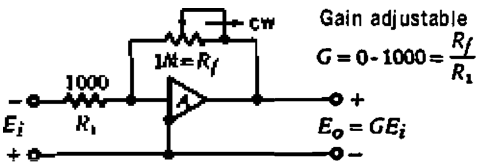


Fig. 3.6b Current sensing circuits with and without amplification.

(B) Voltage Amplifier: Figure 3.7a-b. Varying R_f either continuously or in steps produces a stable, adjustable gain d-c amplifier suitable for a host of laboratory applications, including meter amplifiers, photo-electric cell amplification, recorder drive amplifier, null detector, and potentiometrics. The voltage amplifier can be combined with the impedance transformer to produce a virtually infinite, potentiometric-like input (Fig. 3.8).



(a) Step selectable gain.



(b) Continuously adjustable gain.

Fig. 3.7

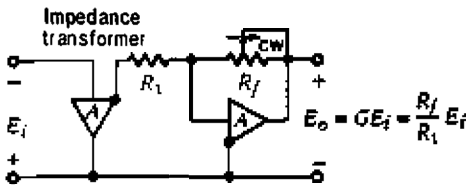


Fig. 3.8 Adjustable gain d-c amplifier with high input impedance.

(C) Voltage Adder (Scaling summer): Figure 3.9.

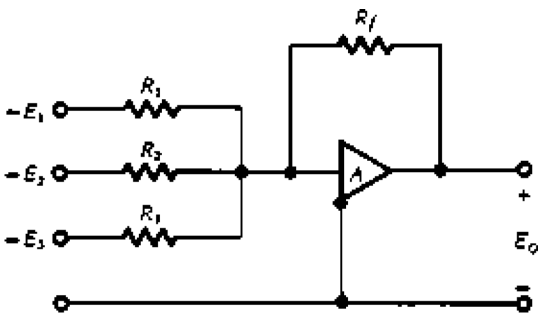


Fig. 3.9 Summing amplifier, $E_O = R_f(E_1/R_1 + E_2/R_2 + E_3/R_3 + \ldots)$.

(D) Integrator or Ramp Generator: Figure 3.10.

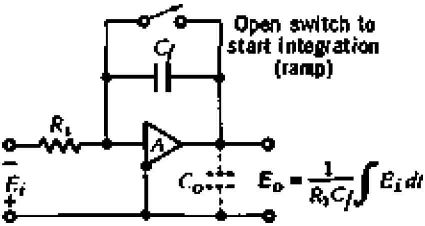
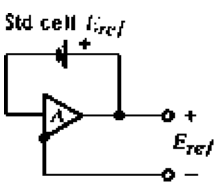


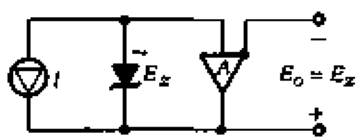
Fig. 3.10 Voltage input current integrator (ramp generator). Note: the output capacitor has not been diagrammed explicitly, but must be accounted for in all time dependent operations. For integration, C_O, shown dashed, should be removed (see Chap. 9), or make $C_f R_1 \gg C_O R_O$ where R_O is output impedance (normally low). The HS (high speed) PAX regulators, and the OPS (operational power supply) contain no output capacitor C_O, and so make very good integrating amplifiers.

(E) Voltage Reference:

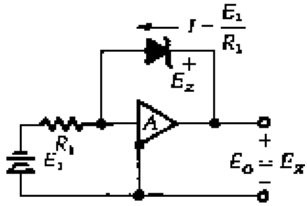
1) A *fixed reference* may be derived from a standard cell, a current fed zener diode, or a mercury battery. The impedance transformation provides a highly stable reference output capable of driving low-impedance measuring circuits directly (Fig. 3.11a-c).



(a) Standard cell buffer. Prevents damage to standard cells from loading. Output terminals are at low impedance and will deliver substantial power with standard cell accuracy.



(b) Fixed reference current driven zener diode.



(c) Fixed reference using the amplifier to regulate the zener current.

Fig. 3.11

2) *Adjustable voltage reference* (power supply), refer to Fig. 3.12. The variable reference can be adjusted very precisely if accurate resistors and a stable voltage reference are available. A simple method for zeroing and calibrating is given in Sect. 3.3(C).

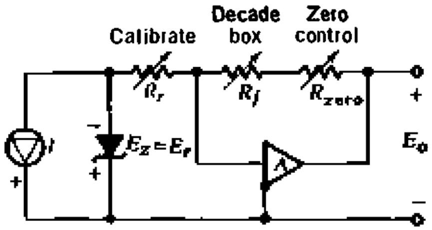


Fig. 3.12 Adjustable reference voltage source. $E_o = E_r\left(\frac{R_f + R_{zero}}{R_r}\right)$. **Set zero and calibrate according to the procedure given in Sect. 3.3(C).**

(F) Small Current Regulator: A practical manifestation of this circuit uses two power supplies connected in tandem (Fig. 3.13).

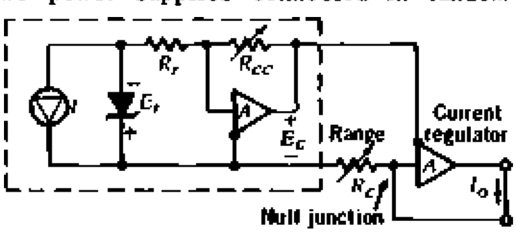
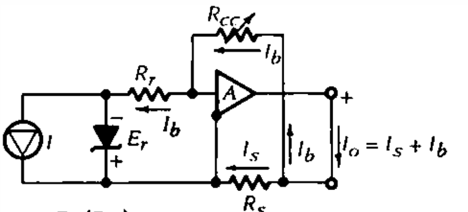


Fig. 3.13 Adjustable small current source. Standard power supply drawn operationally, functions as the driver. Its voltage control, R_{CC}, serves as a current control, adjusting E_C. $I_o = E_C/R_C$. R_C functions as a current range control.

Note that since I_o is a direct linear function of the driver supply's feedback resistor (voltage control), that control is considered a current control for the over-all circuit and is labeled R_{CC}. R_C can serve as a range switch if desired. When connected as shown, excellent small current regulation will be obtained, equal to the "current regulator's" abilities as a normal voltage regulator and limited only by the stability of the driver source. A minimum current, on the order of 5-15 microamperes, flows into the amplifier connection at the null junction of the "current regulator" supply. For control over exceedingly small currents, a cancellation technique is useful to eliminate the effect of this leakage. A method for cancelling the leakage is described in Sect. 3.3(B).

The maximum current is limited only by the ability of both supplies to generate the desired current. However, since this current must flow through R_c, the heating in this resistor will confine this current to small current application.

(G) Large Current (sampling) Regulator: Figure 3.14. This is the circuit used by almost all current regulating conversions of voltage regulators. The current $I_b = E_r/R_r$ is the limiting minimum current that can be regulated since when R_{cc}, the *current control,* is made zero, only the first term of equation (1) disappears, leaving I_b in the load. For load currents which are large relative to I_b, this limitation is not usually serious, particularly since the offset [see Sect. 3.3(B)] often permits the output to be cut off. In special situations, the minimum current restriction can be circumvented by substituting a variable voltage source (supplemental power supply) for the resistor R_{cc}. (See Fig. 3.14). This obviates the need for R_r and E_r, and does away with I_b.



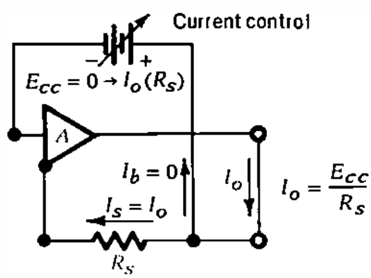
1. $I_o = E_r\left(\frac{R_{cc}}{R_r}\right)\left(\frac{1}{R_s}\right) + I_b$

2. $I_o = E_r\left(\frac{R_{cc}}{R_r}\right)\left(\frac{1}{R_s}\right) + \frac{E_r}{R_r}$

3. $I_o = \frac{E_r}{R_r}\left(1 + \frac{R_{cc}}{R_s}\right)$

$$I_s = \frac{E_r}{R_r}\left(\frac{R_{cc}}{R_s}\right)$$

(a) Large current regulator.



(b) Alternate current regulator configuration.

Fig. 3.14

3.3 PRACTICAL CIRCUIT CONSIDERATIONS

Irrespective of the analytical approach, bridge concept or operational analogy, employed, certain nonideal characteristics of the nulling circuit should be understood and taken into account.

(A) Offset: In practice, an offset voltage ranging from near zero to several hundred millivolts will be measured between the null junction and the power supply/amplifier's common terminal. This voltage does not enter into the nulling process, but can be accounted for by assuming a small battery in the common terminal, as shown in Fig. 3.15.

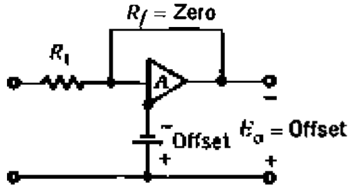


Fig. 3.15

Taking the Kirchoff voltage loop around the feedback and output circuits, with R_f set equal to zero, it will be seen that the offset causes a small reverse potential across the power supply's output terminals. In many power supplies, a small offset is deliberately introduced in order to cause a negatively polarized output when the feedback control R_f is zero. This feature permits the user to insert a zeroing control as shown in Fig. 3.16 to adjust the output of the

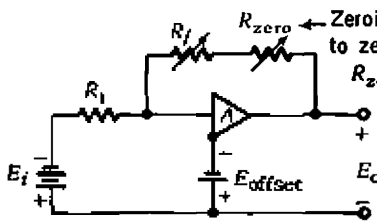


Fig. 3.16

power supply exactly to zero. By having permitted the power supply to go past zero, the offset allows the terminals to be adjusted to a precise zero value with whatever tolerance may be desired. Considering the wire resistance and switch contact resistance that in all likelihood will exist in any practical programming circuit, it is evident that without a deliberate offset, true zero would be impossible (Fig. 3.16-17).

The zeroing control is a small variable resistance placed in series with R_f to program just enough positive output as to just cancel the offset. Plotting output versus the resistance R_f, the characteristic of Fig. 3.17 is obtained. The straight line characteristic is made to pass through the origin of the axes by adjusting R_{zero}. The

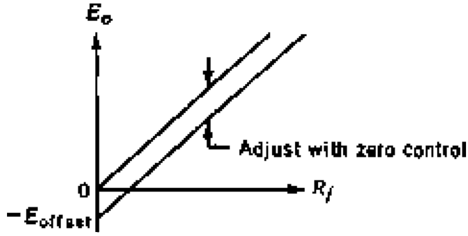


Fig. 3.17 E_O versus R_f, showing effect of the offset voltage.

zero control is required whenever accurate power supply control is wanted. The adjustable reference circuit of Part (e), Fig. 3.12, is an example of an application requiring precise zeroing. Generally, a high resolution multiturn potentiometer is used for a zero control. For 1,000 ohm/V circuits, use a 500Ω control, for 100Ω/V control ratios, a 50Ω control is suitable. The zeroing procedure requires that the programming resistors be shorted, or otherwise set to zero. Place a sensitive millivoltmeter across the output terminals of the power supply and adjust for a zero reading. Offset will vary slightly as a function of time and temperature. However, its contribution to the overall power supply drift and temperature coefficient is small.

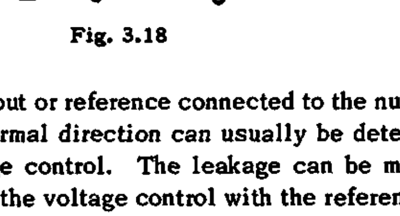


Fig. 3.18

(B) Leakage: With no input or reference connected to the null junction, a small current in the normal direction can usually be detected in the feedback resistor/voltage control. The leakage can be measured by using a large resistor as the voltage control with the reference resistor (input) disconnected. Divide the observed output terminal voltage by the resistor value to determine the size of the leakage current. A 1 megohm resistor may produce from 5 to 15 volts of output, indicating leakage on the order of 5 - 15 microamperes. The current is relatively fixed, and for most applications will be accommodated by adjusting the loop current calibration (the input resistor). In some applications, however, such as the small current regulator circuit (Fig. 3.13), a

special cancellation technique is used to eliminate the leakage Fig. 3.19).

Cancellation is a form of "bootstrapping" or trimming which uses an externally supplied current to just exactly oppose the leakage current (Fig. 3.19). Properly cancelled, the feedback terminals of the power supply present a near infinite input resistance. The cancellation current must flow opposite the normal loop current direction so that the cancellation source must have the opposite polarity from the source or reference normally used to drive the supply. In Kepco automatic crossover (VIX) power supplies, the zener reference for the current bridge is such a source. It delivers a +6.2 volt potential to a terminal on the supply's barrier strip (see Appendix) which may be borrowed to supply the cancellation current. A 1 megohm trimming resistor in series with this potential to the null junction of the amplifier will cause 6.2 microamperes of cancellation current to flow. By suitably adjusting this resistor, exact cancellation can be obtained. For those power supplies lacking the auxiliary reverse potential, a small battery can be used to deliver the needed microamperes of cancellation current.

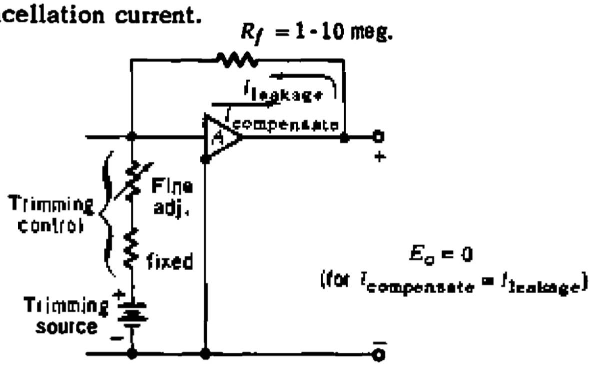


Fig. 3.19 Leakage compensation circuit.

A convenient way of determining when complete cancellation is achieved is to observe the output voltage with the normal input or driver or reference circuits disconnected. Use an adjustable or selectable resistance across the feedback (programming) terminals in place of R_{vc}. Adjust the trimming resistance for zero output while the R_{vc} resistance is increased to very large values. It is usually convenient to construct the trimming resistance in two parts: a large fixed resistor and a small variable one to give enhanced resolution. As the resistance in the programming circuit R_f is made very large, the test becomes quite sensitive. For R_f equal to 10 megohms, a 1 volt change in the output represents 10^{-7} amperes of leakage. For an even more sensitive test, R_f can be replaced by a capacitor, in which case the object is to cause the output voltage to remain constant, indicating that no charge is flowing into or out of the capacitor. Of course, a good low-leakage capacitor is necessary. Actually, in most power supplies there is a capacitor already bridging the feedback terminals which can be used for this test. If the output meter should go off scale in either direction, the feedback terminals should be shorted to restore zero initial charge on the capacitor.

When compensated, the leakage is reduced to virtually zero. If the power supply is then connected as a small current regulator (Fig. 3.13), the lower current limit with careful trimming can be made as small as 50 - 100 nanoamperes. The same compensation technique can be used to raise the open-circuit input resistance of the circuits shown in Figs. 3.4, 3.5, 3.6, 3.8 and 3.11a. By carefully trimming the compensating currents, the open-circuit input resistance can be raised to the vicinity of 10^7 - 10^8 ohms.

(C) Calibration: The slope of the programming ratio (the plot of volts output versus control ohms) depends on the accuracy of the bridge current setting. If the bridge current is exactly 1 ma, the output will vary exactly 1 volt for each 1,000 ohms of R_f. The programming ratio is described as 1,000Ω/V; a 10 ma bridge programs at 100Ω/V. Adjustment of the bridge current made by reference to some external standard constitutes calibration. The current is adjusted by means of R_r. In most power supplies, where a zener diode reference is employed, the reference potential may have any fixed value between rather wide limits (5.9 - 6.5V) typically, so the reference resistor R_r is constructed in two parts: one fixed and one variable, for calibration.

A typical calibration setup is shown in Fig. 3.20. For proper alignment, the supply should be zeroed before calibration. Using switch Position 1 with $R_{VC} = 0$, adjust R_{zero} for a zero reading on the null meter. The reference source is a mercury cell whose terminal voltage is very close to 1.35 volts. To calibrate the circuit, make $R_{VC} = 1350\Omega$ as exactly as possible, preferably by series connecting individual resistors of known exactitude, 1000Ω + 300Ω + 50Ω. Adjust R_r until the null detector reads zero when the switch is keyed to Position 2. Recheck the zero and then the calibration once again to eliminate interaction between the adjustments. A carefully aligned power supply will track precision programming resistors to an accuracy determined by the loop gain (see Sect. 1).

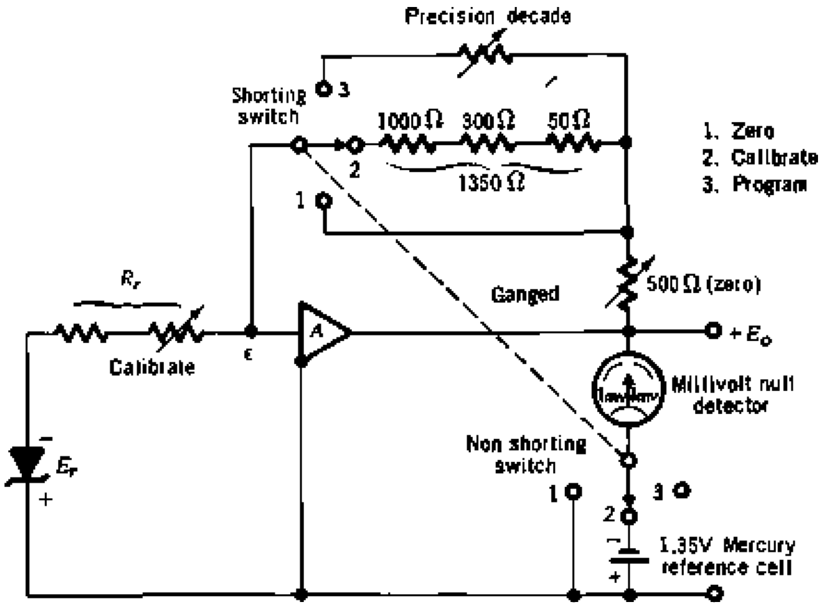


Fig. 3.20

The circuit is inherently linear since, with leakage and offset accounted for, output is solely the product of bridge current and programming resistance, less ϵ, the open-loop null error. ϵ changes slightly, of course, increasing at the rate E_o/A. Numerically, if the output is programmed over a 1000:1 ratio from, say, 0.1 to 100 volts, ϵ varies $1000/A$, which for $A = 10^5$ is only 10^{-2}, or 1%. For $E_o = 100$ volts and $A = 10^5$, ϵ max is 1 millivolt, so $\Delta\epsilon$ is 1% of 1 millivolt or 0.01 millivolts!

The output is very good, straight line reproduction of the programming resistance, capable of following the program of the most accurate resistance decades.

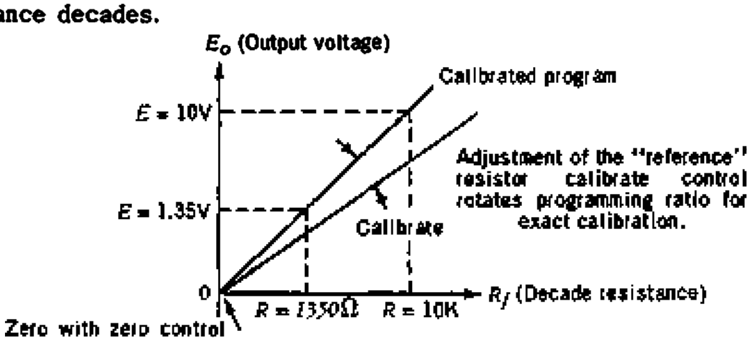


Fig. 3.21 Alignment for 1000Ω/V power supply.

4 SHORT CIRCUIT PROTECTION AND CURRENT REGULATION

4.1 CURRENT LIMITING

All power supplies should be protected in some way from the possibility of damage through overload. In the simplest circuits, protection may take the form of a fusible element, or a thermal or magnetic circuit breaker. In more sophisticated programmable power supplies, the pass element dissipation is usually such as to permit unrestricted operation at any voltage from zero to the maximum output voltage, while the maximum rated current is delivered to a load. This capability suggests a possible overload protection against short circuits. If the power supply could be made to reduce its output voltage when overloaded, it ought to be possible to limit the short circuit current to some safe maximum. Such operation is, of course, a crude form of current regulation. A typical limiter circuit might appear as in Fig. 4.1.

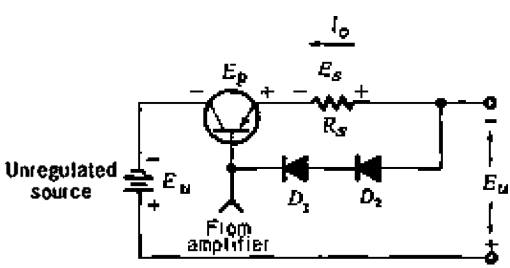


Fig. 4.1 Elementary current limiter.

The diodes D_1 D_2 are nonconducting until the voltage drop across the emitter sensing resistor R_S exceeds their forward conduction potential (approximately .25 volts each for germanium diodes, 0.7 each for silicon diodes). Should the output of a power supply containing this circuit be short circuited, the current I_o would only increase up to a limiting point determined by the feedback through D_1 D_2, tending to cut off the pass transistor; E_p increases until, for a short circuit at the output, $E_p + E_s = E_u$. The maximum current I_o obviously depends on the size of the sensing resistor R_s. A large resistor means a smaller short-circuit current limit; and vice versa, a small resistor means a larger short-circuit current limit. This kind of circuit, or more complex forms of it, is widely used as a current limiter in transistorized and some hybrid regulated power supplies. The Kepco ABC, PAX and SC design groups are examples of power supplies employing a version of this current limiter. A typical E, I plot for such a power supply might appear as in Fig. 4.2.

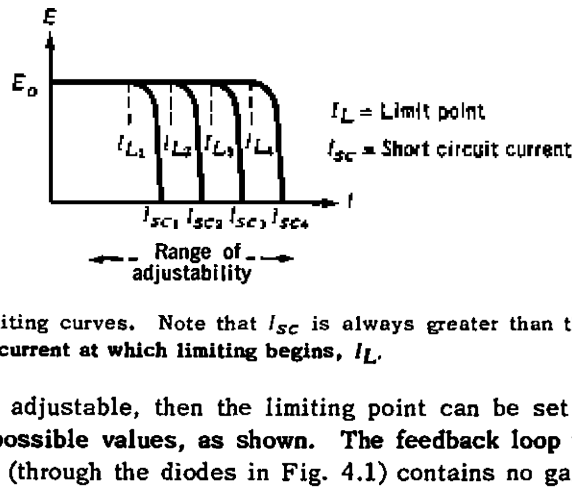


Fig. 4.2 Current limiting curves. Note that I_{SC} is always greater than the current at which limiting begins, I_L.

If R_s is made adjustable, then the limiting point can be set to cover a range of possible values, as shown. The feedback loop for the current limiter (through the diodes in Fig. 4.1) contains no gain, and so the steepness of the limit lines is not very great. Typically, there will be a slow roll-off of voltage regulation performance as current is increased, and the voltage drop across R_s begins to cause some conduction through the diodes. The first symptom is usually increased output ripple as the drive to the pass element is overcome by the local feedback. As the load is increased to a full short circuit, the output current rises only a small additional amount as the pass transistor tends to cut off, and its effective passing resistance increases. The difference between the initial current I_L and the short-circuit current I_{SC} (Fig. 4.2), expressed as a percentage of the maximum current, is the "current regulation" of the limiter circuit. Because of the low loop gain (less than unity), the "regulation" is not particularly spectacular in the simple limiter configuration. Typically, the slope of the current limit would correspond to 10-20% current regulation.

4.2 AUTOMATIC CROSSOVER

In the automatic crossover circuit, the simple limiter of Sec. 4.1 is given a high gain amplifier to boost its regulating ability, and the diodes are replaced by a positive action, or-gate, toggle circuit.

Consider the circuit of Fig. 4.3. Two control bridges are evident. One bridge takes its signal from across the output terminals of the power supply and controlling voltage. The second looks across a current sensing resistor, and will act to regulate current by maintaining a constant voltage across R_s. Each bridge has its own high gain error amplifier which permits an operational analysis of each bridge similar to that carried out in Chap. 3.

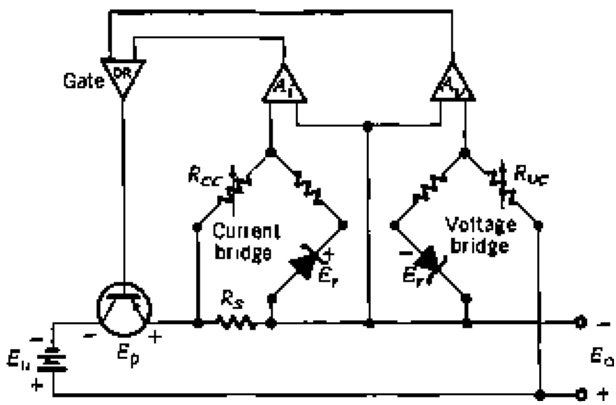


Fig. 4.3 Kepco twin bridge automatic crossover circuit.

It is interesting to observe that in order for the two error amplifiers to have a common point at the negative output terminal (for a PNP all-transistor design), the reference potential for the current bridge has the opposite polarity as compared to the reference zener in the voltage bridge. So that the two amplifiers present the same polarity at the or-gate, the current amplifier usually has one more stage of amplification and inversion. The extra amplification is needed because of the exceptionally small signals that the current bridge has to work with.

The or-gate consists of a pair of transistors sharing a common emitter resistor. The circuit is designed to respond to only one polarity (usually negative), as shown by Fig. 4.4.

Whichever transistor sees the most negative signal immediately conducts, thereby cutting off the other transistor and blocking the

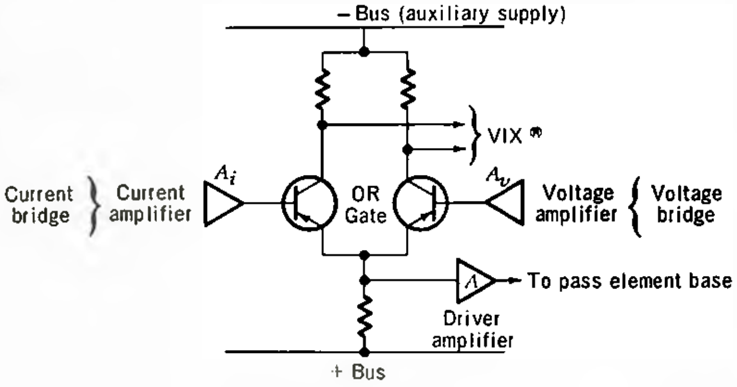


Fig. 4.4 VIX gate circuit.

path through that transistor to the pass element. Error signals which result from an output *lower* than the respective control setting are positive, and so do not get through the gate. If the output voltage is *less* than the setting, or if output current is *less* than its setting, the respective error amplifiers deliver a relatively more positive signal to the gate, and the least positive one assumes control. By such action, the dual-bridge automatic crossover supply sets *upper* limits on the parameters of voltage and current. Control is switched from one bridge to another only when the output tries to exceed one or the other limit. This action and the resultant *E, I* characteristics can be seen in Fig. 4.5a.

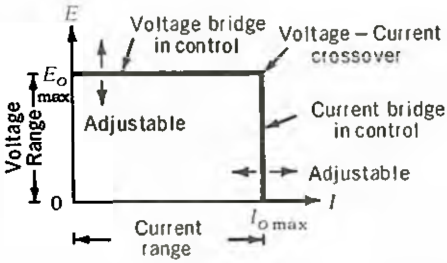


Fig. 4.5a Automatic crossover *E, I* characteristic.

The gate circuit itself incorporates a zero hysteresis, positive feedback loop, so that acceptance of either the voltage amplifier or current amplifier signal completely excludes the other. The result is a toggle-like action which makes the crossover point itself a very unstable position.

Since the voltage and current regulators both have high gain and high stability reference zener diodes, the quality of the voltage and current regulation is similar. This is the principle distinction between the automatic crossover circuit and the limiting circuit described previously. An intermediate form of current limiting-regulation with moderate gain is used in the Kepco PBX modular design.

An automatic crossover power supply is unique in that it cannot be overloaded. It operates with full control into any load resistance from infinity to zero ohms. From infinity to the crossover resistance, it behaves as a voltage regulator holding its voltage constant as the current increases. From the crossover resistance down to zero ohms, the power supply behaves as a true current regulator, holding its terminal current constant as the voltage decreases. Since the voltage and current controls are each independently adjustable throughout the full voltage and current output range, the crossover point, which is the intersection of the locus of the two control settings, can be located anywhere in the volt-ampere range of the power supply. (See Fig. 4.5b.)

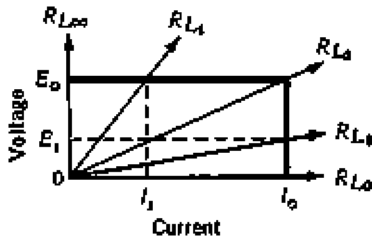


Fig. 4.5b The locus of a variable load resistance is a circle centered at the origin. An infinite resistance (open circuit) corresponds to the voltage (vertical) axis. A zero resistance (short circuit) corresponds to the current (horizontal) axis. Resistors from 0 to ∞ span the first quadrant. R_{L_1} intersects the power supply characteristic along its voltage regulated characteristic, the output of the supply is E_O and I_1. A resistor intersects at the crossover point, and is called the *crossover* resistance. The crossover resistance draws the maximum power from the power supply, $E_O I_O$. Resistor R_{L_3} intersects the characteristic at E_1, I_O along the current regulated line.

Another overload protection circuit employed in Kepco power supplies is the current cut-off circuit found in Kepco's PWR Modules. The cut-off characteristic is illustrated by Fig. 4.6. Here, positive feedback is employed to reduce the current output in proportion to the overload. Since the voltage and current diminish together, the overload power in the load decreases rapidly. At a full short circuit, the output current will be typically 5% of the maximum rating. For cer-

tain kinds of sensitive loads, this sort of power supply behavior can be very important as a safety measure.

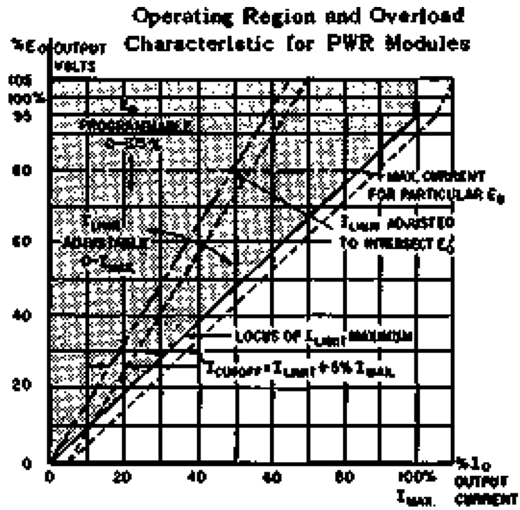
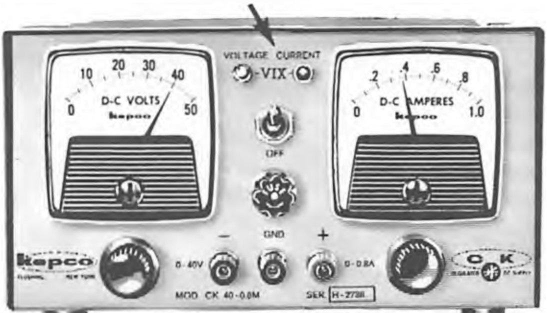


Fig. 4.6 I_{limit} is the maximum operating current for any selected cutoff current. I_{limit} is approximately equal to I_{cutoff} minus 5% I_{max}.

4.3 VOLTAGE-CURRENT CROSSOVER: INDICATION AND SIGNAL

The operation of an automatic crossover power supply can be somewhat confusing without specific knowledge about which mode the power supply may happen to be regulating. This information is particularly important in systems interconnection, where several power supplies are responding to a variable drive, and may be feeding power to a variable load. The Kepco VIX®circuit is an answer to the problem. VIX signifies: (*V*) Voltage, (*I*) Current and (*X*) Crossover; and consists of a pair of mode indicating lamps mounted on the power supply's panel. The pair of VIX lamps replace the usual pilot lamp to provide more sophisticated information than the simple on-off condition of the supply.

When the voltage lamp is lighted, the VIX signal tells us that the voltage amplifier of the power supply is in control, and the output has the low resistance characteristic of a voltage regulator. If the power supply's load or control settings causes the output to pass the crossover point, and gives control to the current amplifier, the voltage lamp extinguishes and the current lamp lights. Simultaneously, the VIX circuit causes a polarity reversal in a voltage presented to a pair of pin jacks on the rear apron of the power supply. The signal



Model CK 40-0.8M

is ±8V d-c at up to 0.8 ma (10K loading). The remote VIX signal can be used in certain applications to provide remote overload signalling, or to signal the end of a process (like a battery charge). An accessory amplified relay device (Model VIX-1) is available to convert the remote VIX signal into a heavy-duty relay closure.

4.4 CURRENT REGULATING CIRCUITS

Current is regulated indirectly by regulating the voltage drop across a series resistor, called a *sensing resistor.* The circuit used is the same, whether it is internal to the supply, as in automatic crossover units, or external. Borrowing from the operational analysis of Sect. 2.2(G), the sampling circuit can be drawn, as in Fig. 4.8 This is the standard current regulating circuit used in all power supplies. Using bridge notation, the same circuit can be redrawn as in Fig. 4.9

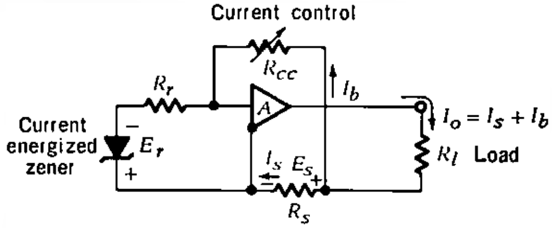


Fig. 4.8 Current regulating circuit.

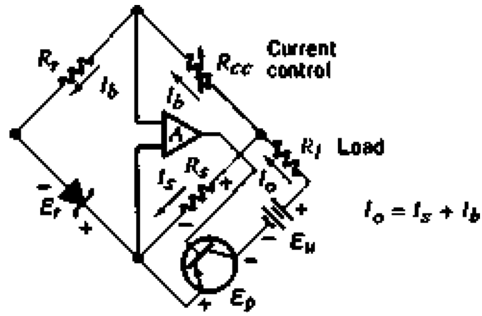


Fig. 4.9 Current regulating circuit.

The voltage to be regulated is the drop caused by the load current flowing through R_S, the sensing resistor. The feedback resistor in this case is R_{CC}, a current control. The only current flowing through the current control is the bridge current. When the voltage drop in R_{CC} due to I_b equals the drop in R_S due to I_S, the conditions for bridge balance are met, and a null exists across the input terminals of the error amplifier.

The output current of this circuit can be programmed by varying R_{CC} or changing I_b, either event causing the voltage drop across R_{CC} (or the R_{CC} terminals) to change. In order to restore the balance condition, the voltage drop across R_S must change. This happens when the feedback to the pass element alters its conduction, so as to permit a change in load current in the proper direction and amount as to restore the balance. Here, as has been the practice throughout this manual, the illustrative circuits are drawn for a PNP pass transistor. With proper polarity alteration, they are equally applicable to NPN transistors or vacuum tubes in hybrid units.

The fact that current is regulated indirectly, as the drop across a sensing resistor gives rise to some problems which did not affect the performance of voltage regulator circuits. For one thing, the output current depends on the constancy of the ratio E_S/R_S, in which the resistance temperature coefficient of a power dissipating element enters. Dissipation in the sensing resistor means a temperature rise, which will naturally cause some change in resistance according to the temperature coefficient of the resistance material used. Ideally, the sensing resistor should have a very large surface area, so that it can dissipate the sensing power with a minimum temperature rise. Practically, this means that the power rating of the sensing resistor should be much, much larger than the actual power to be dissipated.

10:1 is the minimum excess rating that should be considered; 100:1 is better if feasible from a size point of view. In really critical applications, the sensing resistor may be immersed in an oil bath to carry away heat at a more rapid pace.

To minimize the heat dissipation requirements, the size of the sampling drop should be made as small as possible. But dissipation in the sensing resistor is not the only difficulty encountered in current regulation: the problem of regulating a very small voltage must also be taken into account. While reducing the sampling voltage seems to be a good idea with respect to the resistor, it is a problem with respect to the amplifier. Whatever drop is chosen, its magnitude represents the total amount of the load current. If, for example, 0.5 volts is the maximum sensing drop for a 100 ampere supply, then changes of 1 ampere manifest themselves as a *5 millivolt* change in the sensing sample. 100 ma corresponds to a half millivolt, while a 1 ma load current change produces only *5 microvolts* change in the sample. 10 ma is a 0.01% change in 100 amperes, so to regulate current to 0.01%, the amplifiers must be sensitive to a good deal better than 50 microvolts. On the other hand, the 0.5 volt drop is generating 50 watts of dissipation at 100 amperes. This is quite a substantial amount of power to be dissipated without significant temperature rise. The amount of the sample can be increased to give an increased error amplitude to the comparison amplifier only at the sacrifice of increased power dissipation, which means either a larger resistor or an increased temperature rise in the sensing resistor. Likewise, the sensing voltage can be reduced for less dissipation only by forcing the error amplifier to work with impossibly small signals, with a consequent poor signal to noise ratio.

This conflict between the requirements for good current regulation and low sensing resistor dissipation taxes the designer's ingenuity in devising good current regulator circuits. The numbers used in the preceding example happen to be actual conditions prevailing in a Kepco power supply. In the Model KS 8-100M Power Supply, a 0.5 volt drop (at 100 amperes) is the actual sample used. The miniscule error signals give some idea of the resolution and gain requirements placed on the comparison amplifier in order to obtain true 0.01% current regulation performance throughout the current range of such a power supply.

As the current control is used to reduce the output current from the maximum value, the sampling voltage is naturally reduced. Eventually, as the sample becomes very small, the variations in current can no longer produce a recognizable error signal, and regulation performance suffers somewhat. For most supplies, employing a fixed sensing resistor, current control can be exercised over a 10:1 or 20:1 ratio of currents before the degradation becomes noticeable. Control over the current will continue right down to near zero, but the regulation performance below one tenth of the nominal output rating will be reduced. The major advantage of external current sensing is the fact that the user can extend the regulation range down to very small currents merely by reselecting the value of the sensing resistor to drop a large sample. With only three resistors, each covering a 10:1 ratio of current, a regulator can be made to cover a 1000:1 current ratio when external sensing resistors are used.

The size of the sample depends, as we have seen, upon the power dissipation requirements. For external sensing, the size of the sample will also depend on the current ratio to be programmed and on the amount of compliance voltage that can be sacrificed for the sensing voltage. Since the sensing resistor is in series with the load, voltage dropped across it cannot be used at the load. For most all-transistor power supplies, Kepco recommends a 1 volt drop. For hybrid models, whose current ratings are usually somewhat lighter, a 10 volt sample is recommended. One reason for whole number values, as for example, 1 volt and 10 volts, is that it makes computations somewhat easier. The internal sensing resistors used in automatic crossover power supplies are generally selected to drop 1 volt for supplies with current ratings below 5 amperes.

The following sample calculation illustrates how to calculate the value of the sensing resistor and current control for an external sensing application.

Example: A power supply programmed at 1000 ohms per volt (bridge current equals 1 ma). Set up to program 0.2A to 2 amperes using external sensing.

Solution: Choosing to drop 1 volt at 2 amperes, R_s must be 0.5 ohms. The power dissipated will be 1V × 2A = 2 watts. Use a 20-200 watt resistor. Since the power supply programs at 1000 ohms per volt, a 1000 ohm rheostat-connected potentiometer is needed at R_{CC} to generate the 1 volt needed for 2 ampere output. To obtain half current, make R_{CC} = 500 ohms, programming 0.5 volts across 0.5 ohms, giving 1 ampere. 250 ohms causes the power supply to deliver ½ ampere, and so forth.

Intuition tells that a current sampling system cannot regulate zero current. A glance at Figs. 4.8-9 will show that since the bridge current I_b circulates through the load, I_b represents the actual minimum current that can be generated; even if R_s is made large and R_{cc} made zero, I_b remains the minimum current.

4.5 SMALL CURRENT REGULATION

The inherent, minimum regulated current limitation can be overcome by using one of the operational current regulator techniques introduced in Chap. 3. By employing a separate power supply as a control, the need for a bridge current is eliminated.

The circuit shown in Fig. 4.10 eliminates I_b by replacing R_{cc} with an external voltage source, which might be a small variable power supply. Assuming that R_s is still chosen for a 1 volt drop at the maximum desired load current, E_{cc}, the programming source would be made to vary 0-1V d-c. As no control currents circulate in this circuit, the output current can be programmed right down to true zero by making $E_{cc} = 0$. The problems associated with regulating extremely small voltages still limit performance, although so far as the E_{cc} source is concerned, the difficulty can be relieved by making E_{cc} the output of a resistive divider across a larger voltage. Because the comparison amplifier draws only a very small leakage current (which may be cancelled), simple resistive dividers can be used without fear of loading.

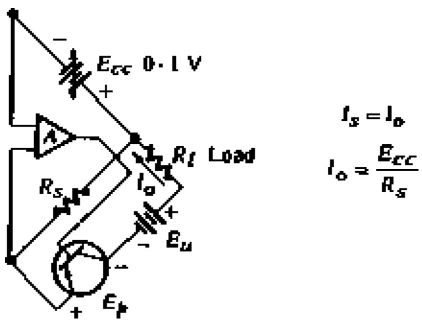


Fig. 4.10 Alternate method of current control, not limited by a bridge current minimum.

Another operational regulator technique, useful for the control of ultra small currents, dispenses altogether with the need for a small voltage sample by taking advantage of isolation properties of the null junction.

Since the null junction is a virtual ground potential, the current I_b across it depends only on the ratio of E_r/R_r. If E_r is replaced by a source of variable voltage E_1 (another power supply), and R_r is replaced by an external range control resistor R_1, a load can be connected in place of the voltage control across the feedback terminals of the circuit, as shown in Fig. 4.11.

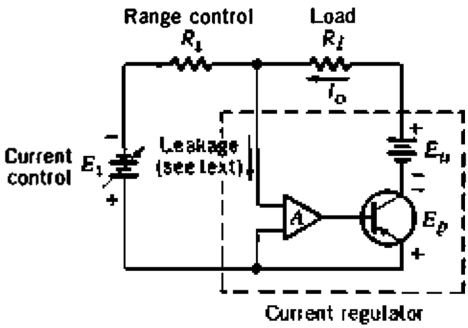
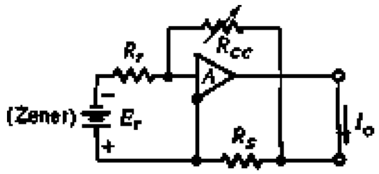


Fig. 4.11 Small current regulator.

The current flowing in these terminals is equivalent to the bridge current in the conventional circuit arrangement. Just as I_b was independent of R_l, the load current is independent of R_l and current is regulated. The regulator output terminals of the power supply are not used for anything in this configuration. They do repeat the voltage across the load, however, and being at a low impedance, serve as an excellent way to monitor the voltage compliance. The adjustable voltage E_1 serves as a current control, while R_1 can be used as a range control. The minimum current that can be programmed with this method is set by the small leakage currents into the comparison amplifier null junction. This current is small, on the order of 5-15 microamperes. It can be nulled for extremely small current regulation using the cancellation technique described in Sect. 3.3(B). The technique, briefly, consists of a small opposing current, derived from auxiliary sources and injected into the null junction in such a way as to cause the no-excitation current in the feedback (load) circuit to be truly zero.

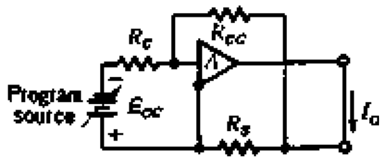
The major advantage of this current regulating technique is that the comparison amplifiers are allowed to operate directly upon the load current, rather than through the attenuation of a sampling resistor network. Also, power supply E_1 can operate at a relatively high voltage where its own error signals are sufficiently large to insure good regulation by its own comparison amplifier. The major disadvantages of this circuit are that both the current regulator and the supply used for E_1 must both be capable of delivering the maximum load current. That current also flows through R_1, where dissipation considerations will limit the maximum current that can be efficiently controlled. Fig. 4.12 summarizes the various current regulating circuits.



$$I_o = E_r\left(\frac{R_{cc}}{R_r}\right)\left(\frac{1}{R_s}\right) + \frac{E_r}{R_r}$$

1. Resistance controlled
2. Suitable for large current
3. I_o minimum $= E_r/R_r = I_b$

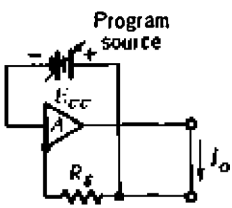
(a)



$$I_o = \frac{E_{cc}}{R_c}\left(1 + \frac{R_{cc}}{R_s}\right)$$

1. Indirect voltage controlled
2. Suitable for large current
3. I_o minimum = 0

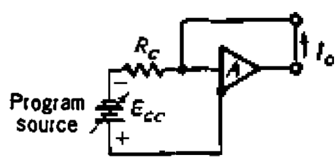
(b)



$$I_o = \frac{E_{cc}}{R_s}$$

1. Direct voltage control
2. Suitable for large current
3. I_o minimum $\cong 0$

(c)



$$I_o = \frac{E_{cc}}{R_c}$$

1. Indirect voltage control
2. Suitable for small current
3. I_o minimum $\cong 0$

(d)

Fig. 4.12 Tabulation of four basic current regulators and their characteristics.

5 METHODS OF CONTROLLING POWER DISSIPATION

5.1 SERIES AND SHUNT REGULATORS

The power handling portion of a d-c power supply consists of a power generator (the transformer, rectifier and filter), plus a power absorber (the regulator's pass element). The output is equal to the power generated less the power absorbed by the regulator. There are basically two kinds of absorptive regulators, a series regulabor and a shunt regulator. Both are diagrammed in Fig. 5.1.

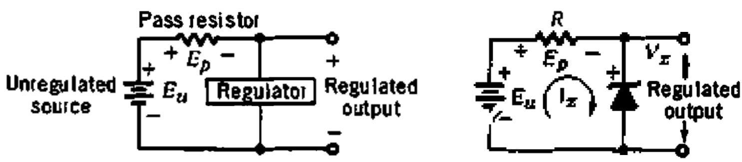


Fig. 5.1a-b Shunt regulator. The zener diode regulator circuit is a typical example of a shunt regulator. It is widely used in simple low-power applications, and as a reference potential generator for more elaborate circuits.

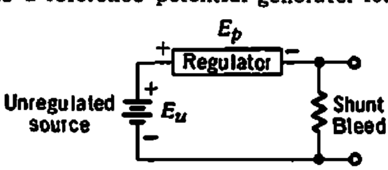


Fig. 5.1c Series regulator, widely used for high powered adjustable power supplies.

In reality, both control the output in the same way, by causing a voltage drop E_p which is varied automatically to subtract just the required amount from the raw unregulated d-c, E_u. The shunt regulator operates by causing a current flow through the pass resistor to drop the voltage E_p. When not externally loaded, a shunt regulator conducts the maximum current needed to establish E_p. When such a power supply is loaded externally, the regulator conduction is reduced by an amount equal to the load current, so that (neglecting changes in the raw d-c), the total current through the series pass resistor remains constant. Such a design dissipates the maximum amount of internal power when its output terminals are open and unloaded. When fully loaded externally, the current through the regulator element reduces to a minimum. Dissipation in the pass resistor is constant irrespective of loading.

A series regulator functions in much the same way as the shunt device, except that the regulator itself now takes the voltage drop E_p directly across its own terminals. A bleeder resistor may be used to establish a minimum load across the output, but in practice, a transistor or tube can be made to cut-off so well that only a very small bleeder current is required. For all intents and purposes, the bleeder dissipation is negligibly small. (In actual Kepco power supply design, the bleeder is replaced by a small auxiliary power source to permit operation at zero output voltage.) When a supply employing the series regulator is not loaded, very little current (only the bleeder current) flows through the pass element, so that its power dissipation is quite small. At full load, the dissipation increases directly with the load current.

The output voltage from either type of circuit may be adjusted by causing the regulator to vary the voltage drop E_p. E_p is large for small output voltages and small when output voltage is large. In a series regulator, this means that dissipation in the pass element is minimized at the highest output voltages, and maximized at the lowest. Since the regulator in the shunt circuit sees the entire output voltage, its dissipation is greatest at the maximum voltage setting, and minimum at the lowest.

Because dissipation is generally less in the series circuit, and it lends itself well to pre-regulator dissipation reduction schemes, the series circuit has been favored by most power supply designers. An important factor is that by manipulating the raw d-c in concert with the output setting, the series regulator can be made to work well with just a small portion of the total voltage dropped across the the pass element (Fig. 5.2).

The amount of voltage drop that the pass element needs to sustain regulation is the sum of all the internal losses in the supply, plus 1) the contribution of line voltage variations, 2) the peak-to-peak ripple that gets past the first or "input" filter, 3) a margin for remote error sensing, 4) the minimum voltage needed for full conduction, and 5) a safety factor. The sum of all these factors will usually be much less than the total voltage; i.e., seldom more than

20-25%, and in combination with more sophisticated pre-regulators and good input filters, perhaps as little as 5%. (See Fig. 5.2.)

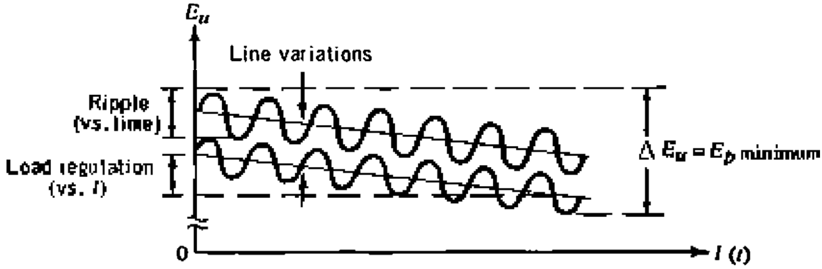


Fig. 5.2 ΔE_u is the combined load and line regulation, plus the p-p ripple. E_p, the pass element drop, must at least equal ΔE_u to maintain control for all input (line) and loading conditions. Other factors not shown on the composite diagram include:

1. Allowance for remote error sensing
2. Minimum conduction voltage required by the pass element
3. Transient loading allowance.

In small power supplies (in the 50 watt and under class), it is often desirable to make no attempt to limit the series regulator dissipation. In this way, the designer can avoid compromising other desirable power supply performance characteristics, such as current regulation, response speed, programmability, linearity, etc. (Fig. 5.3.)

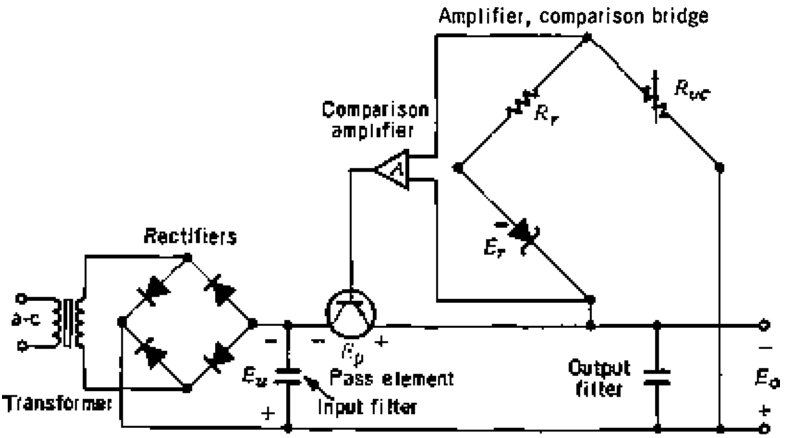


Fig. 5.3 Full dissipative regulator circuit. The reference voltage is shown symbolically as a zener diode; actually, a circuit similar to Fig. 5.1b generates the voltage E_r.

5.2 FULL DISSIPATION REGULATORS

In this kind of design, the pass element tubes or transistors are called upon to dissipate the full output of the power supply under minimum voltage and full-load conditions. Frequently, multiple series or parallel elements are used to increase the dissipation capabilities of the pass section. Typical Kepco designs in this class are the ABC, CK, PAX and PBX designs.

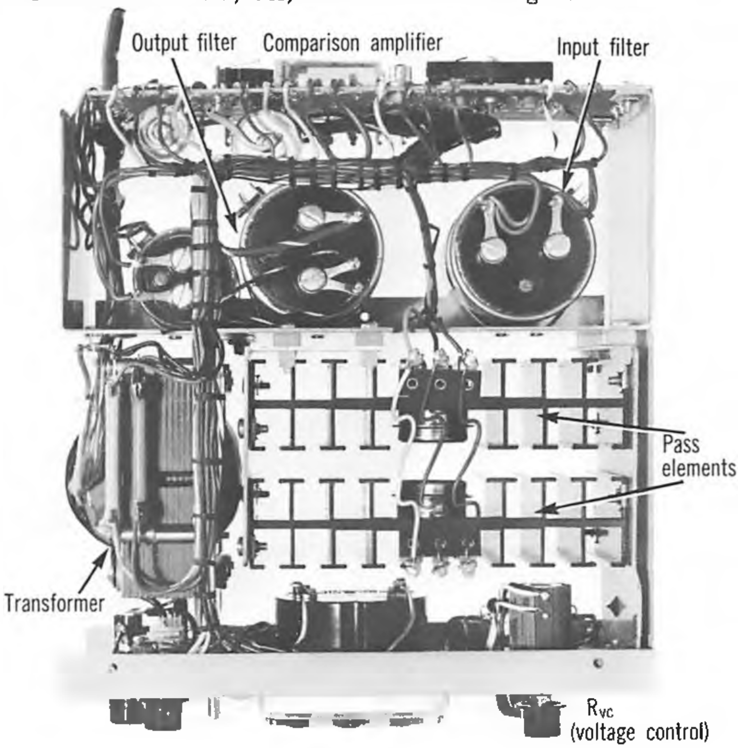


Fig. 5.4 Interior view of ABC design power supply.

The outstanding characteristics of this group are their programmability (including high speed capabilities), current regulation, and short circuit immunity. The latter is associated automatically with full range current regulation because of the need to be able to operate indefinitely at zero compliance – into a short circuit. This ability grows out of the capability of the pass elements to absorb the full raw d-c supply voltage, and is indeed a reason for this kind of design. The ABC, PAX and PBX feature adjustable current limiting (in addition to their ability to be set up as current regulators). When shorted, the series pass elements of such supplies tend to cut off, passing only such current as is allowed by the setting of a current limit control which acts as the result of the voltage drop across a small current sensing resistor. The CK design is an automatic crossover circuit which has very sharp limiting characteristics described as current *regulation.* The PBX regulator is an all-silicon design using NPN pass transistors so polarities are reversed.

5.3 STEPPED SWITCH TRACKING

A simple yet effective way to achieve a degree of reduction in pass element dissipation is illustrated in Fig. 5.5.

In this design, a multiply-tapped transformer secondary is ganged with a step switch in a portion of the control bridge voltage control resistance. The raw d-c input voltage to the regulator is increased or reduced in step with the voltage control selector switch, so that the voltage drop across the pass elements (in this case, the plate-cathode voltage drop of a vacuum tube) is relatively constant irrespective of the output voltage setting.

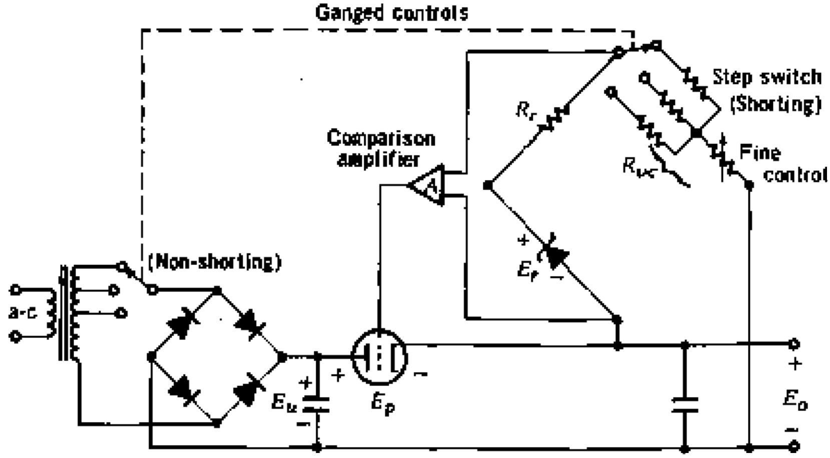


Fig. 5.5 Step switch pretracking, drawn for the Kepco hybrid circuit which uses a vacuum tube for a high voltage pass element. The comparison amplifier is transistorized.

The voltage drop across the pass element must be large enough to accommodate all of the line regulation, ripple and factors previously mentioned, plus an amount equal to the inter-tap voltage, but it does not have to accommodate the full output voltage rating in the steady-state. This is the major advantage of the tap switch design. (See Fig. 5.6.)

This circuit is used in the Kepco HB group of power supplies and has proved a very successful approach to practical dissipation control. One disadvantage is that programmability, and hence, current mode voltage compliance, is limited to the span of one switch position (a voltage drop that is allowed for a "fine control"). Operation beyond the limits of any span is usually possible, however, subject to some derating to account for the increased dissipation as the pass regulator's pass voltage drop is caused to increase. (See Appendix for derating curves applicable to operation of HB power supplies beyond the limits of a single span.)

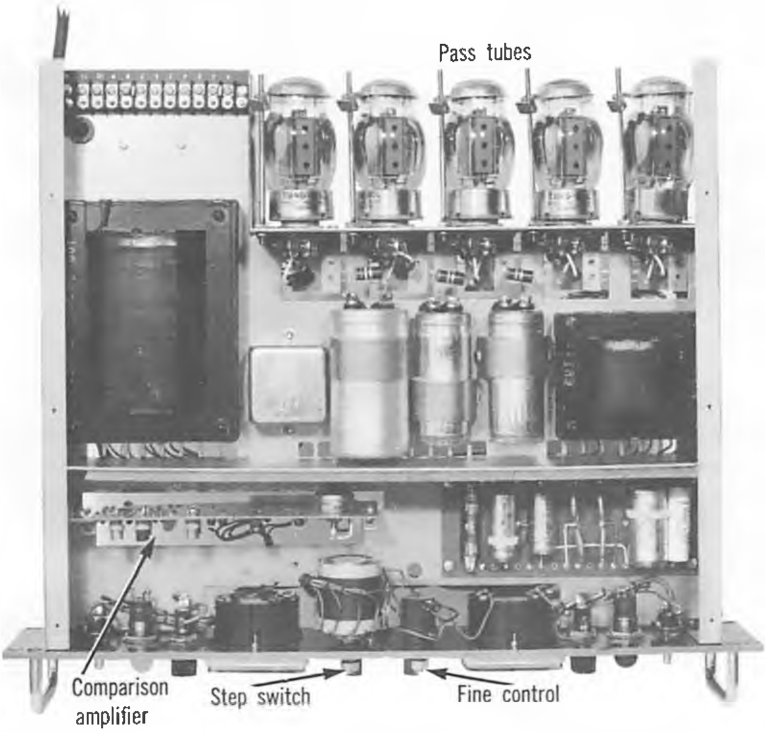


Fig. 5.6 Hybrid (HB) power supply design, employing stepped switch tracking.

5.4 VARIABLE TRANSFORMER TRACKING

Another popular dissipation control technique is to use a variable auto-transformer ganged to the voltage control of the power supply (Fig. 5.7).

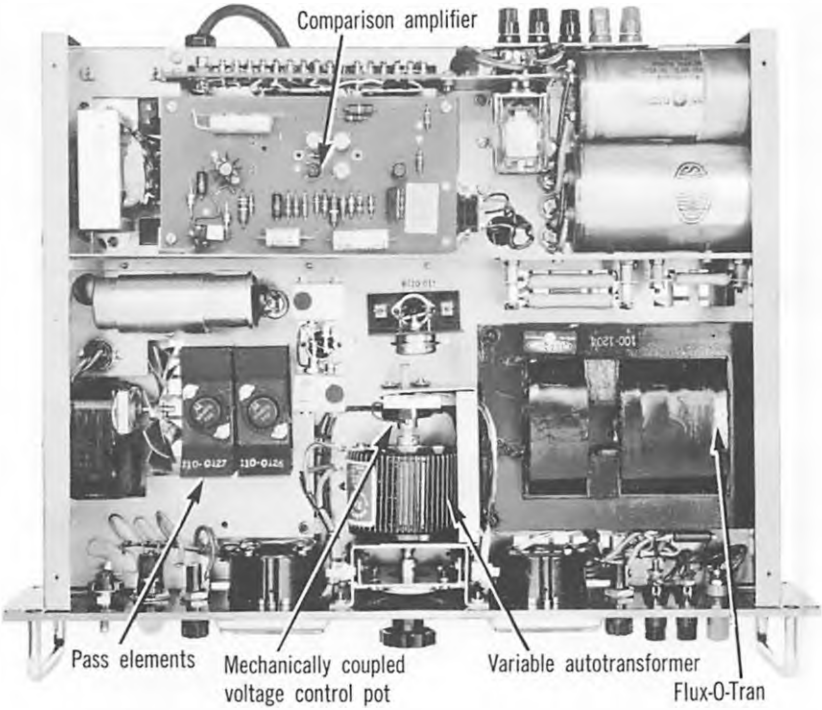


Fig. 5.7 SM design employing variable auto-transformer tracking.

Such a design is a logical extension of the switch selected secondary tap arrangement. Its operation can be visualized as simply a very large number of tap positions on the transformer selected by a shaft rotated brush, instead of a separate switch. This technique is used in Kepco SM and WR power supplies, as shown by Fig. 5.8.

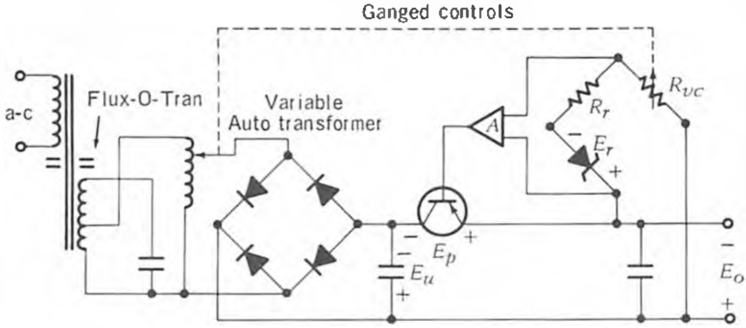


Fig. 5.8 Variable transformer pretracking, SM design.

The variable auto-transformer introduces another lossy element in the output circuit which must be accounted for in the voltage drop of the pass elements. It more than makes up for this loss by permitting the designer to control the pass drop to much tighter limits – and thus reduce dissipation. By controlling the pass dissipation in this fashion, much larger amounts of output power can be regulated than would otherwise be possible. A 40 watt (40 volt, 1 ampere) transistor, for example, might be used to regulate as much as 400 watts (400 volts, 1 ampere), if employment of a variable auto-transformer enables the designer to keep the pass voltage under 40 volts. This imposes fairly stiff limits on the tracking accuracy of auto-transformer and voltage control, but it is, nevertheless, illustrative of the kind of advantage that can be obtained. One question that must occur to the reader is consideration of what might happen under short circuit conditions. Current limiting, or current regulated crossover, is obviously out of the question, since such operation requires the pass elements to be capable of absorbing the full unregulated voltage.

The solution – which makes such circuits practical – is a combination of two circuit innovations. First, the pass elements are caused to saturate during a short circuit. This prevents excess voltage from appearing across the pass but does nothing to limit the current, which, if not controlled, would quickly destroy any real pass element. The current limiting is provided by using a Flux-O-Tran (flux oscillating transformer) as a power transformer. In addition to the line regulation abilities of the Flux-O-Tran outlined in Chap. 1, the transformer also provides a form of intrinsic current limiting, as can be seen from the output curve reproduced in Fig. 5.9. By com-

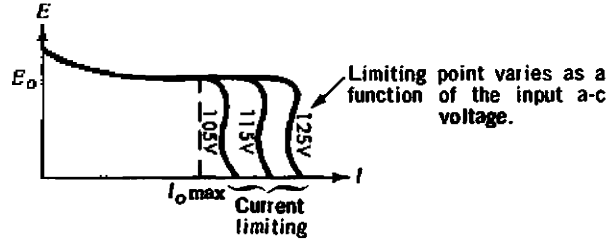


Fig. 5.9 Overload characteristics of Flux-O-Tran.

bining a saturating pass element with the current limiting Flux-O-Tran, complete short circuit protection is achieved. Fuse or circuit breaker back-up is usually provided for long term short circuits, to interrupt primary power and turn the set off.

The use of the line regulating Flux-O-Tran produces the added advantage of reduced variations at the pass regulator. This is of enormous advantage when using relatively low voltage elements in a high voltage circuit, as in the preceding example. In fact, the example wouldn't work unless line preregulation were used. A ±10% line voltage variation would reflect through an ordinary transformer as ±40V or 80 volts that the pass element would have to handle — obviously impossible for the 40V transistor. The interposition of the line regulating flux oscillator can reduce this to ±1% or only 8 volts, well within the capabilities of a 40 volt transistor. In a practical circuit such as the Kepco SM design, the combination of a Flux-O-Tran power transformer, a tracking variable auto-transformer, and the saturation technique for absorbing short circuits, permit an all-transistor power supply design to handle a 0-325 volt output range at currents up to 2 amperes. The Kepco Model SM 325-2M is an example of such a power supply using transistors in the 80-100 volt range.

Because of the mechanical coupling between variable auto-transformer and voltage control, this system of dissipation control is not suited for remote programming, nor can it be used as a current regulator.

5.5 LINE PRE-REGULATION

Another power supply design which makes use of the unique properties of the Kepco Flux-O-Tran is the PWR series of regulated power modules. These units resemble the SM series in that the Flux-O-Tran is followed by a full feedback regulator. Of particular interest is the form of the short circuit protection provided in PWR Modules. When overloaded, the PWR Power Supply has an output characteristic as shown in Fig. 5.10a.

For a full short circuit, the output voltage is zero and the output current is reduced to about 5% of the rated current.

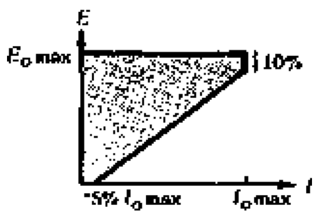
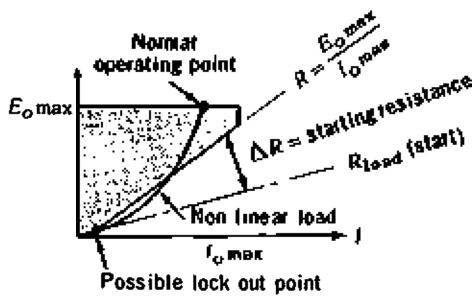


Fig. 5.10a Overload characteristic for the Kepco PWR module.

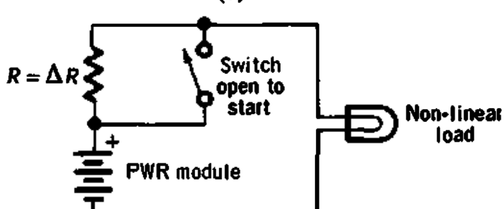
In the PWR circuit, the Flux-O-Tran provides line pre-regulation by "smoothing" the variations in line voltage by more than 10:1, and thus permitting the pass transistor to function with a smaller voltage drop, regulating with a reduced internal dissipation.

A full-sized output characteristic plot for PWR models is reproduced in Fig. 4.3. Note that the output derating curve follows the overload characteristic, so that as the power supply is programmed for lower output voltage (increased pass voltage), the output current limit is reduced by an exactly equal amount. By this means, maximum pass dissipation is maintained constant throughout the programming range. The negative resistance character of the cut-off locus may cause severely nonlinear loads (requiring large starting surge currents) to "lock out" for certain settings of the cut-off current limit. For such loads, a series starting resistance can be used to limit the starting surge, so that operation can be contained within the shaded region of the graph. Examples of nonlinear loads are high intensity filament lamps and motors.

The starting resistor value is equal to the incremental difference between the initial resistance of the load and the equivalent resistance of the cut-off locus E_{max}/I_{max} (see Fig. 5.10b). In the



(b)



(c) Starting circuit. $R_{max} = \frac{E_{o\,max}}{I_{o\,max}}$ when R_{load} (start) = 0.

Fig. 5.10b-c Starting arrangement to avoid lock out of PWR circuit with non-linear loads.

worst case, for loads which present an initial short circuit, the starting resistor equals E_{max}/I_{max}. The starting resistor should be shorted out for normal operation after starting (see Fig. 5.10c).

5.6 SCR PRE-REGULATOR

A third method of safely increasing the output capability of the power supply's pass elements is to add a pre-regulator. This combination has a number of advantages which derive from the division of the regulating function between the pre-regulator and the series pass regulator, taking advantage of the particular capabilities of each type. The series pass regulator is used to reduce ripple, absorb transients, and to provide fine and fast regulation of the output. The pre-regulator is used to handle large and relatively slower regulation demands on the power supply. The pre-regulating mechanism employed by Kepco in its KS design makes use of silicon controlled rectifiers (SCR's) in a bridge arrangement to adjust the raw d-c supply by controlling the conduction angle of the rectifiers. (Fig. 5.11.)

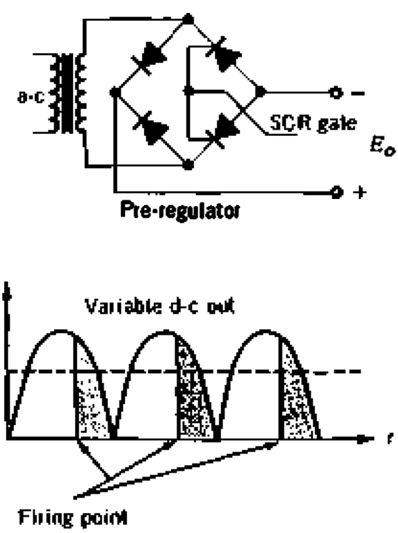


Fig. 5.11 Full wave SCR bridge. By adjusting the timing of the firing (conduction) pulse applied to the SCR gate, the d-c output is varied.

SCR's conduct only after receiving a turn-on signal at their gate, and can be turned off only after the anode-to-cathode excitation has been removed. The gate pulse is generated by a unijunction transistor

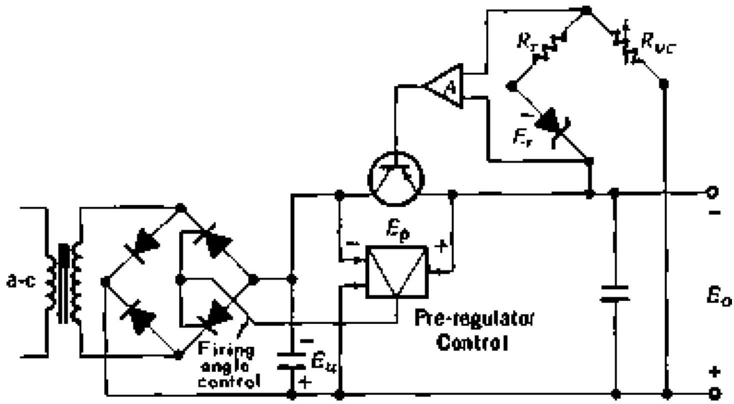


Fig. 5.12 Firing angle (time) determined as a function of the difference between the pass drop E_p and a pass reference.

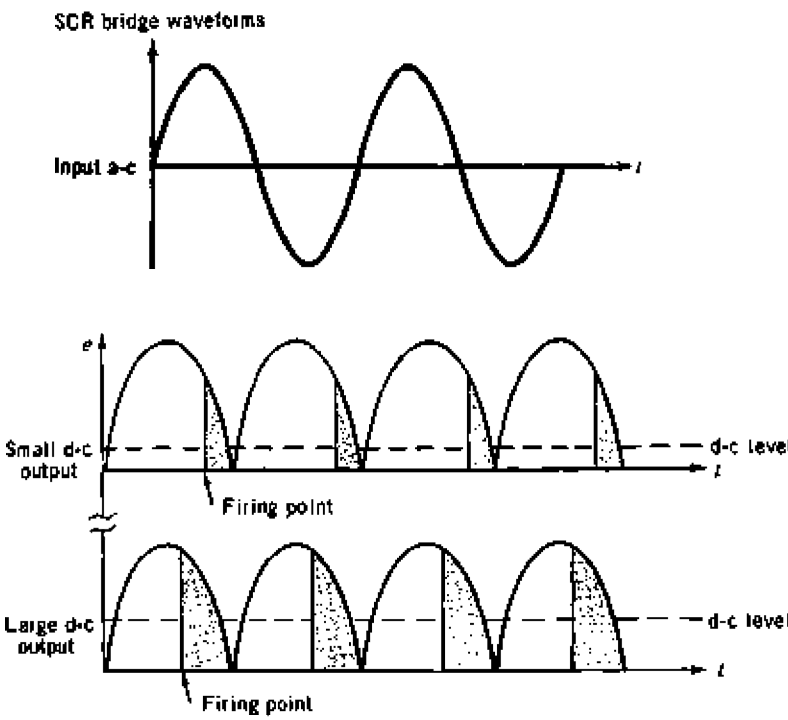


Fig. 5.13. SCR phase angle control over d-c level.

oscillator whose emitter junction is supplied from a synchronized timing ramp. The slope of this ramp, and thus the conduction timing for the SCR's, is made dependent on the error signal derived from the pre-regulator. In operation, the SCR pre-regulator can be visualized as a non-mechanical electronically variable auto-transformer whose output is made to track the output voltage setting of the power supply, in order to control the voltage across the main pass elements. A schematic/block diagram of the entire system is shown in Fig. 5.12. and the SCR waveforms in Fig. 5.13.

Because it is automatically self-adjusting, the power supply can be externally controlled using its voltage control remotely (or any of the many possible ways by which a bridge regulated power supply can be programmed). It follows, also, that this design is suitable for use as a current regulator. In fact, the KS Power Supply design can do all of this. It is one of the most versatile high power supplies made. (Fig. 5.14.)

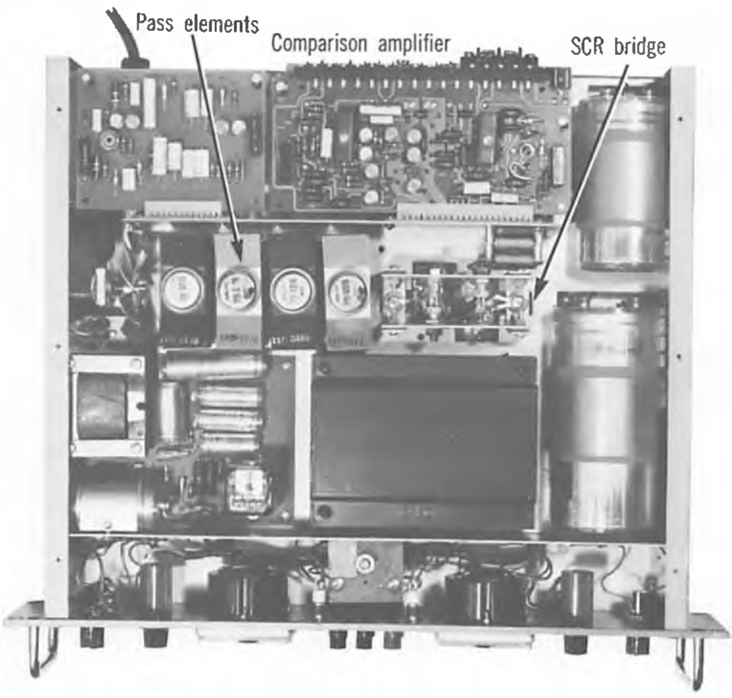


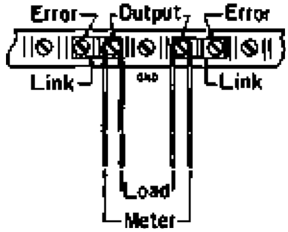
Fig. 5.14. Interior view, typical KS series power supply.

6 MEASURING POWER SUPPLY PERFORMANCE

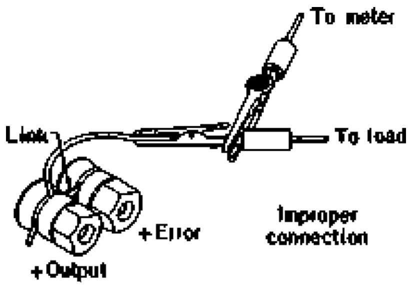
6.1 VOLTAGE REGULATION

A most significant characteristic of power supply voltage regulation is the very small internal or source impedance that such a supply maintains across its output. To calculate just how small this impedance is, first convert the power supply's published or measured load regulation into an absolute voltage change, then divide this value by the output current rating. For example: A 36 volt power supply with "0.01% load regulation" means that the terminal voltage changes no more than 3.6 mv or 3.6×10^{-3} volts for a load change from no load to full load. If that power supply is rated for 0-5 ampere output current, then its internal impedance is $(3.6 \times 10^{-3})/5 = 0.72$ milliohms or 0.00072 ohms. Measuring the load regulation of such a power supply is, in effect the process of measuring the value of this source impedance. As might be imagined, some rather special precautions and techniques are needed to measure 0.00072 ohms with any degree of accuracy. First and foremost, it is absolutely necessary to make certain that the measuring apparatus itself – or the manner of its connection – does not introduce a series resistance comparable to the 0.00072 ohms that is being measured. There are several practical ways to do this, all variations on the familiar four-terminal network (Fig. 6.1).

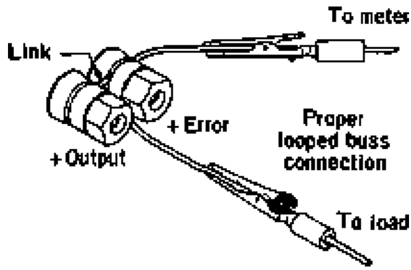
The idea is to avoid passing *any* current through the measuring circuit wires, while at the same time connecting the measuring circuit wires physically as close as possible to the source resistance under investigation. In terms of a physical power supply, the diagrams show some practical four-terminal networks for the kinds of connectors most often used on power supplies. It is surprising how much resistance even the shortest length of wire has – relative to the source resistance of a well regulated power supply. Even worse is the contact resistance of a clip lead or other connector.



Looped buss connection method assures minimum load current in metering circuit.



Improper connection: resistance of the wire and alligator clip will obscure the internal source impedance being measured.



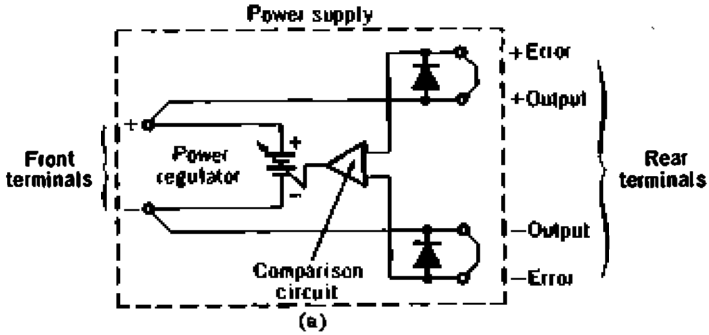
Correct connection: heavy buss wire is threaded through error terminal. One end to the load, the other end to the measuring instruments.

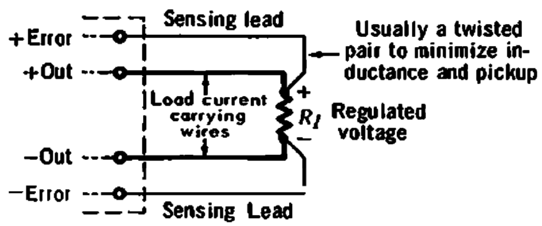
Fig. 6.1

The Kepco wire-loss nomograph, reproduced in the Appendix, permits a rapid calculation of the resistance and voltage drop versus current for any of the most common AWG copper wire sizes.

(A) Remote Error Sensing: Remote error sensing connections are provided on most precision power supplies designed to deliver substantial current. These connections extend the four-terminal network idea to the load itself, allowing the power supply to regulate voltage directly at a critical portion of the load (Fig. 6.2).

The sensing wires are usually connected to a separate set of rear terminations which are externally linked to the output terminals. Notice that in most Kepco designs the output wires go first to the front terminals, including them *inside* the control loop. If a load is connected to the front terminals, the front-to-rear load wire acts simply as an extension of the error sensing so that error sensing is properly accomplished at the power supply's front terminal. Despite this, front terminal regulation for most heavy current supplies is usu-





(b) Remote error sensing.

Fig. 6.2 Diagram of power supply output and error-sensing wiring.

ally not as good as the measured regulation at the rear terminals, because the sensing connection is made at the inside (rear) part of the connector. This kind of connection leaves the body of the connector (passing through the panel) an uncompensated series resistance whose voltage drop cannot be distinguished from the voltage drop due to the internal impedance (load regulation). (See Fig. 6.3.)

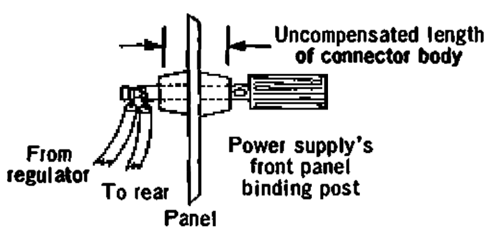


Fig. 6.3 Front panel connector.

To obtain anything like the rated voltage regulation performance of a precision power supply at the load, remote error sensing *must* be used to compensate for the voltage drops caused by the wiring resistance.

The error sensing links at the rear of such a power supply are easily disconnected so that separate light wire can be used to connect the sensing points to the critical part of the load. These sensing wires carry only the control bridge current I_b of the supply which is constant, and so do not contribute a varying IR drop.* The polarity of the load and sensing connection must be maintained, so, if shielded or twisted wires are used, they should be color coded to permit the user to identify the *plus* error sensing wire with the *plus* output terminal, and the *minus* error wire with the *minus* output terminal. In connecting the sensing and output wires to a load, the same precautions described earlier concerning four-terminal networks and wire drops must be faithfully observed. In particular, since the power supply can sense at only one place in the circuit, some thought and care should be given to select the most critical place for the sensing connection. If the circuit permits, a variation of the automatic sensing selection technique, used for the front-to-rear wiring in the power supply, might be used to advantage (Fig. 6.4).

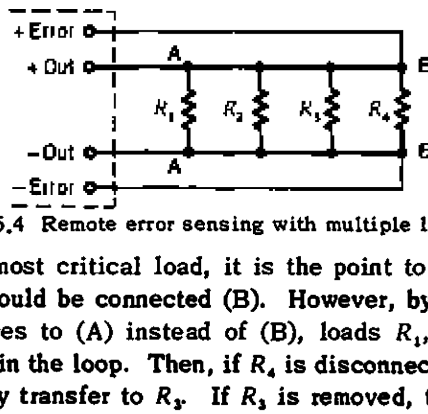


Fig. 6.4 Remote error sensing with multiple loads.

If R_4 is the most critical load, it is the point to which the error sensing wires should be connected (B). However, by connecting the output power wires to (A) instead of (B), loads R_1, R_2 and R_3 can also be included in the loop. Then, if R_4 is disconnected, the sensing will automatically transfer to R_3. If R_3 is removed, the sensing will transfer to R_2, and so forth down the parallel string.

Recalling Fig. 6.2, note that there are diodes connected between each error terminal and its respective output terminal. These diodes are installed to prevent an uncontrolled power supply voltage response should the error connections be inadvertently opened. Without the diodes, an open-sensing circuit would detect no output and so would command the power supply to deliver either its maximum voltage or zero output, depending on which circuit was opened. As the resultant uncontrolled output might be damaging to some loads, sensing diodes are added as an overvoltage protection. Should the voltage between any output terminal and its associated error sensing lead exceed the forward conduction voltage of the diode, that diode conducts and

*In metered units, the meter current, a variable, will also flow in the sensing circuit, but this is small and can usually be neglected.

clamps the error terminal to its output terminal with a voltage rise not exceeding the forward conduction drop of the diode. Sometimes special double diodes are used in this circuit; such diodes consist of two rectifiers in series in a single package. If each is silicon, then the conduction voltage is approximately twice 0.7 volts or about 1.4V d-c. With the usual restriction of the error sensing to 0.5 volts per lead, the diodes do not normally conduct. In special loading situations where it may be desired to increase the error sensing drop above 0.5V per wire, the error sensing diodes should be removed, or extra diodes should be connected in series with the existing ones to increase the forward conduction voltage to approximately 50% greater than will actually be dropped in the load wires.

When remote error sensing is employed, special precautions should be taken to avoid disconnecting the load wires while the sensing leads remain in place. If this happens while power is *on,* the load current will attempt to flow through the error diodes and the light duty sensing leads, and may damage them.

To prevent any possibility of this happening, it is desirable to make all load connections with the power supply turned *off.* If quick-disconnect connectors are used between the power supply and its load, that connector should be equipped with a pair of long pins for the load circuit to make sure the load wires are firmly connected before the sensing leads make contact.

(B) Dynamic System Behavior: One reason for the use of voltage regulators is to make use of the low power supply source impedance to decouple circuits that might otherwise mutually interact. A point often overlooked by designers, however, is the reactive contribution of the wires connecting the power supply to the load distribution point. True, remote error sensing will compensate for the d-c *resistance* of the load-connecting wires, but the error sensing circuits are powerless to cope with the effects of wiring *inductance.* This is because the sensing wires themselves have as much inductance as the load wires (or perhaps even more).

For best performance, the load-to-supply wiring should be arranged to minimize the effect of wire inductance. The self-inductance of a straight wire is given by the formula

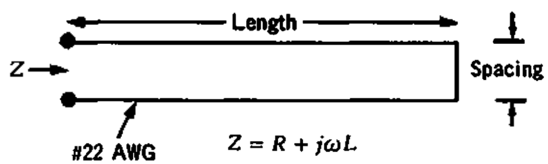
$$L \text{ (in } \mu h) = 0.005l\left(2.3 \log \frac{4l}{d} - 0.75\right)$$

where l is the length of the wire and d its diameter, both in inches. Thus, the shortest, heaviest piece of wire is the least inductive.

Twisting the wires together will further reduce the inductance. Similarly, the remote error sensing wires should be twisted together, and in some applications should be shielded to reduce noise pick-up. The loop inductance of a pair of parallel wires is tabulated in Table 6.1 for various lengths and spacing. These are *straight*, parallel wires, looped or kinked wires have much larger effective inductance.

Table 6.1 Parallel wire inductance table. "L" in microhenries ($10^{-6}h$). #22 AWG parallel wires.

Spacing / Length	Twisted	¼"	½"	¾"	1"	2"	4"	6"	d-c Ohms
1'	0.29	0.47	0.53	0.56	0.59	0.67	0.74	0.80	0.034
2'	0.48	0.82	0.97	1.02	1.12	1.31	1.44	1.60	0.061
3'	0.69	1.12	1.38	1.58	1.69	1.92	2.20	2.36	0.084
4'	0.91	1.47	1.98	2.10	2.23	2.51	2.62	2.80	0.112
5'	1.06	1.92	2.36	2.55	2.63	2.90	3.20	3.45	0.138



To establish a low source impedance at the load, at frequencies where the remaining line inductance is significant, a capacitor by-pass directly at the load terminals is very useful. Such a capacitor acts as a local energy source, compensating for the load wiring inductance.

The Reactance – Impedance Chart reproduced in the Appendix can be used to determine the effect of various amounts of wiring inductance, and the optimum value of capacitance required to establish a specific impedance at any given frequency.

When particularly long sensing leads are used, it is sometimes helpful to connect local sensing bypass capacitors at the power supply. These would be placed between the individual *plus* and *minus* output terminals and their respective error sensing terminal. They serve to bypass the combination of load-wire and sense-wire inductance and can improve the loop stability.

(C) Equipment Requirements: Power supply measurements almost always involve the measurement of a very small d-c voltage change in the presence of a much larger voltage (the supply's output). This calls for differential measuring techniques. While it is possible to spend many dollars on sophisticated differential voltmeters with built-in precision buck-out d-c sources, it is also possible to nearly approach the accuracy and sensitivity of such instruments with far less costly equipment (although at some sacrifice in convenience). When only an occasional measurement is made, the do-it-yourself approach may prove economically advantageous.

The basic concept in all differential measurements is to measure between a known and an unknown potential (Fig. 6.5).

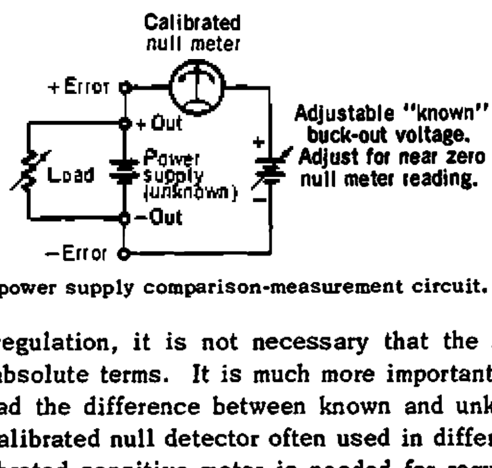


Fig. 6.5 Basic power supply comparison-measurement circuit.

When measuring regulation, it is not necessary that the *known* voltage be known in absolute terms. It is much more important to be able to accurately read the difference between known and unknown. In contrast to the uncalibrated null detector often used in differential measurements, a calibrated sensitive meter is needed for regulation measurements. There are many suitable meters available to most laboratories. The calibrated d-c amplifier in a good high-gain oscilloscope is perhaps the most commonly available. Also suitable, although somewhat less sensitive, are ordinary VOM and VTVMs on their lowest scales. A millivoltmeter, of course, is ideal.

(D) Load Regulation: A typical laboratory set-up is shown in Fig. 6.6. Here, one regulated power supply is used as the buck-out source. A d-c oscilloscope measures the change in the output of the supply being tested, while loading and line voltage are separately varied (Fig. 6.6).

As the load switch is activated, load current is drawn from the supply being tested through its internal impedance. The resulting change in its terminal voltage is detected by the oscilloscope as ΔE. Divide ΔE by the total voltage E as read on a reasonably calibrated voltmeter (1-2%) to obtain the regulation figure $\Delta E/E \times 100$ = percent load regulation. Similarly, $\Delta E/I$ equals the d-c *static* source impedance.

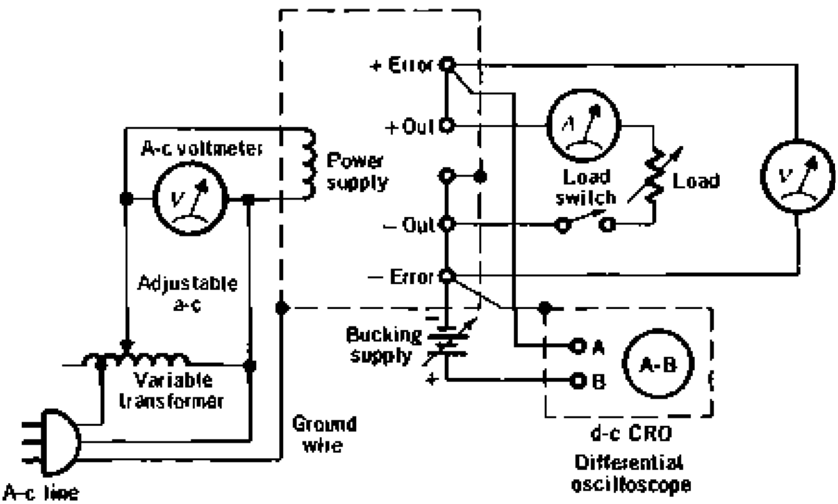


Fig. 6.6 Typical laboratory setup for power supply voltage regulation measurements.

Depending on the isolation of the oscilloscope and the buck-out power supply, some ripple may be introduced by the floating oscilloscope. This can be minimized by careful orientation of the line plugs and grounding. All grounds should be made at a single point, as shown. When testing high voltage power supplies, special precautions must be observed in as much as the oscilloscope case will almost certainly be at high potential from ground. If an oscilloscope with differential input is available, the differential feature should be used with the case of the instrument grounded for safety. Instruments designed as null detectors (millivoltmeters, and other high sensitivity voltmeters) are usually designed either for floating or differential connection. It will sometimes be helpful to use RC networks across the null detectors to minimize their response to spurious ripple pickup. A large amount of filtering is permissible since we are only interested in the d-c (steady-state) response. (Don't forget to remove the filters when using the oscilloscope for ripple measurements.)

For the user faced with the testing of large numbers of d-c power supplies, a specialized instrument called a regulation analyzer is available. Kepco, Inc. manufactures a very fine instrument called the Model 901D Regulation Analyzer. An analyzer consists of a calibrated, digitally adjusted divider with a null millivoltmeter calibrated directly in percentage regulation. An *rms* reading ripple meter is also included in the 901D.

(E) Line Regulation: Line regulation measurements are made using much the same set-up as for load regulation. The only additional equipment needed is a means for varying the primary a-c line voltage and an a-c voltmeter to accurately read the input voltage to the power supply. A variable auto-transformer with a good a-c voltmeter of the iron vane-type are suitable. Be certain that the variable auto-transformer is capable of handling the full load demands of the power supply being tested. Measure line regulation by observing the change in d-c output on the calibrated millivoltmeter as the primary a-c line voltage is varied throughout the prescribed limits. $\Delta E/E \times 100$ equals the percentage line regulation.

(F) Ripple: Ripple measurements can be made either with a true *rms* voltmeter or with an oscilloscope. In the latter case, a peak-to-peak ripple measurement is actually obtained. Translating this peak-to-peak reading to true *rms* is inexact owing to the nonsinusoidal nature of the ripple in a feedback regulated supply. The peak-to-peak ripple and noise will be from two to three times higher than the reading of a true *rms* voltmeter. Low energy rectifier switching impulses, if present, will not be detected by any meter arrangement, but may be observed on a high speed oscilloscope.

To be complete, each regulation and ripple measurement should include the effects of all other possible parameter settings; that is, load regulation should be measured for all values of line voltage. In practice, it is usually sufficient to measure load regulation at the minimum line, and then again at maximum line voltage. Similarly, line regulation should be measured twice: once with no load, and once again with full load. A ripple measurement should be made for each of the above conditions. For adjustable voltage power supplies, the measurements are repeated at full voltage, half output, and 10% output voltage.

Because it is desirable to specify the members of a given power supply group or family with a common set of numerical parameters, regulation is often specified as a percentage of maximum output, rather than as an absolute number. This leads to a specifying problem when variable output power supplies are involved. In particular, the percentage specification increases without bound as the output is reduced toward zero. Since, like all feedback controlled systems, the regulated power supply requires some minimum error signal in order to function, and since (especially for load regulation) not *all* of the series wiring contact and connector resistances can be fully compensated even by the most careful four-terminal network design, there must be some minimum absolute change in the output. This is termed the minimum ΔV (or ΔI, as the case may be). It is usually stated in the specification by the phrase: x percentage change or y volts, *whichever is greater.* y is the minimum ΔV. The minimum ΔV gives the user a convenient way to determine the performance that he can expect from a power supply at other than its maximum voltage setting. By dividing the stated ΔV by the rated percentage regulation (converted to decimal form), the minimum voltage for which the specified holds can easily be determined. For example: a supply rated 0.01% or 1 millivolt, whichever is greater, delivers 0.01% regulation down to 1 mv/0.01% = $10^{-3}/10^{-4}$ = 10V. Below 10 volts, the 1 millivolt figure predominates and would be used by itself to determine the regulation percentage; that is, at 1V output, the regulation would be 0.1%. Graphically, this interplay between specifications can be illustrated as shown by Figs. 6.7a-b.

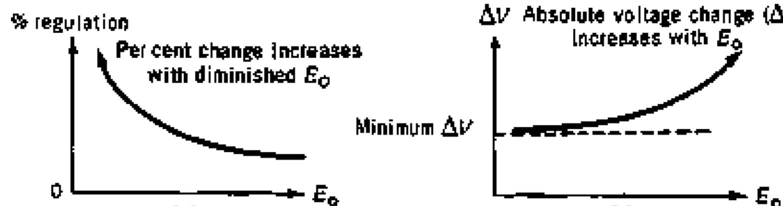


Fig. 6.7

Measurement at the 100%, 50% and 10% output voltage settings permits a ready evaluation of the important minimum ΔV figure. Similar qualifiers (ΔI) are used in specifying current regulation as in the *typical test report* of Fig. 6.8.

6.2 CURRENT REGULATION

The current regulated performance of modern precision power supplies is measured by means of techniques quite

Model________________________ Date Tested________________________

Serial________________________ Tester________________________

TEST RESULTS

OUTPUT Parameter and Value	LOAD REGULATION		LINE REGULATION		RIPPLE			
	Min. Line __V	Max. Line __V	Min. Load __	Max. Load __	Min. Line Max. Load		Max. Line Min. Load	
V I Circle one	Specification	Specification	Specification	Specification	Specification			
					RMS	P-P	RMS	P-P

Fig. 6.8 Typical power supply test report form.

analogous to those used for voltage measurements. In fact, since most instrumentation is voltage sensitive (or is calibrated in terms of voltage), current is usually measured by converting to a voltage by means of sensing resistors. The sample voltage is then compared to a known, stable reference, and the difference – or change – read on a calibrated null detector or sensitive milli/micro voltmeter.

Two complications are introduced by this process. The most serious is that current is not being read directly, but rather is only one part of the product of current and sensing resistance. Variations in the sensing resistance simply cannot be separated from the current variations under observation. Precautions to be observed in the selection of a current sensing resistor have been treated in the chapter on current generation. In general, the same precautions apply to the sampling resistor used in measuring current regulation; that is, they should be amply derated, low temperature coefficient laboratory grade resistors. The second complication is the fact that a very small sampling voltage is usually employed (to avoid detracting from the available compliance, and to avoid excessive dissipation losses). This means that much more sensitive instrumentation is needed for current regulation measurements.

If a 1 volt sample is employed (which is typical), measurement of 0.01% regulation requires the ability to measure 0.1 millivolts with reasonable accuracy. A 0.001% change involves as little as 10 microvolts of d-c level shift – a signal which may be difficult to recover across a hot sensing resistor, passing perhaps a hundred amperes or so, and dissipating 100 watts in the process.

Really accurate measurements require extraordinary precautions, including a large volume, constant temperature oil bath for precision 5-20 PPM sensing resistors, four-terminal network connections to the resistor, an ultra-stable, nulling source (to provide the buck-out voltage), and an accurate microvoltmeter to detect and measure the resulting voltage changes.

The nature of current sources and some peculiarities related to the way current is controlled in the typical power supply dictate some additional measuring precautions when dealing with current regulators.

Of first importance is a need to avoid any sort of shunt conductance across the current regulator's terminals. This is the dual of the often stressed precaution concerning series resistances in *voltage regulators.* Unfortunately, it is easy to forget that a voltmeter, for example, is just such a prohibited shunt conductance. A 20,000 ohm per volt instrument draws 0-50 microamperes full scale.

Such a VOM, reading the compliance voltage of a 100 milliampere power supply, introduces a 0.05% regulation error which can be most significant when you are trying to measure 0.01% regulation. Fig. 6.9 illustrates how this happens.

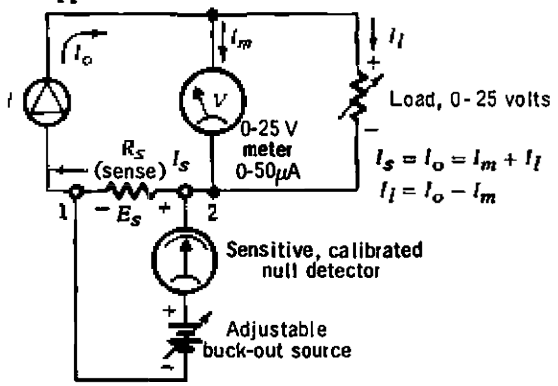


Fig. 6.9 Basic current regulation measuring circuit showing the effect of an output voltmeter.

The degradation comes about because the power supply regulates only its terminal current consisting of the load plus meter currents. The sum is maintained constant, so that if the meter current varies (as the voltage swings from zero to full scale), then the measured load current must change by an equal (but opposite amount).

Locating the voltmeter at (2) instead of (1) side-steps the problem as far as the measuring process is concerned, but the load current regulation is still degraded, even though not detected. One solution is to measure the compliance voltage at a point within the power supply – in front of the current sensing (regulation) resistor. The drawback to this is that the voltage has a built-in error by the amount of the voltage drop across the sensing resistors. A better solution is to use a VTVM or other very high impedance voltmeter – or to forego voltage measurements altogether. The current drawn by the panel voltmeter of Kepco automatic crossover power supplies has been compensated for optimum current regulation.

(A) Dynamic System Behavior: Because the typical power supply is really only a static current regulator (having an output capacitor which lowers the output impedance at any frequency above d-c) precautions are required to avoid unfortunate dynamic situations. It is necessary, for example, to remember that a current source, operating into a large load resistance (at a relatively high compliance voltage) has considerable energy stored in its output filter capacitor; energy

that can destroy a sensitive microammeter should the load be switched to a lower resistance, or shorted through the meter. Fig. 6.10 illustrates the proper way to switch a load through its range.

It should be realized that the rate of discharge of the filter capacitor depends on how much resistance remains at the lowest value of the load (zero in a completely shorted load). Precautions should always be taken to prevent the discharge current from affecting sensitive load components. When the load switch is opened, the voltage will rise to a new compliance voltage = $I_o\ (R_{l_1} + R_{l_2})$. First, however, the supply must charge its filter capacitor. This means a momentary dip in load current until the voltage reaches its final value. If the output current of the power supply is set to a very small value (relative to the size of the capacitor), the charging ramp may be quite slow and easily observed.

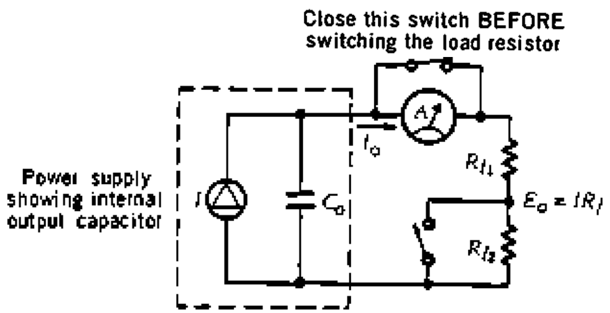


Fig. 6.10 Precautions when load switching a current regulator.

The instantaneous current changes that occur whenever the load resistance varies and the necessarily slow response (because of the output capacitor), can pose some problems in current measurements when unstable resistors or rheostats are used as the load. In particular, when rheostats are used, their brush noise appears as a small, variable resistance in series with the load. This results in a continuously changing compliance voltage (the product of current setting and load resistance). This, in turn, produces a randomly varying current in the filter capacitor (the a-c component generated by the noise, varying load). This current has the same effect that a shunt conductance would have: it causes the measured load current to vary and this, of course, makes it very difficult to observe the real load regulation.

In general, stable, fixed loading resistors are required for current measurements. An electronic variable load may be constructed, using power supplies described in Sect. 7.8, to produce a stable yet adjustable load. Such a load would be very useful if extensive current measurements are to be made.

From the foregoing discussion, it will be seen that the voltage regulator-turned, current generator suffers a severe nonideality by the presence of its necessary output filter-capacitance. In some power supplies equipped with a HS (high speed) regulator or regulation option, the high-speed characteristics are achieved by eliminating most or all of the output filter-capacitance. Such power supplies have much improved dynamic characteristics as current regulators. They should be considered whenever transient loading conditions are anticipated.

Current ripple is another parameter which requires exceptional care in measurement. One must resist the natural urge to measure current ripple by connecting an a-c voltmeter or oscilloscope across the output terminals of the supply. Such a measurement is, of course, meaningless – unless referred to the value of the load resistance at that instant. A power supply producing 1 milliampere of current ripple will read 1 millivolt, 1 volt or 10 volts *voltage* ripple, depending on whether the load resistance happens to be 1 ohm, 1K or 10K. The proper procedure, of course, is to measure the voltage across the same sensing resistor used for current regulation measurements, converting the observed voltage to current using Ohm's Law.

Figure 6.11 shows a typical equipment set-up for current regulation and ripple measurements. One special feature that differs from the equivalent voltage set-up is the requirement for a high-impedance null detector or potentiometer measuring technique. This is necessary to avoid drawing any current from the sensing resistor R_S.

(B) Procedures:

1) *Load regulation* – Observe the change in voltage across the the current sensing resistor as the load compliance voltage is switched from 0 to the maximum voltage. Allow the current to reach equilibrium after each change in load voltage. Protect instrument from current transients as the output capacitor is discharged. Load regulation is computed as $\Delta E_S/E_S \times 100 =$ percent load regulation. (Note that since a dimensionless percentage is computed, it is not necessary to convert the observed voltage to current.)

2) *Line regulation* – Observe the change in voltage across the current sensing resistor as the primary a-c voltage is varied throughout its prescribed limits. $\Delta E_S/E_S \times 100$ is the percentage line regulation.

3) *Ripple* – Measure the a-c voltage e_S across the sensing resistor, converting the measured voltage by dividing e_s by R_S. The ripple is usually specified in *rms* milliamperes, which requires that either a true *rms* reading meter be employed, or that peak-to-peak readings, as made on the oscilloscope, be converted to *rms*. For nonsinusoidal waveforms usually encountered, the equivalent *rms* current is approximately ⅓ the observed peak-to-peak signal.

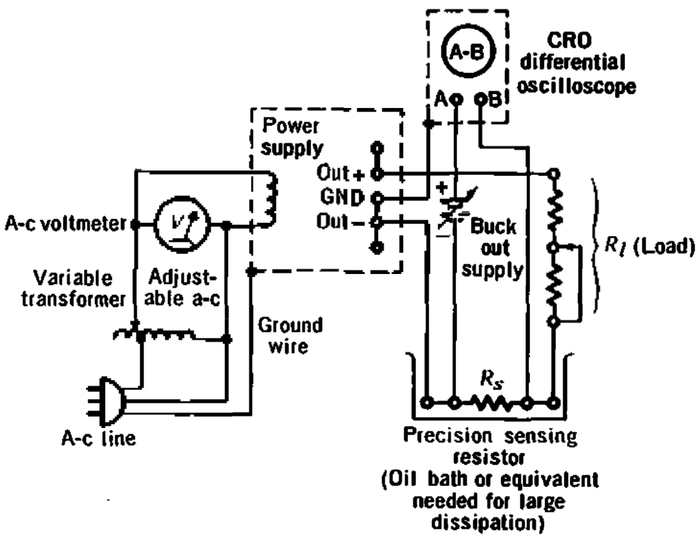


Fig. 6.11 Typical laboratory setup for power supply current regulation measurement.

The various current regulation and ripple measurements should be repeated for at least three points in the output range. Typically: 100%, 50% and 10% of the current rating.

For external sensing circuits, the value of the sensing resistor used to regulate current should be selected as necessary to maintain *at least* the specified voltage sample.

Load regulation measurements should be made both at the minimum and maximum line voltage and, conversely, line regulation measurements should be made for minimum and maximum load compliance voltage. A ripple measurement should be made for each of the above conditions.

When automatic crossover or current limiting power supplies are tested, care must be exercised to make certain that the power supply remains in the intended regulation mode – fully – for all loading conditions. In voltage regulation measurements, for example, it is relatively easy to run the power supply into either its current limiting or its current regulation modes; the latter, if the power supply is equipped with automatic crossover. If this goes unnoticed, the change in voltage will appear as a poor regulation result. Similarly, when automatic crossover supplies are measured in their current regulation mode, the load resistance range must be restricted so that the output does not crossover back into a voltage limiting mode. Supplies equipped with VIX lights and signals make it easy to stay within the desired mode.

Kepco regulated power supplies are designed to meet and exceed their regulation specification by an exceptionally wide margin. Moreover, most failure conditions will result in a very wide variance from the published performance specifications. For these reasons, should a test reveal that regulation is "just a little out of spec," be suspicious of the test set-up. Recheck the four-terminal networks to be sure there is no stray series or shunt loading and make certain that the power supply's current limit (or voltage limit) setting is well above the maximum load current (or compliance voltage). A quick check of the ripple voltage at the output terminals of the power supply can tell a lot about its behavior. Normally, the ripple is measured in fractions of a millivolt (for feedback voltage regulators), and consists of a fair amount of 120, 240 cycles and higher harmonics. 60 cycle components will likely be pick-up, possibly introduced through the grounding. If a supply is loaded past its current limit point, it will first be detected as a sudden, sharp increase in the ripple amplitude. Ripple components not harmonically related to the line frequency are cause for concern, as they may indicate induced voltage (as from nearby oscillators) or phase-gain instability within the power supply.

6.3 OUTPUT IMPEDANCE

As has been shown, a load regulation specification for voltage regulated d-c power supplies can be presented in terms of the internal source impedance, in ohms. If the load current is made to vary sinusoidally at an increasing frequency, the value of the load regulation will be observed to change also because the power supply source impedance is a variable function of load frequency. The output impedance versus frequency specifications for a power supply is a way of stating the load regulation for a spectrum of possible load frequencies.

There are a number of interrelated factors involved in the composite output impedance. They are:

1) Amplifier gain

2) Output capacitance

3) Actual internal resistances, wire, etc.

4) Output circuit inductance

Amplifier gain is mainly important at the lower frequencies, and is the most important factor at d-c (assuming proper error sensing techniques). The output capacitor is mainly effective in the mid-frequency region, although there is considerable overlap with the the amplifier's gain-frequency characteristic. At high frequencies (above the amplifier's cutoff), the series wiring inductance plays the major role in determining output impedance. Knowing the equivalent output inductance permits the source impedance to be calculated for all higher frequencies (Fig. 6.12).

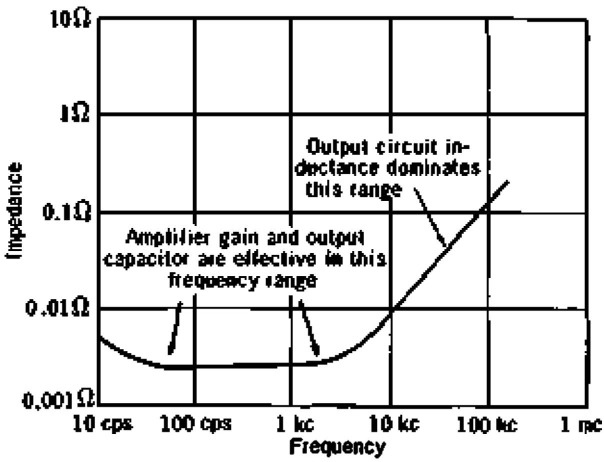


Fig. 6.12 Typical power supply output impedance plot on log-log paper.

(A) Impedance Measurements: Several techniques prevail for the measurement of power supply output impedance versus frequency. All involve some method of inducing a sinusoidal current variation

in the output circuit of the power supply. An oscilloscope is used to measure the resultant a-c voltage across the supply's terminals. Dividing this voltage by the driving current gives the output impedance.

One way to vary the load current is to use a sinusoidally modulated electronic load. A precision, noninductive sensing resistor is used to measure the peak-to-peak current using a dual beam oscilloscope or switch at the input of a single beam scope (Fig. 6.13).

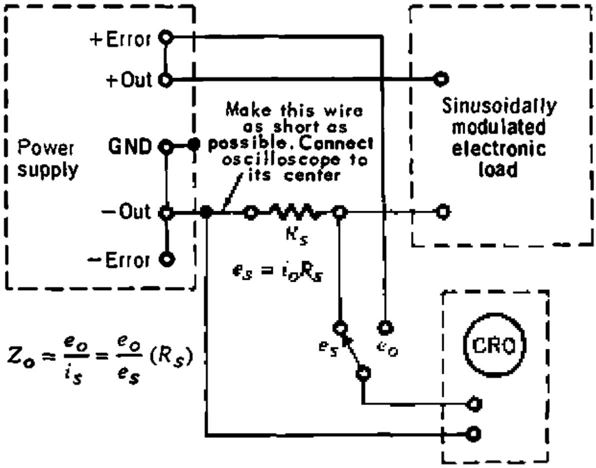


Fig. 6.13 Output impedance measured, using a modulated electronic load.

An alternate procedure employs a power amplifier (Hi-Fi amplifier), together with a fixed load resistance to modulate the output current. In this circuit, a large capacitor is required to keep the d-c out of the amplifier's output transformer secondary. This capacitor must pass the a-c load current and be able to support the d-c voltage. A-c type oil capacitors, such as used for power factor correction, are suitable. A total of several hundred microfarads is required for sufficient low-frequency response — and to minimize tilt for transient square-wave measurements. A resistor in series with the capacitor provides a matching load for the a-c amplifier which should be used on its lowest output impedance tap — highest current (Fig. 6.14a).

A preload resistance is also used to assure that the net current in the power supply is always in the proper direction. A special current sensing resistor is placed in series with the power supply and is positioned physically very close to the output terminals of the power supply.

The most important characteristic of this sensing resistor is its

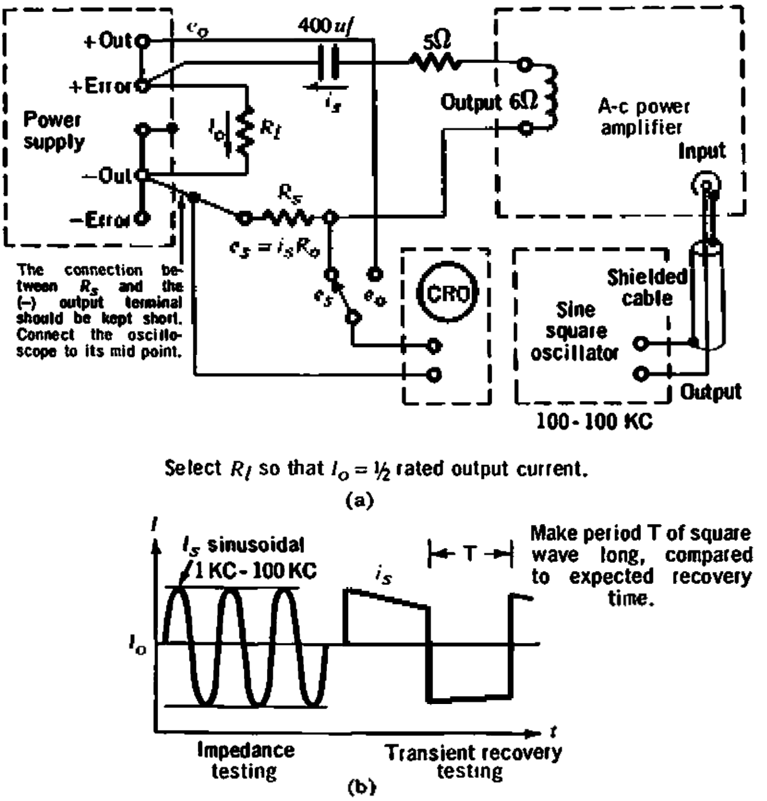
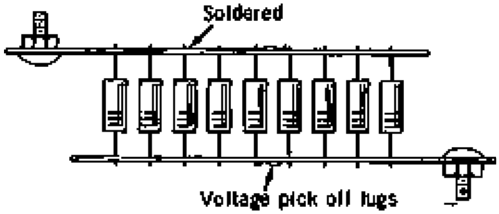


Fig. 6.14 Circuit and waveforms for a-c impedance testing and measurement of the transient recovery time.

non-inductive construction. A suggested method of construction is shown in Fig. 6.15.

The series output inductance of a well-constructed power supply will be measured in microhenries or even tenths of a microhenry. It is, therefore, very important that the sensing resistor contribute an inductance much less than 0.1 μh. Because the impedance is computed by dividing voltage response by the current drive, any inductance in the sensing circuit appears in the denominator of the equation and will cause the output impedance to appear smaller than it actually is. Since the connection between the output terminal of the power supply and the current sensing resistor is the common terminal of the oscilloscope, the resistor must be placed physically as close as possible to the power supply's output terminal, and be connected with the shortest possible wire to minimize the inductance. In addition, the connection to the oscilloscope should be placed at the physical mid-point of the connecting wire. In this way, approximately equal amounts of inductance will appear in the numerator and denominator, cancelling one another.



27 parallel 2.7Ω 1 watt carbon resistors sandwiched between two strips of tinned copper.

Resistance: 0.10Ω (measure with a bridge)

Inductance: < 0.01 μh

Fig. 6.15 Minimum inductance construction for a sensing resistor to measure a-c impedance.

The procedure for measuring impedance, using the oscillator-power amplifier circuit, is to set the power supply to approximately half voltage (observing the voltage limits of the blocking capacitor). Load the output to one-half rated current. Adjust the a-c amplifier output while monitoring the voltage across the current sensing resistor for the maximum peak-to-peak amplitude before distortion occurs. Do not permit the output amplitude to exceed the maximum current rating of the supply, nor pass below zero (reverse current). Refer to Fig. 6.14b.

The actual amplitude of the current is not important – so long as it produces a reasonably large response waveform across the voltage terminals. At low frequencies, this may be difficult to do, particularly if the power supply has an exceptionally low output impedance. There should be relatively little difficulty in obtaining an observable response for frequencies above 1 kc; the current will probably have to be reduced to prevent distortion.

When a satisfactory current waveform has been established, switch the oscilloscope across the output terminals of the oscilloscope and measure the voltage response waveform. Substitute the measured values into the equation to determine Z_o(the output impedance):

$$Z_o = \frac{e_o R_s}{e_s}$$

If R_s has been made an easy number, like 0.1 ohm, then the ratio e_o/e_s is simply measured and multiplied by the factor 0.1.

The measurement is repeated at various frequencies in the range 1 kc to 100 kc, and the result plotted on log-log graph paper. The slope of the impedance plot near 100 kc can be used to determine the output inductance. Figure 6.16 plots the slope of some ideal inductors versus frequency and impedance on the impedance chart reproduced in the Appendix.

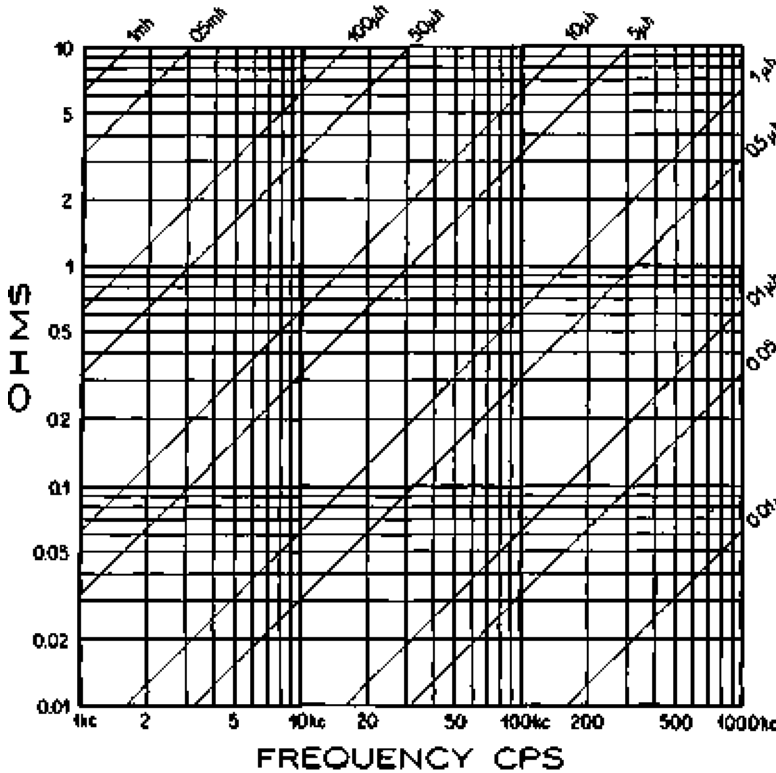


Fig. 6.16 Inductive reactance versus frequency.

A cathode ray oscilloscope is the preferred way to monitor the a-c voltage in an impedance measurement. This is because the eye can serve as a fairly selective a-c filter – to reject 60 and 120 cycle ripple frequencies which are not significant. Measurements made on wide band a-c voltmeters are often erroneous because the 60-120 cycle amplitude all but masks the minute voltage response of a low impedance voltage regulator. If a selectively filtered oscilloscope plug-in like the Tektronix 53/54B Plug-in unit is available, the signals of interest can be viewed without interference.

The output impedance of a power supply is most significant as a method of examining the phase-gain characteristic of the feedback control loop. The shape of the curve is of greatest interest. The impedance plot should be a smooth curve – no short bumps or dips anywhere. Such irregularities, should they be observed, particularly in the region where the curve tilts up (around 10-100 kc), are indicative of potential instability because of a misadjusted lag network in the power supply's amplifier. Proper adjustment of the phase-gain is indicated by a smooth impedance plot.

6.4 TRANSIENT RECOVERY TIME

The *transient recovery time* is defined as the time required for the voltage – of a voltage regulated power supply – to return to within the tolerance band established by the d-c regulation specification (see Appendix, Kepco Glossary of Power Supply Terms).

The amount of excursion outside of the d-c specification band is governed by the a-c regulation specification – the output impedance versus frequency. Since the impedance is very much a function of frequency, it follows that the peak transient excursion will depend entirely upon the rate of change of the load current. Usually, the rate of change is so fast that reference can be made directly to the inductive component of the output impedance. The voltage peak $v = L di/dt$ equals the inductance multiplied by the time rate of change of current. The relatively large error signal generated by this excursion will cause a vigorous response in a feedback regulator. The shape of this response resembles the universal damping curves for the transient response of any feedback system. A trade-off in design is sought between the sluggish response of an overdamped system, and the oscillatory response which results from underdamping. The optimum solution is usually found to be an adjustment which places the system somewhat on the oscillatory side of critical damping with perhaps 2 or 3 cycles of over and undershoot quickly decaying within the regulation tolerance band (Fig. 6.17).

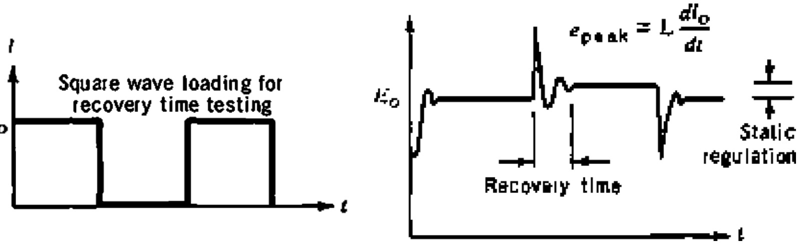


Fig. 6.17 Transient recovery waveforms.

The transient recovery time can be observed and measured using the same equipment and techniques described for impedance measurements, except that the square-wave generator is substituted for the sinusoidal oscillator. The *RC* coupling circuit, between power amplifier and power supply test sample (Fig. 6.14), will introduce some tilt in a low-frequency square-wave. By making the time constant long, and concentrating the oscilloscope display on the first 5-100 microseconds, the tilt will introduce a negligibly small error. Both the load-on and load-off transient characteristic can be observed by this technique. A square-wave frequency between 100 and 200 cps will be convenient for most purposes.

The current sensing resistor R_S is used to establish the amplitude of the current step, but does not otherwise enter into the measurements. It is necessary that the power amplifier used in these measurements be capable of a 1-5 microsecond rise time – without significant overshoot – in order to avoid masking effects in measurement.

Other techniques are sometimes used for transient and a-c impedance measurements. When measuring slow responding supplies such as sampled data types, SCR-only power supplies, or Flux-O-Tran (PR, PRM) regulators, the square-wave generator-power amplifier method requires a very low-frequency repetition rate in order to permit the response waveform to complete before the next transient occurs. This calls for an impossibly long *RC* time constant in the coupling network. To measure the transient recovery time of such power supplies, modulated electronic loads, electronic switches, or even large knife action switches are used to interrupt the load.

The above discussion is strictly limited to voltage regulator behavior. The response in current regulated mode is quite a different story. It is not normally specified in the literature because it is not an intrinsic power supply parameter; rather, it is a complex time expression involving the current setting and the load circuit. In current mode, the recovery time would be defined as the time required for the output current to return to within the regulation band for static current regulation following a step load resistance change. Unfortunately, this response depends on the rate of output *voltage* change. The process by which current is regulated calls for the output *voltage* change – in response to a load resistance change – in order to maintain a constant ratio $E_{load}/R_{load} = I$ constant. As we have already seen, the terminal voltage of a standard power supply – whether it regulates current or voltage – changes at a relatively slow pace. [The HS (high speed) regulators, of course, respond more quickly.] The rate of change *dv/dt* is governed by the ratio *I/C*. The slope of the response is a direct linear function of the output current and is an inverse function of the output filter capacitance. While the time function is difficult to express as a simple number, independent of load, (as it is in voltage mode), the *slope* of the response (in volts per second) is easily observed and measured for particular loads. This is treated at greater length along with the subject of programming speed in Chap. 9.

6.5 TEMPERATURE COEFFICIENT

The temperature coefficient (TC) of a regulated power supply is a measure of its *thermal regulation* or the degree of independence between output (voltage or current) and its environmental temperature. The TC is every bit as important as line or load regulation and, in fact, it is often the *largest* single factor in determining the long term constancy of output.

As an example, consider a voltage regulator featuring 0.01% line regulation, 0.01% load regulation, and temperature coefficient of 0.01% per °C. In the typical application, the line voltage and load rarely vary throughout their possible extremes – and hardly ever at the same time. Yet, the supply is more than capable of meeting the challenge and will vary no more than 0.02% under the worst conditions. On the other hand, it is not at all uncommon for the environmental temperature to change by 5° or 10°C. This would result in a 0.05% - 0.1% output change, quite a bit larger than the line and load regulation taken together.

This characteristic is likely to be of major concern when a power supply is being used as an absolute voltage reference source (an application not uncommon for modern digitally programmed supplies). In such application, the equipment should be protected from drafts, and the direct radiation of heat from adjacent equipment.

Temperature coefficient is measured by permitting the power supply to stabilize at one temperature, after which the temperature is abruptly changed to a new value. After permitting the power supply to restabilize at the new temperature, the resulting voltage (or current) change is divided by the causative temperature swing and expressed as a percentage of the output (percent change per °C).

The major equipment required is an environmental chamber capable of producing the desired temperature changes, plus a means for recording the output voltage of the power supply during the test. Continuous strip chart recordings are preferred to spot meter readings, since the recorder trace more accurately reveals the transient behavior during the temperature step, and permits an accurate evaluation of when stabilization has occurred. A millivolt recorder bucked against a stable reference source will yield the requisite sensitivity. Since thermal time constants tend to be relatively long (depending on the mass of the supply in question), a TC measurement may take an hour or more to complete. This places a burden of stability on the reference (buck-out) source. While a highly stable power supply could be used for this purpose – as is often done for regulation measurements – the possibility of systematic error is reduced by the used of a non-electronic reference. Mercury batteries have been found excellent for this purpose. When placed in a simple constant temperature air-bath, mercury reference batteries yield secondary cell performance. An excellent choice for this function is the Mallory type 303113, 8-cell reference battery. For less demanding measurements, a battery box made up of type RM 502R and type TR 134R batteries will give very good results. A small (22 ohm) resistor in series with the output terminals gives short-circuit protection and does not affect the voltage in potentiometric (zero current) application. For precise bucking, an extra cell, bridged by a 10 k, 10-turn potentiometer provides an adjustable output.

6.6 LONG TERM STABILITY

A power supply's long term stability (LTS) describes the residual changes in output after the known effects of line regulation, load regulation, and temperature coefficient have been accounted for. The remaining drift is observed for a stated period and given as a specified percentage of output, or as a maximum absolute quantity – whichever is greater. Kepco rates power supply stability over a period of 8 hours; this period was chosen because it represents a typical working day, and thus gives a more accurate representation of the practical stability than much longer or shorter periods.

The long term stability like the temperature coefficient of a modern zener-referenced power supply is mainly traceable to the quality of the reference diode against which error comparisons are made. This comes about because of improvements in the design and construction of solid-state d-c amplifiers which have nearly eliminated the amplifier as a major source of drift – with or without choppers.

For the measurements point of view, LTS and TC parameters are often difficult to separate, since both have about the same order of magnitude. The long term stability of a power supply rated 0.01% per 8 hours cannot be measured accurately unless the temperature coefficient (typically 0.01% per °C) is suppressed during the measurement.

To measure such long term stability with an accuracy of 1%, the environment temperature must be held constant to better than 0.01°C during the entire period of the measurement. To further reduce error, LTS measurements are usually conducted for several consecutive 8-hour periods, and the total drift divided by the number of periods. Temperature control apparatus capable of holding 0.01°C for 40 hours (while the test sample itself dissipates hundreds of watts) can be constructed using power supplies themselves in a feedback arrangement described in Chap. 8.

As in the case of TC measurements, a major problem for accurate results is a stable (reliable) reference. For long term measurements, it is helpful to enclose the reference source or battery in a constant temperature air-bath to eliminate the slow, periodic variations caused by plant air-conditioning/heating equipment in its day/night cycle. The air-bath itself is a miniature environmental chamber using power supply feedback arrangements. Temperature is maintained at about 30°C ± 0.01°C. Several mercury reference batteries and standard cells can be maintained and calibrated in this environment to serve as the null references for the recorders. As is the case for TC measurements, the sensitive potentiometric recorder is the basic tool employed in measuring the long term drift characteristic.

7 POWER SUPPLY INTERCONNECTIONS

This chapter will treat some of the many ways in which two or more d-c regulated power supplies can be connected together to extend their usefulness.

7.1 SERIES CONNECTIONS

(A) Voltage Regulator, Independent Control: The ordinary series connection is one of the simplest ways to interconnect voltage regulated power supplies. Voltage regulators can be series connected to increase the total output voltage to the sum of their individual voltage settings up to the maximum voltage limit allowed by the isolation voltage specification for each power supply. The maximum isolation voltage specifies the amount of *additional* voltage that can be connected between either terminal of the power supply and its grounded chassis, *not* including the power supply's own output voltage. This means, for example, that two 1000V supplies, each isolated for 1000 volts, *can* be interconnected for 2000 service.

The isolation specification derives from the voltage breakdown rating of the printed circuit boards, connectors, sockets, meter bezels, switches, transformer, and even the wiring insulation within the supply. There is also, usually, a capacitor connected from one output terminal to ground (sometimes with a series or parallel resistor) whose purpose it is to provide a low impedance a-c path from chassis to circuit (for shielding purposes), yet maintaining d-c isolation for grounding convenience. The working voltage rating of this capacitor is often instrumental in determining the isolation rating of the supply.

When dissimilar supplies are connected in series, it is desirable that wherever possible the power supply with the smallest isolation rating be "on the bottom" of the stack, that is, nearest ground. If, for example, a 325 volt supply with a 600 volt isolation rating is series connected with a 1500 volt power supply rated for 1000 volts of isolation, the 325 volt supply should be connected nearest to ground (Fig. 7.1).

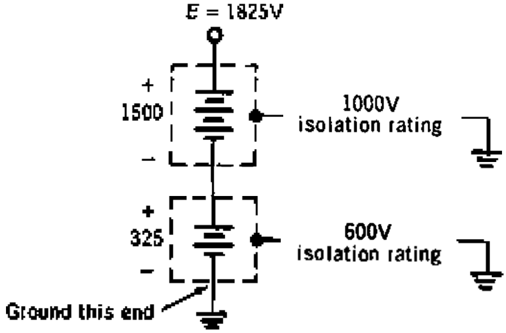


Fig. 7.1 When high voltage power supplies are connected in series, ground the one with the smallest isolation rating whenever possible.

When it is not possible to stay within the rated isolation voltage limits, a power supply can be "floated" with its chassis connected to one or the other output terminal, provided that an isolation power transformer is used, and precautions are taken to safeguard personnel from contact with the *hot* chassis.

Whenever regulated power supplies are series connected, it is a good idea to protect them against the possibility of a reverse potential. Such a momentary reversal could occur if, for example, the load were short circuited (Fig. 7.2).

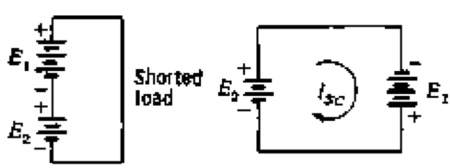


Fig. 7.2 Should the supplies be set to unequal voltages or should one be switched off, and the other not, reversal of the terminal polarity might occur.

Reverse polarity is undesirable because of the polarized components in the output circuit of a typical regulated power supply (e.g., the filter capacitors). The standard precaution is a pair of semiconductor diodes connected in the normally nonconducting direction across the output terminals of all power supplies in a series string (Fig. 7.3).

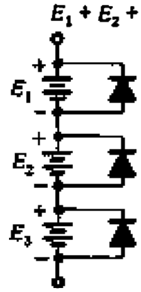


Fig. 7.3 Reverse diode protects a series string of power supplies.

These reverse diodes will conduct the moment a reverse potential appears, and will provide a safe path for the short-circuit current *around* the power supply. Each diode must have a PRV rating at least equal to the maximum supply voltage of the individual supply to which it is connected, and should be able to conduct the maximum short-circuit current indefinitely.

(B) Series Connection Automatic Crossover Power Supplies: When voltage/current regulated power supplies featuring automatic crossover between voltage regulation and current regulation are series connected (observing the foregoing precautions concerning reverse potential protection), an interesting "stair step" characteristic can be obtained. Figure 7.4 illustrates the characteristic for power supplies equipped with VIX automatic mode indicating lamps.

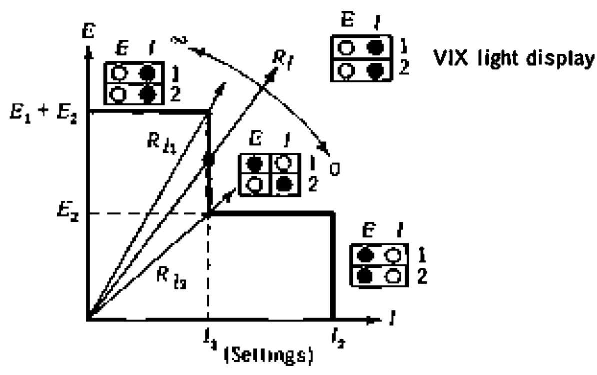


Fig. 7.4 The pair of lamps represent the VIX signal lights found on the front panel of a VIX power supply. When the left lamp is lighted, the power supply is a voltage regulator; when the right lamp is lighted, the power supply is a current regulator.

In the E, I characteristic, a load appears as a radial line, such as the ones marked R_l. For loads limited from R_{l1} to R_{l2}, it is possible to traverse the I_1 setting from a maximum voltage of $E_1 + E_2$ to a minimum of E_2; i.e., a current source with maximum and minimum voltage constraints. By suitably choosing the limits for R_l and the control settings, operation over any of the four available characteristics is possible. If the I_1 and I_2 settings are made equal, or nearly so, it is possible to operate as a current source with a voltage compliance equal to the sum of the voltages $E_1 + E_2$.

(C) Series Connections Voltage Regulation — Master-Slave Single Knob Control: From Chap. 2, it will be recalled that a power supply can be programmed by any arbitrary voltage of the proper polarity, provided that a suitable coupling resistor is chosen to generate the proper bridge current.

As shown in Fig. 7.5, another bridge controlled power supply designated the *master* can readily supply the bridge current for a second supply designated the *slave*.

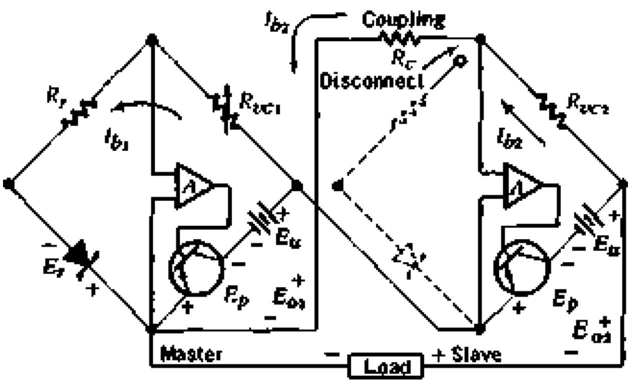


Fig. 7.5 Master slave series connection for the all-transistor power supplies (using PNP pass transistors).

If the coupling resistor is computed to generate the design value of I_{b2} (1 ma or 10 ma) for the *maximum* output of the master supply, the output of the two power supplies will be related by the formula $E_{o2} = (E_{o1}/R_C)(R_{vc2})$. The sum of $E_{o1} + E_{o2}$ appears across the load.

Figure 7.6 shows the same circuit, except for hybrid power supplies, whose vacuum tube pass element requires a reversal of the polarities. All-transistor designs employing NPN pass transistors use the same polarities as hybrid models (Figs. 7.5-6),

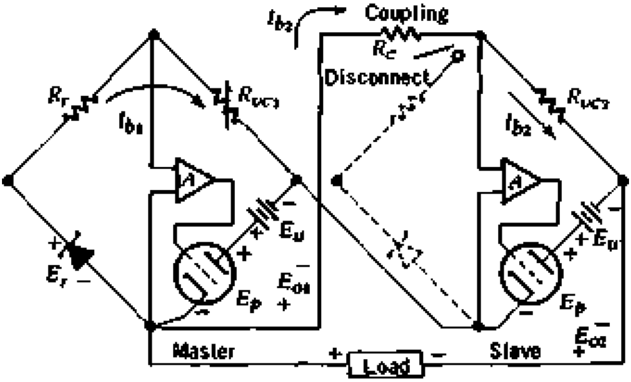


Fig. 7.6 Master-slave series connection for hybrid power supplies using a pass tube. The polarities are also correct for NPN transistors.

Consider the following calculations:

Example: We have 2 power supplies, each capable of 0-40V d-c, 0-500 ma, with a design bridge current of 1 ma. Connect in series using master-slave control for 0-80V, 0-500 ma.

Solution:

1) Select R_c (coupling) = 40V/1 ma = 40,000 ohms.

2) Disconnect the internal reference source and reference resistor in the slave following instructions supplied with the power supply.

3) Connect as shown in either Fig. 7.5 or Fig. 7.6, and adjust the voltage control of the slave for equal output voltage. The two power supplies will now track the master's voltage control being referred to a common reference.

(D) Complementary Tracking: An interesting variation on the master-slave series interconnection is a system of complementary connections which results when two separate loads are connected to each supply, as shown by Fig. 7.7.

Note that the individual load voltages E_{o1} and E_{o2} are controlled in unison by the voltage control of the master supply. The voltages can be set in any ratio, not necessarily equal, by appropriate variations in the coupling resistor R_c and the voltage control #2, R_{vc2}. Once the ratio is selected, the outputs can be adjusted in a constant ratio, using R_{vc1}. This particular master-slave programming technique can be extended so that the master supply drives several slave units simultaneously, as shown in Fig. 7.8.

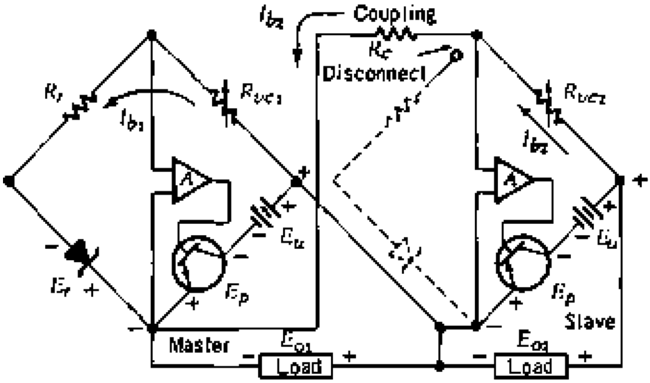


Fig. 7.7 The complementary connection, E_{O1} and E_{O2}, increase and decrease in unison in a ratio determined by the relative settings of R_{VC1} and R_{VC2}. R_{VC1} controls both outputs.

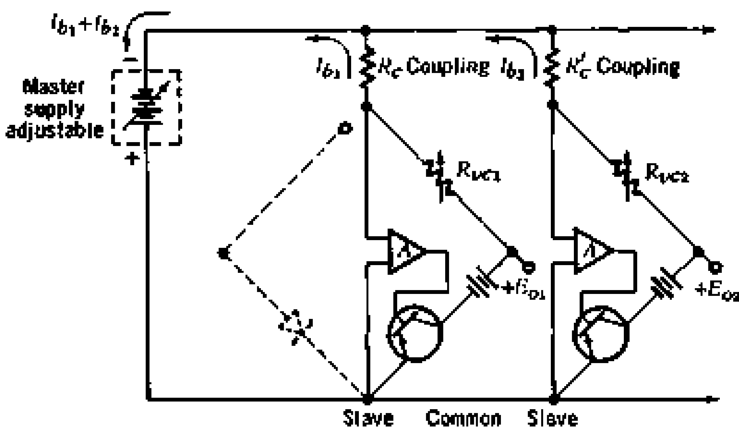


Fig. 7.8 Master-slave control in which a single master voltage (battery symbol) controls many slave bridges simultaneously.

(E) Series Connected Current Regulators: Current regulating power supplies cannot ordinarily be series connected, since in a series circuit, the terminal currents must be identical, precluding independent current determination (a part of the definition for current sources). Two current regulators set to different currents (no matter how minute the difference) simply cannot exist in series. Because the ideal current source has a very high internal resistance, the slightest difference in current settings causes a very large compliance

voltage, saturating one or the other of the supplies. To overcome this limitation, two different methods of series connections have been designed. One makes use of the self-determination ability of automatic crossover power supplies to limit the saturation to any desired voltage. The other method employs a master-slave interconnection in which only one power supply is set up as a current regulator. This method is described in Sect. 7.1(C).

(F) Series Connection Current Regulation — Master-Slave Control (Compliance Extension): The master-slave interconnection described in Sect. 7.1(C) can be made to have the characteristics of a current source, merely by connecting the *master* power supply as a current regulator. Since the slave supply simply duplicates the terminal voltage of the master, it too will behave as a current regulator — by imitation.

The *master* supply can either operate using *external* sensing or *internal* sensing if it is an automatic crossover power supply equipped with VIX (Figs. 7.9a-c).

Figure 7.9c shows a typical circuit drawn for external sensing to make its function clear.

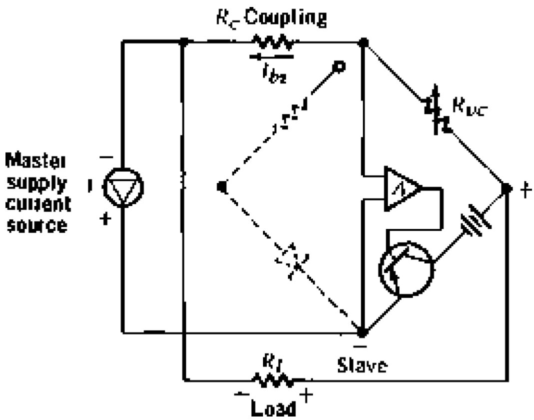


Fig. 7.9a Compliance extension for current regulators. The master and slave are in series across the load. The slave bridge is programmed by the voltage compliance of the master current source. The coupling resistor is a shunt load on the master which degrades its current regulation by the amount of I_{b2}. To avoid this loading, it is advisable to connect the coupling resistor *in front of* the current sensing resistor as shown in (c). Even if internal sensing is used for the master, access to the inside terminal of the current sensing resistor is still available via the power supply's terminal strip. (See appendix.)

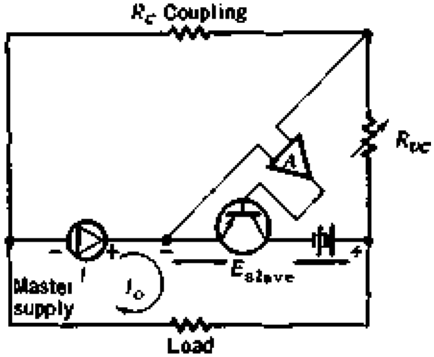


Fig. 7.9b Compliance extension. Same circuit, drawn so as to emphasize the series connection.

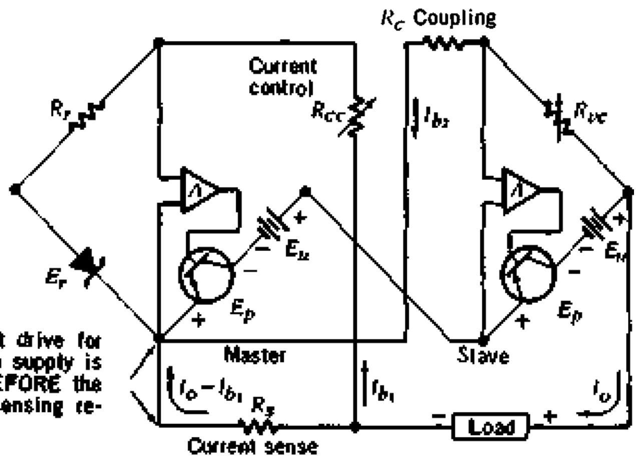


Fig. 7.9c Compliance extension with the sensing circuit for the master supply shown explicitly.

The circuit for hybrid or NPN-transistor power supplies is similar except that the polarities are reversed.

For the compliance extender, as for all master-slave interconnections, the reference section of the slave is disabled, and an external drive system is substituted. This time the drive is from a power supply operating as a current regulator. The drive originates with the compliance voltage of the *master*. Since the output voltage of the current regulator is proportional to the load resistance in such a fashion as to maintain the load current constant, and since the voltage from this driver is used to program the slave, which simply duplicates the variations of its master, the total series voltage behavior is also that of a current regulated supply, except now, of course, the compliance voltage range is the sum of the two supplies' capabilities.

The coupling resistor R_c is computed by dividing the maximum compliance voltage capability of the master supply, by the normal bridge current requirements of the slave supply.

7.2 PARALLEL CONNECTIONS

The problems associated with the parallel interconnection are just the dual of those discussed for series connection. Whereas voltage regulators could readily be series connected while current regulators require some special precautions, here it is the parallel connections of current regulators which is simplest, while voltage regulators require special treatment before they can be paralleled.

(A) Parallel Current Regulators: Any number of current regulators can be paralleled at will without the slightest precaution, except to note that the supplies should all have the same maximum compliance voltage rating (Fig. 7.10).

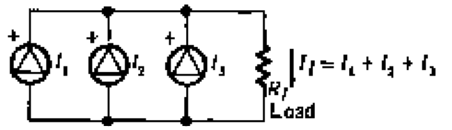


Fig. 7.10 Parallel current regulators.

If the voltage ratings are different, and if the load is open-circuited, the terminal voltage will rise to the maximum output of the highest voltage rated supply. If this is greater than the capabilities of one of the parallel partners, damage could result. It is a good idea to confine parallel operations to supplies of the same rating wherever possible. If this is impossible, the power supplies can be individually protected by means of a diode in series with each output (Fig. 7.11).

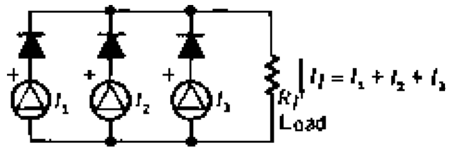


Fig. 7.11

The diode is normally conducting and must, therefore, be capable of withstanding the short-circuit current of its regulator. The diode PRV should be at least as large as the maximum open-circuit potential of the highest rated supply in the group.

(B) Parallel Connection – Automatic Crossover Power Supplies: When voltage/current regulated power supplies featuring automatic crossover between current regulation and voltage regulation are paralleled (observing the diode precaution if warranted), an interesting stair-step characteristic results which is just the dual of the one observed for series connection. Figure 7.12 illustrates this characteristic along with the VIX mode indicator light pattern that will be observed. Note that the maximum voltage is the highest setting of either power supply while the currents add along the horizontal axis. One precaution that should be observed is to always have at least 10% of the voltage control setting as the minimum voltage setting E_2.

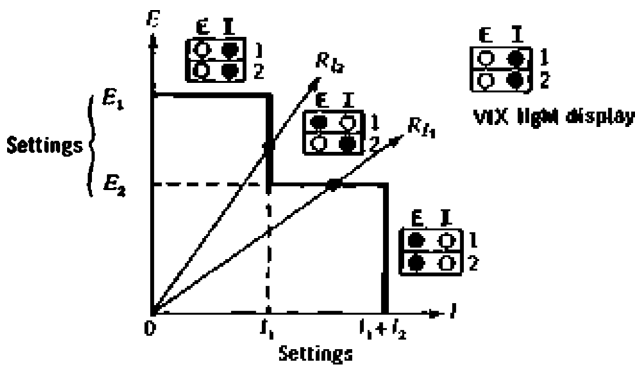


Fig. 7.12 VIX diagram shows the panel VIX light pattern that will be observed in the three major sections of the *E, I* characteristic.

The reason for this is that the voltage control "sees" the entire output terminal voltage since the parallel terminal voltage is controlled by the larger of the two voltage settings. The current through the smaller control setting, R_{VC2}, is E_1/R_{VC2}. We wish to prevent R_{VC2} from approaching zero in order to prevent burn-out of this component.

In Fig. 7.12, R_l is a radial line as shown, and may be made to vary in such a way as to intercept only a limited portion of the *E, I* characteristic, so as to generate either a constant voltage with a minimum or maximum current, load R_{l_1}; or a constant current with a minimum and maximum voltage, load R_{l_2}.

(C) Parallel Connection of Voltage Regulators Using a Master-Slave Configuration: When voltage regulators are paralleled, their terminal voltages are all, of necessity, identical. Since it is unlikely that independently adjusted supplies will have the same, identical terminal voltage before being paralleled, the minute differences that are bound to exist in even the most carefully adjusted supplies will inevitably cause excessively large circulating currents through the near zero source resistance of the good voltage regulator. This characteristic prohibits paralleling without recourse to one of the following:

1) An automatic crossover power supply, or one containing a good current limiter, can be paralleled by employing the self-limiting feature as has been described in Sect. 7.2(B).
2) Alternately, one of the two master-slave interconnections can be employed to assure the equal sharing of load current by causing the terminal voltages to be exactly equal.

Pass Element Drive

In the voltage regulated power supply, the output of a high-gain comparison amplifier drives the pass element with an amplified error, varying the conduction of the pass element in order to control the output voltage. If one comparison amplifier is made to drive two sets of pass elements, then both conduct equally, and if parallel, will generate the identical terminal voltage without a circulating current. It is generally sufficient to allow a 10% derating in the *total* paralleled current rating to account for any unbalance in the current sharing (Fig. 7.13).

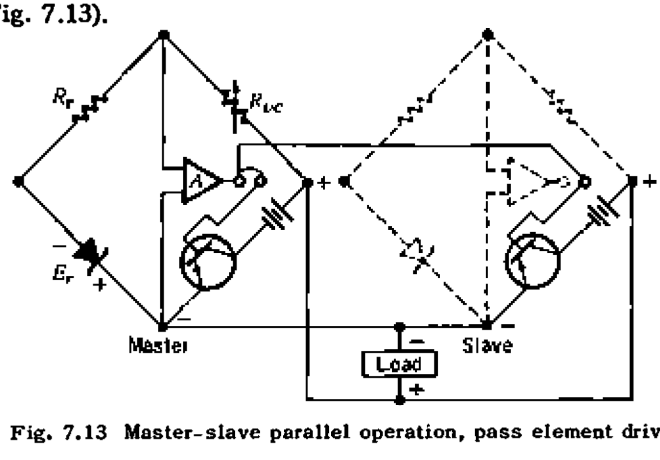


Fig. 7.13 Master-slave parallel operation, pass element drive.

To preserve the balance, it is helpful to make both sets of load wires of about equal length. Remote error sensing from the *master* may be employed if desired. It is also advisable to have both supplies operated from a common, switched, primary a-c source so that they can be turned on and off together.

The number of identical supplies that can be paralleled in this way depends on the amount of drive that can be obtained from the active comparison amplifier. There is always enough drive for at least two sets of pass elements (one more besides its own), and usually there is enough for three or four sets. It is not advisable, however, to depend on this kind of interconnection when many supplies are to be paralleled. Instead, the method known as *parallel programming* is recommended.

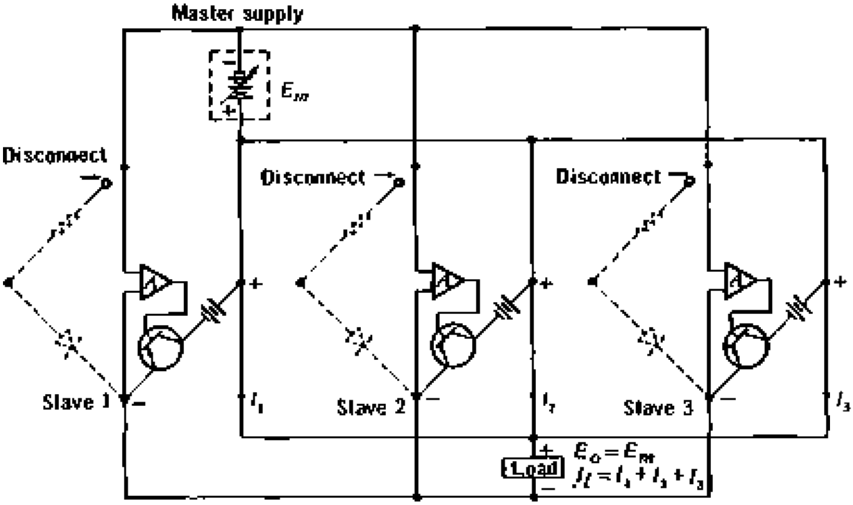


Fig. 7.14a Parallel programming. The master supply is a variable voltage source which programs each of the slave supplies to have identically the same output voltage connected in parallel. Their currents add through the load.

Parallel Programming

This interconnection employs a separate master control supply to drive all of the paralleled power supplies. The master control supply itself does not deliver any power to the load and need only be capable of generating the desired voltage. It may, therefore, be a relatively small, inexpensive supply. The basis for this method of control was developed in Chap. 2, where it was pointed out that the voltage drop across the voltage control (which is equal opposite and controls the output voltage of the supply) is not solely limited to the voltage drop across a resistor. Indeed, *any* source of voltage substituted for the potential drop across the *feedback* or voltage control terminals will effectively program the power supply as a unity gain amplifier whose input impedance is high, but whose output impedance is low, and is suitable for delivering substantial power to a load.

If several identical power supplies (any number of them) are connected as such unity gain amplifiers with their outputs in parallel, then a single *master control* supply will program them in unison for unrestricted addition of their current capabilities. Again, the total current of the paralleled supplies should be derated approximately 10% to account for the residual unbalance in the current sharing (Fig. 7.14).

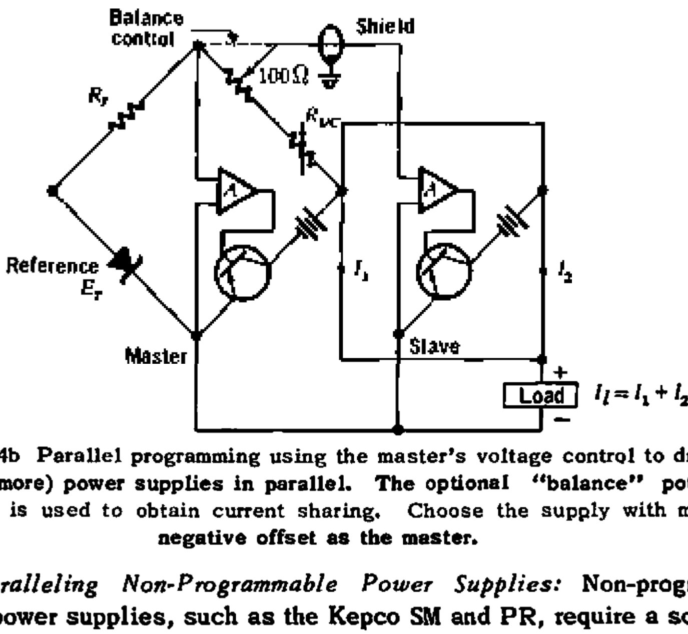
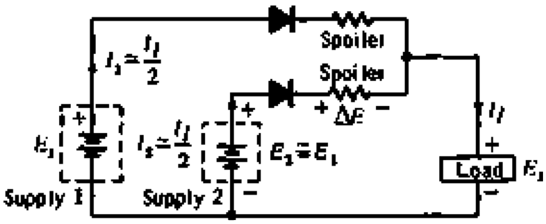


Fig. 7.14b Parallel programming using the master's voltage control to drive two (or more) power supplies in parallel. The optional "balance" potentiometer is used to obtain current sharing. Choose the supply with most negative offset as the master.

(D) Paralleling Non-Programmable Power Supplies: Non-programmable power supplies, such as the Kepco SM and PR, require a somewhat different approach for paralleling. None of the master-slave approaches will work because of the lack of programming capability. Such supplies can be paralleled, however, if the user is willing to sacrifice a little load regulation capability. Line regulation, ripple, and other capabilities would be unaffected. Simply stated, *spoiler* resistors would be used to *spoil* the low source resistance characteristic which makes for excellent load regulation, but also makes paralleling impossible (Fig. 7.15).



$I_l = I_1 + I_2 - 10\%$

Fig. 7.15 Pick spoilers to drop the maximum anticipated difference between the voltage settings, ΔE, with approximately one half of the load current.

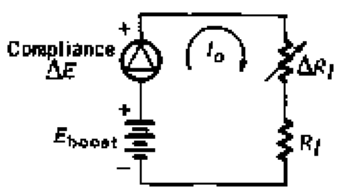
The purpose of the diodes in Fig. 7.15 is to prevent a reverse current flow back into whichever supply has the smaller voltage setting. The procedure for setting up this interconnection requires that the individual output voltages first be adjusted to as nearly equal settings as possible. Then, after connecting them in parallel through the diodes and spoiler resistors, touch up the voltage setting to get the best possible current sharing. The value of the spoiler resistor depends on how accurately you can set the individual voltages (the resolution of the controls, plus the amount of leeway allowed the operator). The effect of line regulation should also be taken into account. For example, to parallel a pair of 160 volt, 4 ampere power supplies whose voltage controls are capable of 0.1% resolution, and whose line regulation is 0.01%, we add the possible deviations, determining that there can be as much as 0.22% difference between the supplies. 0.22% of 160V d-c is 0.352 volts. A 0.18 ohm resistor will drop 0.36 volts at 2 amperes, so that each spoiler resistor needs to be about 0.18 ohms, or perhaps 0.2 ohms, just to be on the safe side. Of course, this leaves no room for any error in setting the respective voltages. It may be more convenient to pick a larger resistor, say 0.5 ohms, and allow a bit more margin for misadjustment. It is generally advisable to assume that there will be at least a 10% unbalance current, and reduce the rated available current by this amount.

The method of spoiler resistors is most useful in fixed load applications where the effects of the degraded load regulation are not so important. The most severe drawback is that the voltage is not variable; once set, it must be left set until it is readjusted all over again to a new value. As before, it is a good idea to turn all paralleled supplies on and off simultaneously, by means of a switched, common a-c power line.

PR and PRM power supplies, and other similar types which lack feedback regulators can easily be paralleled without the need for additional spoiler resistance. Their source resistance is generally not low enough to cause excessive circulating current. Like SM units, however, they must first be set to the same voltage, and adjusted for equal current sharing. A 10% allowance for unbalance currents is also a good idea. Parallel only identical models in the fixed voltage PRM module series.

7.3 SERIES BOOST

Another useful form of series interconnection results when a relatively uncontrolled power supply is used to boost the voltage control range of a more sophisticated supply into a higher voltage region. An example of this is illustrated in Fig. 7.16, in which the compliance range of a current regulated power supply is boosted to a higher voltage by a voltage regulated unit E_{boost}. A possible load resistor varying ΔR_l is to be driven by a current regulator. We have already seen (Chap. 4) how the dissipation requirements of a current regulator are greatly increased as it operates near zero voltage. The method shown provides a way to produce the necessary compliance $I_o(\Delta R_l)$ when compliance down to zero volts is not needed (Fig. 7.16).



$I_oR_l = E_{boost}$ and $I_o\Delta R_l = \Delta E$ compliance

Fig. 7.16 Series boost.

A simple numerical example will illustrate the possible advantages.

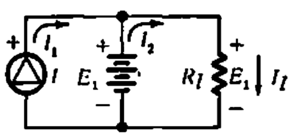
Example: We have a load resistance varying 200-210 ohms. We require a constant current of 1 ampere through the load.

Solution: The compliance required is only 10 volts in the region from 200 to 210 volts, and so a fixed 200 volt power supply in series with a 10 volt compliance current regulator satisfies all requirements. Both supplies must be rated for the desired 1 ampere of current flow. The pairing of the two supplies, as indicated, will often be much less costly than the purchase of a single 1 ampere current regulator with 0-210 volt compliance. In fact, current regulators with large compliance may be difficult, or very costly, to produce.

7.4 PARALLEL PADDING

The dual of the series boost connection is the parallel padding interconnection which can be used to extend the current range of a voltage regulator by paralleling it with a current source which functions to add a fixed amount of current into a load.

As shown in Fig. 7.17, the current in the load is simply I_1 and I_2, where I_1 is I, the current setting of the current source, while I_2 is $(E_1/R_l)-I_1$. The current source is needed whenever E_1/R_l exceeds the current capability of the voltage source. The voltage across the terminals of the load is fully regulated by the voltage source power supply.



$I_l = I_1 + I_2$

$I_1 = I$ (source)

$I_2 = (E_1/R_l) - I_1$

Fig. 7.17 Parallel padding.

Parallel current padding and series compliance boosting will probably be recognized as a form of the series/parallel interconnections described in Sect. 7.1(B) and 7.2(B). By limiting the operation to only a small portion of the stair-step diagram (Figs. 7.4 and 7.12), the interconnection can be made between power supplies that lack the automatic crossover feature.

7.5 VOLTAGE CORRECTOR

A somewhat more specialized form of the series boost configuration results when the controlled supply is connected as a voltage regulator (rather than as a current source). If feedback is applied around the entire series string, the fixed, or uncontrolled supply, E_{boost} simply becomes a part of the unregulated source, E_u, in the output circuit of the regulator. This particular configuration is called a voltage corrector because the regulator *corrects* for variations in the boost supply by causing its own terminal voltage to vary in an equal and opposite fashion.

A useful function for the corrector circuit is as a ripple corrector for a large, poorly regulated source. Subject to the amplitude frequency limits developed in Chap. 9, the corrector circuit will do a good job in opposing variations in the boost supply as can be seen in Fig. 7.18.

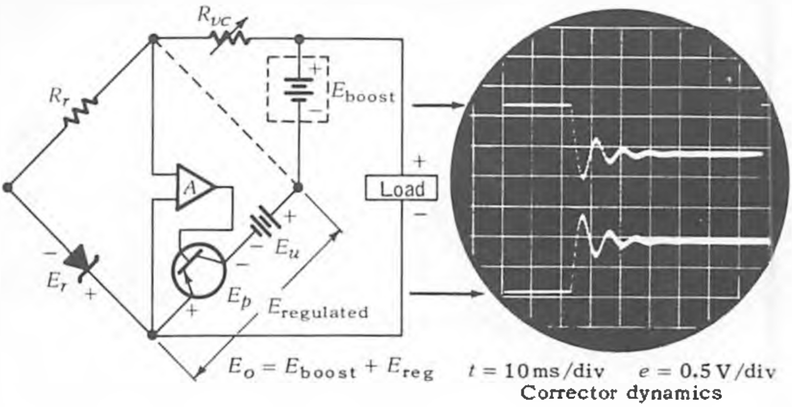


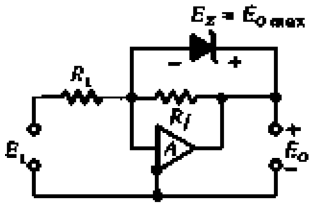
Fig. 7.18 Voltage corrector. *Caution:* the sum of $E_u + E_{boost}$ may exceed the voltage rating of the pass transistor E_p. If accidental short circuits are likely, and if E_{boost} is not the sort of power supply whose terminal voltage goes to zero when shorted, then E_p should be bridged by a power zener diode to provide over voltage protection.

7.6 VOLTAGE LIMITING

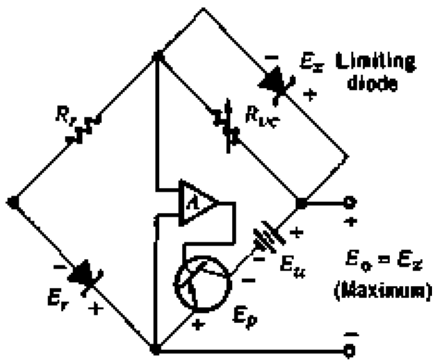
A form of overload (over voltage) limiting is a desirable feature for power supplies operating in the current regulation mode. Adjustable limiting is provided in all automatic crossover (VIX) power supplies, where the voltage control setting serves as the upper limit to the voltage compliance. If external error sensing is used to generate the current regulation, or if non-automatic crossover supplies are set up for current regulation, there is no limit on the voltage compliance. When such power supplies are overloaded by having their load circuit opened, the compliance voltage rises to the maximum unregulated voltage, E_u, which is usually a good bit higher than the normal maximum rating. When such power supplies drive voltage-sensitive loads, it may be desirable to limit the maximum compliance to some specific value, perhaps less than the design rating of the supply.

When a d-c regulated power supply is programmed operationally as in any of the control systems described in Chap. 8, a large error signal will drive the power supply into saturation until a balance can be established. A way of controlling the maximum saturation voltage is highly desirable.

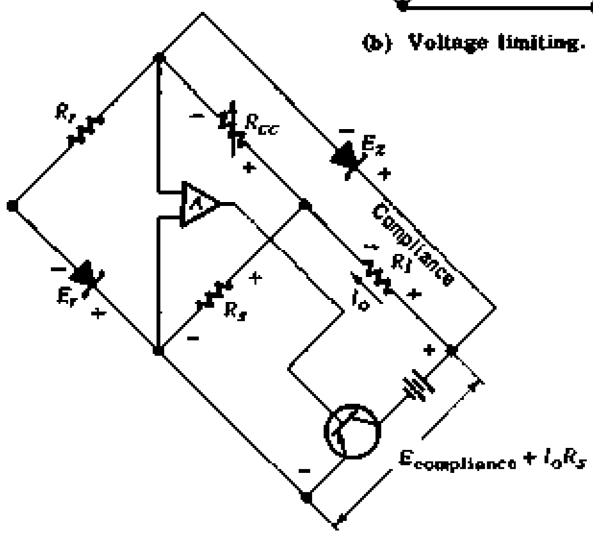
Voltage limiting, which satisfies all of these potential needs, is easily introduced by means of appropriate zener diodes in the feedback leg of the power supply. A zener diode of appropriate voltage breakdown rating, connected across the power supply's voltage control (as in remote programming), will conduct whenever the feedback potential exceeds its zener rating. The zener thus limits the maximum feedback (voltage control) voltage to a specific value. As we have seen in Chap. 2, the magnitude of the power supply output voltage always equals the drop across the voltage control, so that limiting this voltage effectively constrains the output.



(a) Voltage limiting for operationally programmed supplies.



(b) Voltage limiting.



(c) The compliance voltage (across the load, R_l) is limited to E_Z (maximum) minus the voltage across R_{CC}, normally 1 volt or less.

Fig. 7.19

The major advantage of this method is that it permits the use of relatively small, inexpensive zener diodes since the current in the diode will be no more than the bridge current for the supply, typically between 1 and 10 milliamperes. Zener diodes can be obtained in a wide variety of breakdown ratings, and can be connected in series to obtain still more choices.

Figure 7.19a shows the zener limiting circuit applied to an operationally programmed supply. Figure 7.19b shows the same limiting circuit used to place a maximum voltage limit on the output of a conventionally resistance programmed power supply. The application of this method to current regulators is shown in Fig. 7.19c.

When used in circuits having a very small bridge current (1 ma or less), the zener "knee" is not likely to be particularly sharp, and so the zener should be selected for at least 10% excess voltage above the maximum desired output. Of course, for critical loads, positive fail-safe protection is provided by the Kepco Model VIP load protectors.

7.7 LOAD CONTROL

A favorite power supply application is as a load controlling device for other power sources (e.g., batteries). While a power supply cannot itself dissipate power, it can control the voltage across, or current into, a load resistance and thus control dissipation.

(A) Constant Current Battery Discharge: A current regulating power supply placed in series with a battery and a suitable load resistor is capable of maintaining a constant discharge *from* the battery by adjusting its own terminal voltage *automatically* to compensate from a decaying battery voltage (Fig. 7.20).

$$\frac{E_b + E_c}{R_l} = I_{discharge}$$

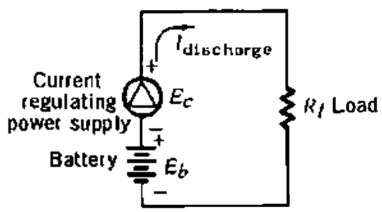


Fig. 7.20 Power supply used to control the discharge current from a battery.

The procedure for calculating the needed power supply compliance, and resistor R_l is best illustrated by a numerical example.

Example: We have a 12V d-c battery to be discharged to exhaustion at a constant 1 ampere rate. When fresh, the battery has a terminal voltage of 12V d-c; when exhausted, zero volts.

Solution: A current regulator power supply in series with the battery can only *add* to the terminal voltage of the load resistor R_l; therefore, choose R_l so as to require slightly *more* than 12V to pass the desired 1 ampere, say 13V. R_l = 13V/1A = 13 ohms. When discharge is initiated, E_b = 12V, E_c = 1V, $R_l = 13\Omega$. When the battery has fully discharged, E_b = 0, E_c = 13V. The required compliance is, therefore, 13 volts.

(B) Reverse Current Loading: In several of the systems applications discussed in Chap. 8, series opposing power supplies are employed to drive an operationally programmed power supply. The driving current (bridge current) obviously has got to pass through one of the series opposing supplies backwards. This is a fairly common occurrence in which a power *supply* must behave as a current *sink*. Recalling the discussion of Chap. 4, the output circuit of a series regulated power supply is unidirectional; it simply will not pass current in the reverse direction (except through the electrolytic capacitors, and other highly undesirable paths) – yet a path must be found for the power supply to absorb a reverse current. Diodes connected to provide a reverse path conduct only when the power supply's terminal voltage reverses (or tries to), and so would not work when the terminal voltage remains correctly polarized. The solution to the problem of reverse current loading is to preload the power supply either resistively or with a current generator, so that the current outward from the supply always exceeds the reverse current magnitude inward.

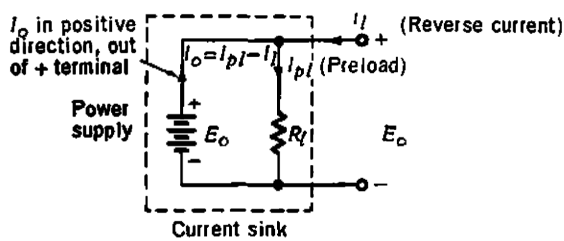


Fig. 7.21 Reverse current load sink.

Consider Fig. 7.21. R_l is adjusted so that

$$I_{pl} \text{ (preload)} = \frac{E_o}{R_l} > I_l$$

So long as the preload current I_{pl} exceeds I_l, the net current at the power supply's terminals is positive out of the plus terminal, the power supply output $I_o = I_{pl} - I_l$.

If the preload resistor R_l is considered a "part of the power supply" as indicated by the dotted lines, the combination can be used as a current sink for I_l. Because I_{pl} is related by Ohms Law to E_o, the preload resistor R_l must be adjusted to draw the proper preload current whenever E_o is varied. This problem can be avoided by adding a current regulator in series with R_l set to draw a specified I_{pl} without regard to the value of E_o (Fig. 7.22).

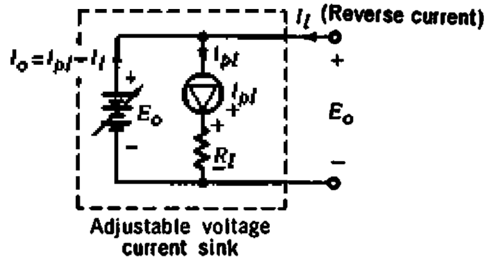


Fig. 7.22 Select R_l so that the voltage drop across R_l, $I_{pl}R_l \geq E_{o\,max}$. This keeps the voltage correctly polarized across the current source I_{pl}.

7.8 ELECTRONIC LOADS

In Chap. 6, in the discussion of loads suitable for the measurement of current regulator performance, it was pointed out that *brush noise* disqualified most adjustable rheostat variable resistance loads. This was because the output capacitor charging or discharging current (in the current regulator being measured), upset the load current enough to make precision measurements difficult. This problem is related to the relatively slow recovery of current regulators and is treated at length in Chap. 9.

Fortunately, there is a way to obtain a jitter-free constant voltage electronic load for testing current regulators. A constant current electronic load for voltage regulators can be constructed with equal ease.

The circuit of Fig. 7.23 is a constant voltage load suitable for loading current regulators in test situations. The power supply E_l

is directly across the "load" terminals and so determines the voltage (e.g., load for a current regulator). The current source I_{pl} serves as a preload and R_l, of course, dissipates the power. If E_l is a well regulated power supply, the load voltage, and therefore, the equivalent load resistance "seen" by the current regulator, is constant, stable and free of jitter. Moreover, if E_l is adjustable, or even programmable, the load terminal voltage can be adjusted throughout the compliance range of the current regulator being tested – precisely and without jitter. In effect, the electronic equivalent resistance seen by the current regulator can be programmed merely by programming E_l. The preload current source I_{pl} is set so that it generates a current through R_l slightly larger than the maximum current of the current regulator being loaded. This insures that the net current flow in E_l is in the positive direction. The resistor R_l is chosen so that the voltage drop across its terminals due to I_{pl} is greater than the maximum value for E_l. This insures that the current source I_{pl} operates into the correct polarity.

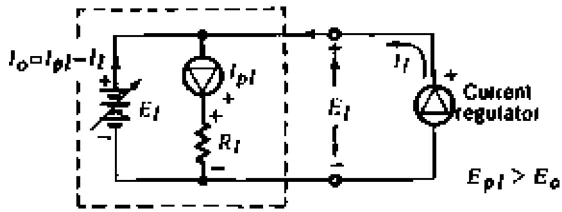


Fig. 7.23 Adjustable voltage electronic load for current regulators.

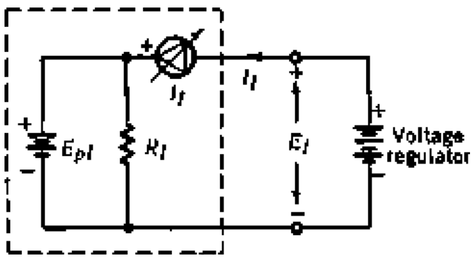


Fig. 7.24 Adjustable current electronic load for voltage regulators.

The dual circuit is shown in Fig. 7.24. Here a current source I_l provides the load for a voltage regulator under test. If I_l is programmable and of sufficient current capacity, then the loading can be programmed throughout the current rating of the voltage regulator. As before, the voltage regulator sees an electronically variable equivalent load resistance. The preload E_l causes a voltage drop across R_l slightly larger than the maximum voltage for the voltage regulator being loaded. This keeps the voltage across I_l properly polarized.

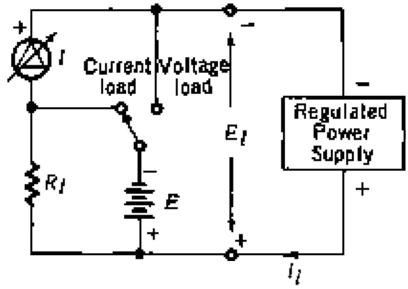


Fig. 7.25 Electronic load, switchable for constant voltage or constant current (shown) loading.

A simple switching arrangement as shown in Fig. 7.25 will allow the same supplies to be used as a constant voltage or constant current load interchangeably.

8 EXTENSION OF THE REGULATING LOOP

Regulated power supplies, as feedback controlled devices, are designed to control the electrical parameters of voltage or current. In many supplies the regulating control loop can also be extended to include a variety of physical elements such as position, speed, temperature, pressure, chemical activity and the like. Assuming the necessary transducers and sensors, servo-like control mechanisms can easily be constructed using d-c regulators as translators, summers, amplifiers, impedance converters – most of the electronic components of a typical servomechanism.

Chaps. 2 and 3 showed how the comparison bridge circuit can be analyzed operationally, yielding specifically the concept of a voltage gain between the input or reference and the output terminals of the device. Moreover, the concept of the constant bridge current and the relative output and feedback circuit impedances were explored at some length.

These properties of the Kepco power supply can be put to use in an enormous variety of useful ways, some of which will be outlined in this chapter, in the hope that it will serve to stimulate the imagination of systems designers.

8.1 BASIC CONTROL CIRCUIT

The simplest controller circuit employs the operational gain of a d-c power supply ($G = R_f/R_1$) to operate on the difference between two voltages, one fixed or controllable (the command, E_c), and the other dependent on the load, E_b. Their difference $E_c - E_b$ is multiplied by the operational gain G, and appears as the output of the power supply, $E_o = G(E_c - E_b)$. If E_o drives a load so that increasing E_o increases E_b somehow, then the necessary condition for negative feedback will have been established for $E_c - E_b$ to diminish, reducing E_o. Assuming the operational gain G can be made sufficiently large, the summing difference $E_c - E_b$ will be a very small quantity, designated ϵ, the error. If ϵ approaches zero, then E_b approaches E_c. In other words, the feedback signal approaches the command signal in magnitude. If E_c is deliberately changed (to exercise command), the resultant difference, ϵ, multiplied by G, provides a large signal to the load which lasts only until the sensor feedback E_b (a function of the load) once again (nearly) matches E_c.

All of this very nearly parallels the discussion in Chap. 2, wherein the process of voltage regulation was explained in terms of a simple comparison circuit. The ideas are exactly the same. In fact, if G is sufficiently large, another *null* or summing junction is created external to the power supply. The internal null is dependent on the open-loop gain of the power supply, while the external feedback null is dependent on the operational gain. Figure 8.1 shows a simple feedback block diagram illustrating all of the vital elements.

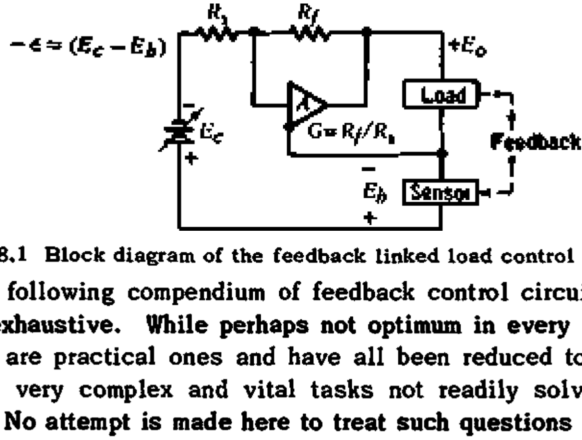


Fig. 8.1 Block diagram of the feedback linked load control circuit.

The following compendium of feedback control circuits is by no means exhaustive. While perhaps not optimum in every respect, the circuits are practical ones and have all been reduced to practice — some in very complex and vital tasks not readily solved by other means. No attempt is made here to treat such questions as transient behavior, stability or the like. The systems all obey the Nyquist and Bodé criteria, and the literature on control systems engineering is fully applicable.*

8.2 SPEED CONTROL

A speed control mechanism is shown in Fig. 8.2. It consists of a d-c motor driven by a compatibly rated d-c power supply, operationally programmed. If the power supply is operated "open loop," that is, controlled by means of a fixed reference to produce a regulated voltage, then the motor's shaft speed is very difficult to

*Control Systems Engineering, Del Toro & Parker. Servomechanism & Regulating Systems Design, Chestnut & Mayer.

control. The speed will be wholly dependent on the output voltage, frictional losses, and most critically, shaft loading. Precision speed control is quite impossible, particularly at the slower speeds. (Fig. 8.3a-b).

If the driving power supply is connected to a feedback circuit consisting of a permanent magnet tachometer and reference supply, so that it is operationally programmed by their difference, $E_c - E_b$, precise speed control can readily be obtained. The output of the tachometer E_b is directly and linearly related to the motor shaft speed through the mechanical coupling which may be geared, or a belt drive. The complete loop equation includes the gear ratio, tachometer speed-to-voltage conversion constant and the operational gain of the driving power supply. Since the power supply's gain is the most easily varied (using R_f), it is usually made variable, and is adjusted for optimum speed control (without hunting or other phase-gain instabilities). The command voltage E_c can be generated by a small variable voltage power supply. It is varied to control the motor speed. Should E_c, for example, be increased, the error ϵ is increased, and so is the motor drive voltage, by an amount $G(\epsilon)$. This in turn speeds the motor, and the tachometer, producing a rise in E_b which just equals the original change in E_c.

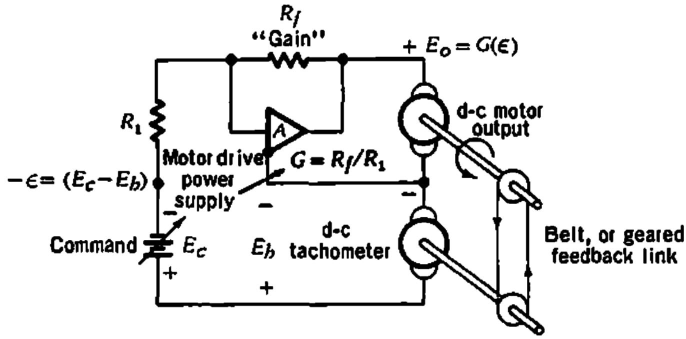


Fig. 8.2 Closed-loop d-c motor speed control.

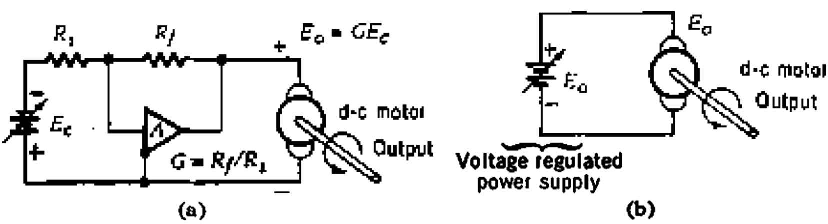


Fig. 8.3 Open-loop motor drive. The circuits (a) and (b) are fully equivalent.

Motor speed control is a common engineering problem. It finds application in process control systems, controllable pumping, fan control, and a host of similar situations. The simple speed control circuit just described is capable of controlling a small d-c motor from a fraction of a revolution per minute to several thousand with a precision not otherwise as readily obtainable.

8.3 ILLUMINATION CONTROL

Another common control problem is the regulation of light intensity. The sources may be filament lamps, high intensity mercury or xenon arc lamps. The uniform illumination of a monochronometer is a typical objective. Since light is a power function, its intensity is regulated by controlling the output *power* from a d-c supply, not either voltage or current. Power regulation is not an easy task; however, there are some circuits which have the hyperbolic E, I characteristic of a constant power curve (Fig. 8.4). The tank circuit of a Flux Oscillator (see Chap. 1) is a good example; it is widely used to drive arc lamps. Another solution to the problem of regulating power is to take the feedback from a power dependent function like heat (temperature) or light (intensity).

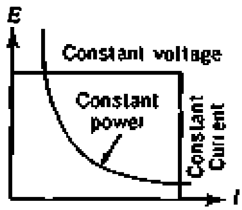


Fig. 8.4

Since light intensity is what we wish to regulate, this latter approach is a direct one capable of yielding very good results. Photo multiplier tubes, photo voltaic cells or photo resistive semiconductors are all suitable feedback sensors. Their diversity also illustrates some interesting power supply applications as signal processing devices. A photo voltaic cell, for example, produces a relatively feeble voltage, requiring some voltage amplification before it can be summed differentially with the command E_C. A small power supply, programmed operationally by such a cell, can amplify its voltage by any convenient amount. Such amplification would be in the feedback loop, contributing to the loop gain (Fig. 8.5a-b).

If a photo resistive cell is used to sense the lamp emission, its output is in the form of a variable terminal resistance as a function

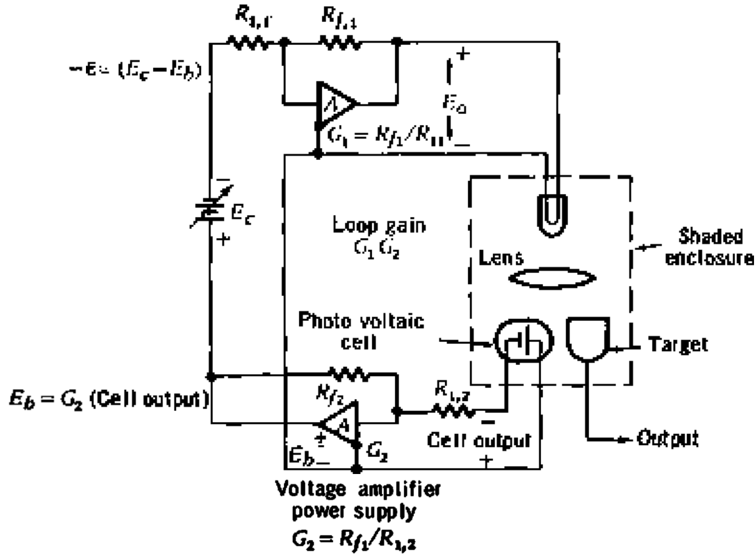


Fig. 8.5a Illumination control using photo voltaic sensor with an operationally programmed power supply as a voltage amplifier.

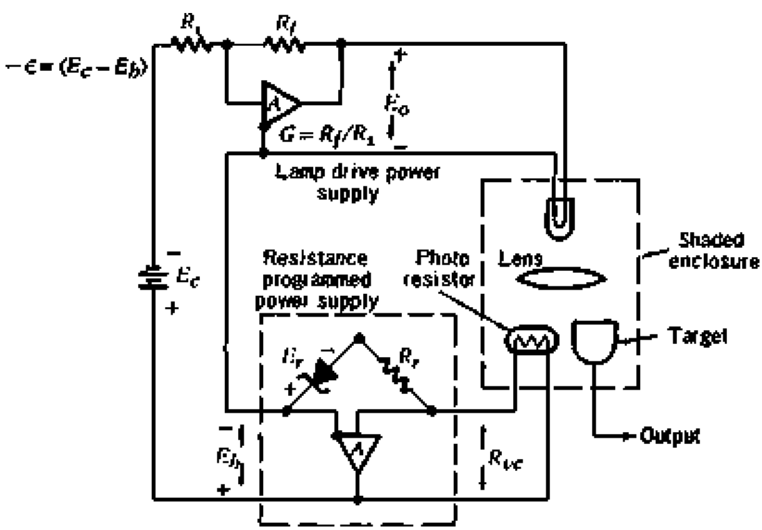


Fig. 8.5b Illumination control using photo resistive sensor with a programmed power supply as a resistance-to-voltage translator.

of the incident illumination. This introduces a translation problem, *resistance to voltage* which is readily solved by using a resistively programmed power supply to convert the cell resistance to a voltage – directly, and linearly,

By whatever means, a voltage E_b can be derived, proportional to light intensity, and suitable for nulling against an external command source E_c. Their difference controls the electrical drive to the lamp through the operational gain G. As shown in Figure 8.5, E_c is adjustable and serves as an intensity control. Depending on the optics and physical arrangements, this sort of control can be used to maintain constant illumination despite varying ambient lighting levels. It can be used to correct for aging lamps (darkened envelope) or for photographic exposure control.

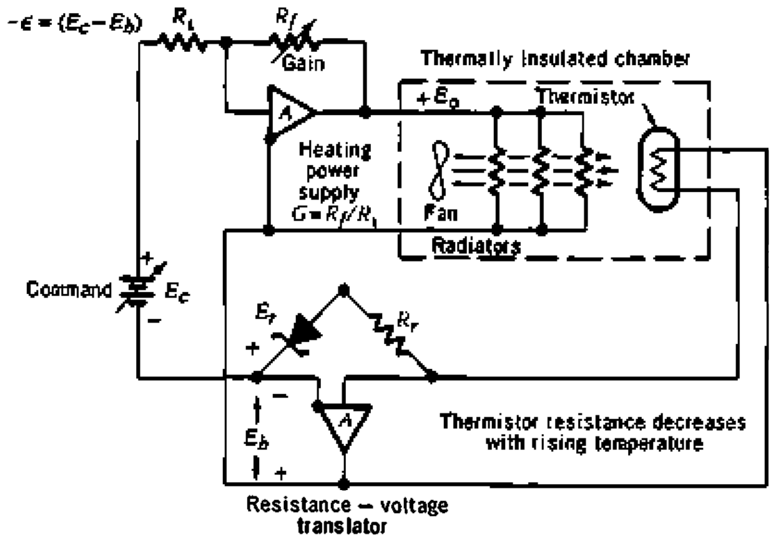


Fig. 8.6 Temperature control using thermistor sensor. Feedback polarities are reversed because thermistor has negative temperature coefficient.

8.4 TEMPERATURE CONTROL

Temperature control, using operationally programmed d-c power supplies for proportional control, is an application of major importance. Power supplies combining high gain amplifier characteristics, with high power output, can be used to generate heat directly as the dissipation in suitable load power resistors. This dissipation, in an insulated enclosure, can be used to control temperature (Fig. 8.6).

Thermistors are a sensitive and yet convenient temperature sensor. With a resistively programmed power supply to convert the thermistor resistance to the sensor voltage, E_b, and another adjustable source for command voltage, E_c, the differential sum $E_c - E_b$ is formed to generate ϵ. The power supply which generates the heating is itself programmed by ϵ through its operational gain $G = R_f/R_1$. By suitably adjusting G, a well insulated chamber using a few circulating fans is quite capable of achieving 0.1°C-0.01°C temperature stability for exceptionally long periods.

The idea of feedback motor speed control can be combined with the temperature control circuit to add a new dimension to the capabilities of a simple insulated environmental chamber. When prolonged temperature tests are conducted on dissipative material (i.e., objects which generate their own heat), some form of cooling or venting to the outside is required. Such venting can be accomplished without upsetting the precise temperature control by adding a second feedback loop as shown in Fig. 8.8. The second loop is made to respond to reverse polarity, ϵ, such as would occur if the chamber were hotter than set by the command control. A d-c motor driven fan, venting either to the outside air or to a specially cooled sink, provides a cooling capability to complement the heating facility. Together they can be made to precisely control the chamber temperature despite varying and substantial amounts of self-dissipation within the chamber. Figure 8.7 shows some Kepco temperature chambers, employing power supplies in the function described.

A schematic block diagram of the two-loop heating-cooling cycle proportional temperature controller is shown in Fig. 8.8. The basic power supply functions are:

1) Thermistor resistance to voltage translator (E_b)

2) Control reference (temperature control, E_c)

3) Heating control amplifier/supply

4) Cooling (air pump) control amplifier/supply

The outputs of supplies 1 and 2 are compared at the summing junction. Their difference, ϵ, drives the two control supplies, 3 and 4. Each control supply responds to only one polarity of the error. A positive error corresponds to a chamber temperature which is higher than commanded, activating the cooling pump. If the chamber is cooler than commanded by E_c, the error activates the heater supply. To

Fig. 8.7a Small temperature control chamber with fresh air venting.

Fig. 8.7b 27 cubic foot temperature control chamber with refrigerated cooling. Recorders at the left are used to measure temperature performance of test samples in the chamber. A miniature air bath chamber with the same basic control mechanism is located under the shelf of the center rack and contains the standard and reference cells.

increase the cooling capabilities of such a system, the air pump might be connected to a *cold sink* refrigerated mechanically.

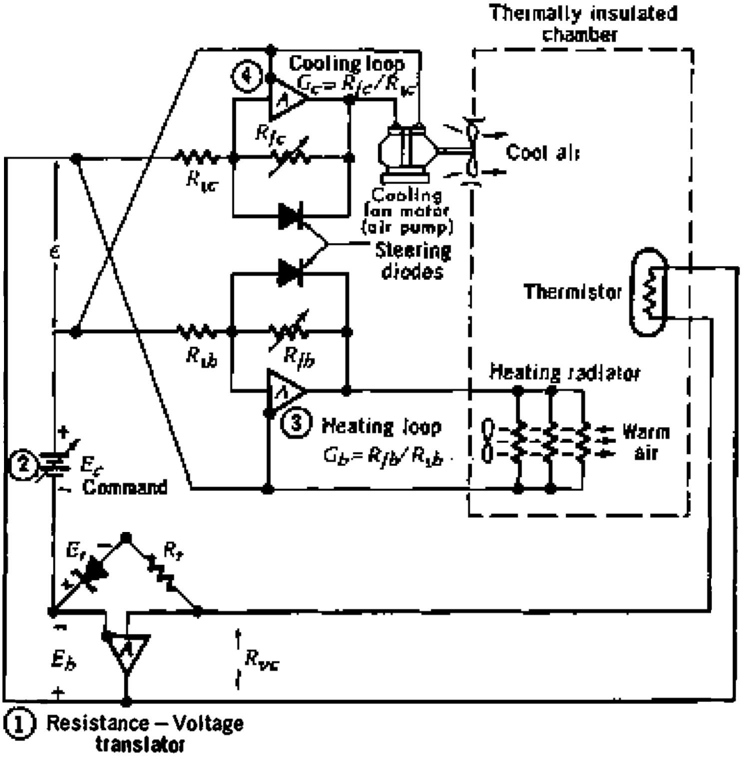


Fig. 8.8 Twin-loop heating and cooling control.

8.5 CHEMICAL POTENTIOSTATS

In the field of chemical analysis, power supplies find application in potentiostat configurations, as shown in Figures 8.9 and 8.10.

In constant potential electrolysis, an operationally connected power supply is used to keep the electro-chemical potential at a special control electrode constant at an externally determined value. By operating the feedback power supply open-loop, this can be accomplished without causing any current flow whatever in the reference electrode. The technique for leakage cancellation, described in Chap. 3, permits the circuit to operate with an input impedance in the vicinity of 10^7 - 10^8 ohms.

In constant current electrolysis, as diagrammed in Fig. 8,10, an impedance transformer power supply configuration will repeat the reference electrode voltage for monitoring by external instrumentation. Again, for exceptionally high input impedance, the leakage cancellation technique is used (Chap. 3).

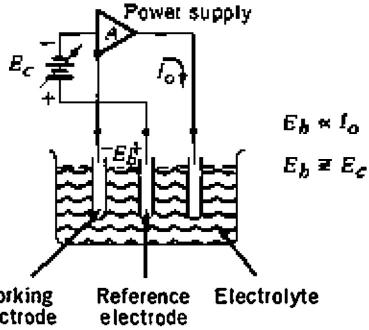


Fig. 8.9 Potentiostat for controlled potential electrolysis.

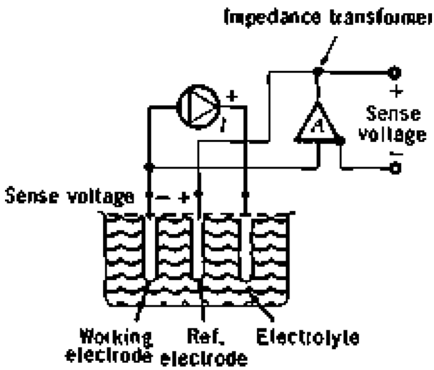
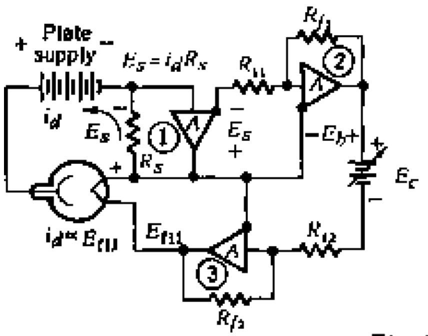


Fig. 8.10 Constant current electrolysis using an "impedance—transformer" power supply to repeat the sense voltage.

8.6 THERMIONIC ELECTRON BEAM CONTROL

Referring to Figs. 3.6a-b, we see how the impedance transformer power supply can be used to detect the current in a high impedance circuit, without loading that circuit, in the examples cited a *diode*. If the diode's current is proportional to filament heat, and if an operationally programmed power supply is used to supply the filament power, then the diode current can be *regulated* by comparing the output of the impedance transformer supply E_b with some adjustable reference E_c, and using their difference to program the filament supply. A circuit for this is shown in Figure 8.11. An application is in electron beam zone refiners in which the schematic *diode* is an evacuated chamber containing the electron gun and target.



(1) Impedance transformer repeats E_S without loading R_S.

(2) Inverting amplifier, changes polarity of E_S. May be set for unity gain if desired, $R_{f1} = R_{i1}$.

(3) Filament power supply, operationally programmed by $E_c - E_b$.

Fig. 8.11

In most of the circuit examples discussed, a differential voltage summation between E_c and E_b has been used to illustrate the ideas involved. That is not the only way an error signal can be generated, and may not always be the most convenient way. For one thing, it requires an external source of E_c.

In all of the circuits, the error ϵ is converted to a control current by the input resistor and virtual ground null junction of the operationally programmed power amplifier. Actually, then, it is this control current (equivalent to I_b in the bridge circuit) that is being nulled. Realizing this, the differential summation can be carried out in the feedback (gain) resistor by establishing a control current I_c and a sensor current I_b. The process is similar to the adder circuit of Fig. 3.9.

As shown in Fig. 8.12, I_b is E_b/R_b, and I_c is E_c/R_c. The ad-

$$I_c = \frac{E_c}{R_c} \qquad I_b = \frac{E_b}{R_b}$$

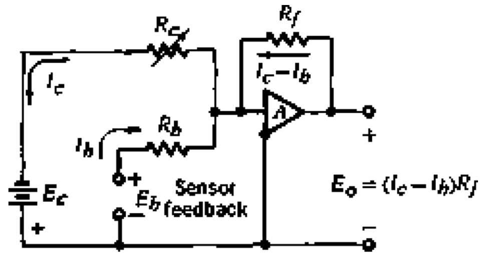


Fig. 8.12 Control current nulling.

vantage obtained by this method is that for a limited range of I_c adjustment, E_c need not be a variable power supply, but could be merely a fixed potential. I_c would be adjusted by changing R_c.

In some situations, the existing reference voltage E_r found within the power supply can be used as the control voltage E_c, if the reference resistor is replaced by an externally adjustable resistor as R_c. Figure 8.12 shows this circuit. Of course, R_c cannot be adjusted to cause I_c to be too much larger than the designed control current I_b (1 - 10 ma usually). Otherwise, the zener diode which controls E_r may be "starved" for current and may lose control of E_r. For applications where the required control span is not excessive, the current nulling technique works very well indeed. The technique is also useful in combination with a voltage comparison – nulling circuit. For example, in the two-loop heating-cooling circuit, the built-in reference voltage E_r coupled through a suitable resistor can be used to inject pre-bias current into each of the control loops, so that each loop becomes activated just *before* the error signal changes polarity. This prevents a temperature dead band by providing for overlapping control near zero error (ϵ).

9 AC CHARACTERISTICS OF DC SUPPLIES

9.1 PROGRAMMING SPEED

In any of the systems application for power supplies, or, for that matter in any programming situation, the question of programming speed must be considered. Programming speed is defined as the time required for the output *voltage* of a programmed power supply to get from one fixed value to another. The time may be measured for a 10% to 90% swing, or for one time constant, or for the whole time, depending on the waveforms, the nature of the response, and the information needed. In general, we will find that the *slope* of the response during the interval between two fixed commands can be closely approximated by the 10% to 90% slope. However, translation to sinusoidal response requires knowledge of the slope to a single time constant. (See Fig. 9.1.)

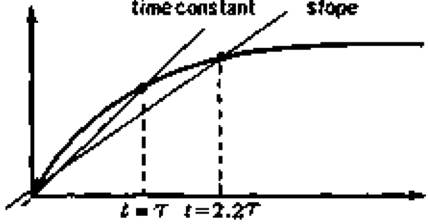


Fig. 9.1 Time constants.

The programming speed provides a considerable insight into another very important power supply characteristic: the *recovery time* for *current regulated* mode of operation. The two are related by the fact that they are ways of describing the rate of change of voltage across the load terminals.

Current regulation may be achieved by means of internal or external sensing, switch selection or automatic crossover. Whatever method, the *idea* of current regulation demands that the terminal voltage of the power supply be capable of automatically assuming any value (within the rated compliance range) as may be needed to regulate the current through a load. In a sense, the power supply (as a current regulator) is asked to automatically program its terminal voltage throughout a range of possible values as demanded by the load. When the power supply does this, it is subject to the same frequency response restrictions that govern the programming speed response to an external program.

The dominating influence on this behavior is, of course, the output capacitor of the power supply. In Chap. 6, we saw how this capacitor introduced an error current (the capacitor charging or discharging current) into the regulation of a current source. This error current is observed during a change in load resistance and follows from the corresponding change in compliance voltage. (See Fig. 9.2.)

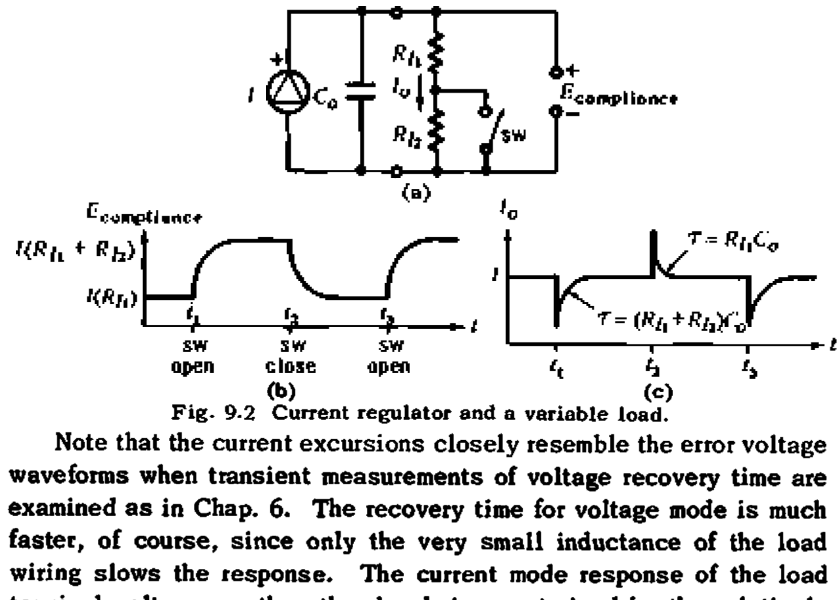


Fig. 9.2 Current regulator and a variable load.

Note that the current excursions closely resemble the error voltage waveforms when transient measurements of voltage recovery time are examined as in Chap. 6. The recovery time for voltage mode is much faster, of course, since only the very small inductance of the load wiring slows the response. The current mode response of the load terminal voltage, on the other hand, is constrained by the relatively large output filter capacitor.

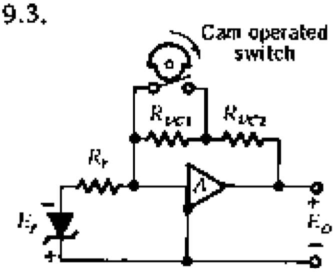
9.2 DYNAMIC TRANSFER FUNCTIONS

Both programming speed and current mode recovery time can be studied by examining the dynamic transfer function or power supply frequency response when considered as a voltage controlled amplifier (see Chap. 3). The transfer function sets the limit on the maximum sinusoidal programming frequency that a power supply will follow, and on the shape of its response to a step program.

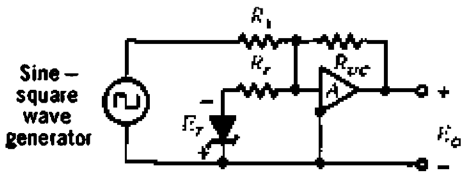
The modern power supply is necessarily a compromise between the most nearly ideal characteristics of its various forms. As we have seen, the ideal characteristics for a good voltage source are almost diametrically opposed to the most suitable characteristics for a current source. Similarly, the strictly voltage-source "power supply" considerations of low ripple, low output impedance, large energy storage and good stability in the presence of arbitrary phase angle loads, conflict with the desirable "amplifier" characteristics of wide bandwidth and fast rise times. Since the devices at hand *are* basically power supplies, and since they are mainly used as voltage regulators, the conflicting requirements are usually resolved in favor of the most nearly ideal voltage-power supply. This chapter will attempt to show the effect of such orientation, how to compute the constraints for a given application, and how to readjust the basic dynamic transfer function to more nearly suit those applications requiring faster response.

Dynamically, a capacitor most nearly resembles an ideal voltage source. It has a low internal impedance, decreasing with increased frequency. Its terminal voltage resists instantaneous change, and it has good energy storage for transient loading. No wonder then that a voltage regulated power supply relies heavily on the presence of a large filter capacitance across its output terminals. If the capacitor is sufficiently large, the cutoff frequency of the output circuit discriminates sharply against ripple frequencies, which is also very desirable in a power supply. The cutoff frequency is the reciprocal of the time constant that the filter capacitance makes with the load resistance and source resistance in series, or parallel, depending on whether the capacitor is being charged or discharged. This RC time constant *(break point)* is the chief factor controlling the programming speed and current mode recovery time. A second somewhat higher frequency break point is contributed by the capacitor across the feedback terminals of the power supply (the voltage control, normally). The two capacitors together look like a large *lag*, approaching 90° if the capacitors are sufficiently large. By providing the power supply with this built-in lag, the designer is able to make the loop gain very large and yet preserve sufficient phase margin to assure stability for almost any arbitrary phase angle load. This means that the power supply is able to drive inductive, resistive and capacitive loads equally, without risk of instability or oscillatory tendencies.

It is convenient to analyze the effects of the output and feedback capacitors in terms of the slope of the maximum ramp slope *dE/dt (max)* for a single time constant τ. A little algebra transforms the ramp into a sinusoidal frequency response by the relation $dE/dt\ (max) = \omega E_p$, where E_p is the peak value of the sinusoid (½ peak-to-peak), and ω is the radian frequency $2\pi f$. These relations are useful because *dE/dt (max)* can be approximated in a number of different ways. Experimentally, it is possible to program the power supply for a step voltage change and observe the exponential response. The power supply can also be programmed with a sinusoidal function, measuring the frequency-amplitude ratio at which distortion begins. Finally, it is possible to compute *dE/dt (max)* from the published schematic value for output capacitance using the energy relationship in a capacitor: $dE/dt\ (max) = I/C$; that is, the slope of the maximum response ramp can be approximated by dividing the power supply's current rating (in amperes) by the size of the filter capacitor (in farads). Circuits for generating step and sinusoidal programs are shown in Fig. 9.3.

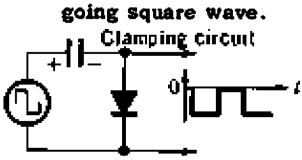


(a) One method of generating a step program.



$$\frac{1}{2}\ \frac{E_{pp}\ (\text{sq wave})}{R_1} < \frac{E_r}{R_r}$$

(b) A better method employs a square wave oscillator. If the oscillator used has no zero reference, a clamping circuit can be used to obtain a negative going square wave.



(c) Square wave clamped negative. Note: some tilt may be introduced by this method, particularly if the coupling resistor (R_1) is small.

Fig. 9.3

The break point associated with the output filter capacitor usually occurs at the lowest frequency, so the RC time constant that this capacitor makes with the load and source resistors dominates the power supply's a-c response. The load resistance is easy to define, but difficult to use in computations since it is a variable. Moreover, the internal or source impedance is also a variable; in fact, it is bi-valued. In many modern current limiting or automatic crossover power supplies, the normal low source impedance as a voltage regulator electronically "switches" to a high source impedance (current regulator) whenever the current setting of its limiting control is exceeded. The internal resistance is also unidirectional (like a diode), in that the power supply's regulated circuit will not pass current in the reverse direction.

Consider the equivalent circuit for the programmed power supply and its load. R_i is the internal *source* resistance of the supply, R_l its load, and C_o the output capacitance. If this supply is programmed with a positive step, the output voltage E_o will experience an increasing exponential, $e_o(t) = E_{oss}(1 - e^{-t/\tau_r})$, where E_{oss} is the steady-state (final) value of E_o, $E_{oss} = [R_l/(R_i + R_l)]E_i$, and τ_r is the rising time constant (R_i in parallel with R_c)C_o.

If the peak charging current in C_o does not exceed the current limit setting of the power supply, the R_i is much, much less than R_l, so that $E_{oss} \approx E_{i_2}$ and $\tau_r \approx R_iC_o$. In about four time constants, the output rises to 98% of the steady-state value E_{oss}. If the power supply's rise time is defined as the time required for the output to rise from 10 to 90 percent of the final steady-state voltage after excitation by a unit step, the rise time, found by substituting these limits into the equation, equals 2.2 τ_r. (See Fig. 9.4.)

Having reached the steady-state value, the power supply can be programmed downward with a negative going unit step. Since the charge on the output capacitor cannot flow back through the source impedance, R_i can be considered infinite for the discharging step, leaving a simple RC circuit consisting of the load resistance, R_l, and the capacitor, C_o. The output decays according to the equation $e_o(t) = E_o e^{-t/\tau_f}$, where τ_f is R_lC_o. The fall time can also be approximated by 2.2 τ_f for a 90% to 10% decay or as 4 τ_f for a 98% decay.

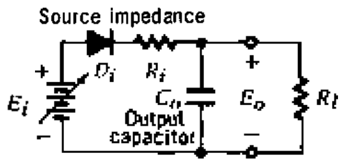
It is significant to note that the rise time constant τ_r and the falling time constant τ_f are not the same, and depending, respectively, on the source resistance for τ_r and the load resistance for τ_f.

Should the sum of the load current and capacitor charging current exceed the current limit setting during a rising step, the source would assume the characteristics of a current generator. (See Fig. 9.5.)

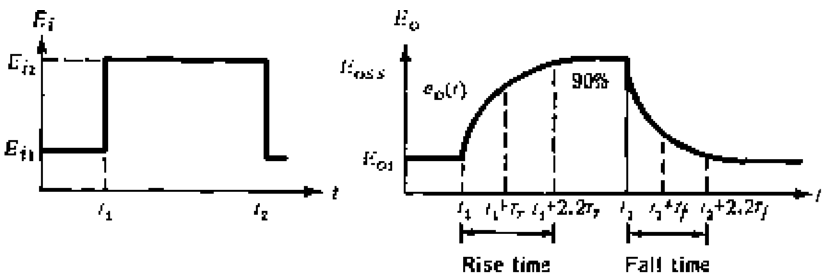
During this interval the charging of the capacitor has the form $e_o(t) = IR_l(1 - e^{-t/R_lC})$ with the time constant R_lC.

If the load current alone is not sufficiently large to keep the power supply operating in its current regulated (or limited) mode, this situation may only prevail for a portion of the charging cycle, lasting until the charging current into the capacitor diminishes sufficiently as to reduce the total current demand below the current limit point, whereupon the circuit reverts to its low source resistance configuration of Fig. 9.4.

The rising time constant is two-valued, having the value $\tau_r = R_lC$ while charging toward IR_l, and then switching to $\tau_r = R_iC$ while charging to E_{oss}. Since the constant current portion of the charging cycle occurs near the beginning of the rise, while E_o is still small, the current through R_l is not yet significantly large and may be



(a) Equivalent circuit for series regulated power supply.



(b)

Fig. 9.4 Response to a square wave program.

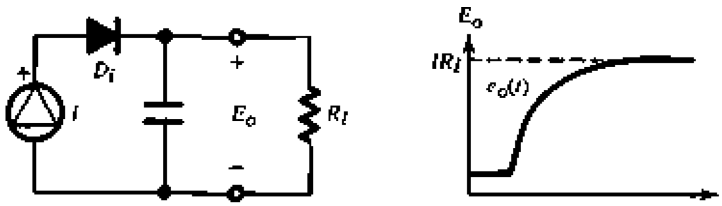


Fig. 9.5 Charging waveform as a current generator.

neglected for purpose of simplification. This leaves an unloaded current source to charge the capacitor producing a linear ramp for that portion of the cycle as shown in Fig. 9.6.

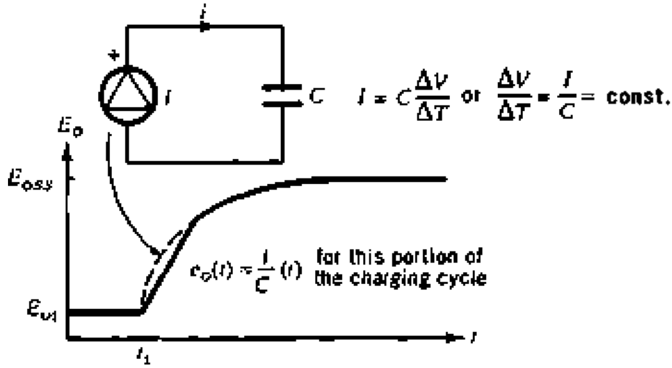


Fig. 9.6 Current limiting during a portion of a rising program.

The final (exponential) portion of the rise is completed with $\tau_r = R_l C$.

It is apparent that the setting of the current limiting point relative to the load controls the rise time, while the value of the load resistance primarily affects the fall time.

9.3 SINUSOIDAL RESPONSE

The slope of the rise or fall time to one time constant closely corresponds to the slope of a single integrator $dE/dt\ (max) = I_o/C_o = -\omega E_p$.

The actual time required for a given voltage swing depends on the nature and size of the load. Capacitive loads will respond more slowly than purely resistive or inductive loads, making it difficult to specify the programming time for an arbitrary load. Nevertheless, it can be approximated. The size of the typical output filter capacitor is related to the magnitude of the power supply's output current rating by such factors as ripple, current, and the need to establish sufficient energy storage for an adequate voltage mode transient load response. For a surprising variety of power supply ratings, the ratios of I_o/C_o tend to fall within a reasonably restricted range of values, sufficiently concentrated as to permit the choice of one value as representative for some approximate computations. The ratio I_o/C_o ranges from about 100 volts per second for low voltage supplies to as much as 1000-2000 volts per second for high voltage equipment where, presumably, excessive capacity is an expensive luxury. The mean value, and the one chosen for computation, lies near 250 volts per second. This is the slope that will be measured to one time constant, when the load resistance is adjusted for nearly equal rise and fall time. That load setting is usually found at about half-rated load.

The slope constraint can be converted to sinusoidal frequency response by letting $dE/dt\ (max) = -\omega E_p = 250\text{V/sec}$. Frequency and peak excursions are seen to be dependent quantities; specifying one sets the maximum value for the other. For example, for a 2 volt peak-to-peak output excursion, $E_p = 1$ volt, and the maximum frequency,

$$f = \frac{250}{2\pi} \times \frac{1}{E_p}$$

yields

$$f = 40 \text{ cps.}$$

Peak frequency and excursion can be traded off linearly, so that halving the peak voltage doubles the maximum frequency or vice versa.

The maximum frequency derived by this relationship is determined by the steepest slope of the sinusoid when this slope just exceeds the maximum programming speed. When this happens, the power supply no longer follows the rise and fall of the sinusoid but instead takes on its own exponential form for a portion of the cycle. The sinusoidal rising voltage is delayed until it meets itself going downward, whereupon, with the source diode open, the voltage decays exponentially, giving rise to slope or diagonal distortion. (See Fig. 9.7.)

When the ratio dE/dt is used to describe the recovery time occasioned by a load change in the current regulated mode, account must be taken of the nonconstant load resistance R_l. When the downward step was *programmed* into the power supply, the decay time constant was $\tau_f = C_o R_l$. This is also true of a downward step oc-

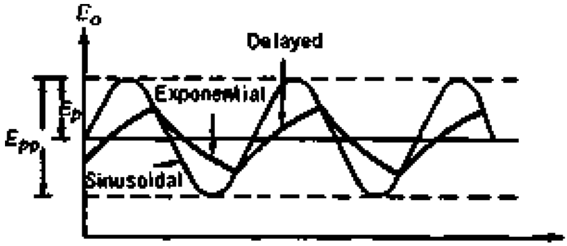


Fig. 9.7 Sinusoidal input $E_p \omega > dE/dt\ (max)$, output does not follow the driving waveform.

casioned by a sudden reduction in R_l for current source regulation, except that now R_l will have just changed to a smaller value of resistance and so τ_f will be faster. (Fig. 9.8.)

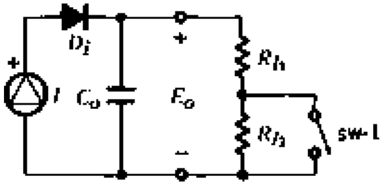


Fig. 9.8 For a downward step program in I,

$$R_l = R_{l_1} + R_{l_2},\ \tau_f = C_o(R_{l_1} + R_{l_2}).$$

For constant I, if switch (Sw-1) is closed,

$$R_l(0^+) = R_{l_1},\ \tau_f = R_{l_1}C_o.\quad R_{l_1}C_o < (R_{l_1} + R_{l_2})C_o.$$

Whatever the speed of decay, the discharge current from C_o must pass through the load R_l and may represent a hazard to delicate loads which should be accounted for by the careful designer.

9.4 MODIFYING THE DYNAMIC CHARACTERISTICS

The design value of C_o is based upon the needs of a good voltage regulator – low source resistance, ripple, and large energy storage. As has been pointed out, these characteristics are not always the most desirable for current regulator operation or high speed programming. For particular applications it may sometimes be desirable to judiciously reduce C_o. Such reduction will usually have its most significant drawback in the loss of some phase margin stability, and an increased sensitivity to the reactive nature of the load. It is generally necessary to readjust one or more of the internal circuit lag networks to re-establish a stable operating point.

As C_o is reduced, the role of the feedback capacitor which bridges the voltage control becomes increasingly significant. Adjustment of C_o alone can produce as much as a 10:1 increase in the programming speed. If the feedback capacitor is simultaneously reduced or eliminated entirely, the programming speed can be extended even further. Ripple will increase somewhat due to a-c pick-up across the relatively high resistance of the voltage control, but good shielding practices can usually keep the increase in ripple within acceptable limits. Such changes in the basic lag structure of a power supply will almost certainly require readjustment of its stabilizing circuits. This accomplished, however, the power supply will take on the characteristics of a pretty good *wideband* amplifier with full response up to several hundred cycles, and even several kc for some models.

In some model lines, Kepco manufactures high-speed (HS) regulators in which this modification has been performed. In the PAX series, for example, a choice of plug-in regulator cards allow for "normal" or "high-speed" regulation. HS operation permits up to 50,000 volts per second programming speed (slewing rate). The HS regulators usually provide some means for adjusting the lag networks because of their greater sensitivity to reactive loads. In addition, limitations are usually placed on the maximum capacitive loading in order to assure stability even with lag adjustment.

Another class of power supplies/amplifiers is the Kepco OPS (operational power supply). OPS have been optimized as amplifiers with up to 100,000 volts per second programming speed, very high open-loop gain, and excellent stability. They are particularly useful in some of the very low-power repeater, transformation and other signal processing tasks described in Chaps. 2 and 7.

Other groups are – or can be – converted to high-speed applications, except that in the case of some preregulator supplies (particularly those employing intermittent – sampled data – SCR preregulators), substantial derating may be required to retain the designed dissipation limits.

When a power supply is used in a HS version, it acquires the characteristics of a wideband amplifier with somewhat higher ripple noise – spread over a wider spectrum. Lacking the output capacitor, the output impedance will be significantly higher, particularly in the upper frequency range, and transient loads will feel the lack of energy storage. The momentary voltage excursions will be larger, but the recovery time is not changed. Such a power supply/amplifier is a much more nearly ideal current regulator (when set up to regulate current). By virtue of its ability to slew voltage very rapidly, an HS current regulator has much smaller transient excursions and much faster recovery. A 10 volt change at 50,000 volts per second, for example, is accomplished in 0.2 milliseconds, which is an improvement of better than 100:1 over the speed of a standard (fully filtered) regulator.

9.5 DUTY CYCLE LOADING

When a voltage regulator is subjected to a transient loading, the size of the output filter capacitor determines the amount of voltage *droop* that will occur during periods when the output current exceeds the power supply's rating. In this situation, the larger the filter capacitor, the better. During the interval while the output circuit is overloaded, the output capacitor must supply a part of the load current equal to the required current, less the maximum output of the supply.

Fig. 9.9 illustrates the distribution of currents, assuming that the overload period is short relative to the discharge time constant of the capacitor. The output voltage droop can be considered linear, falling ΔV in Δt.

$$\frac{\Delta V}{\Delta t} = \frac{I_c}{C_o}$$

where I_c is the net capacitor current $I_l - I_o$, and C_o is, of course, the output capacitor.

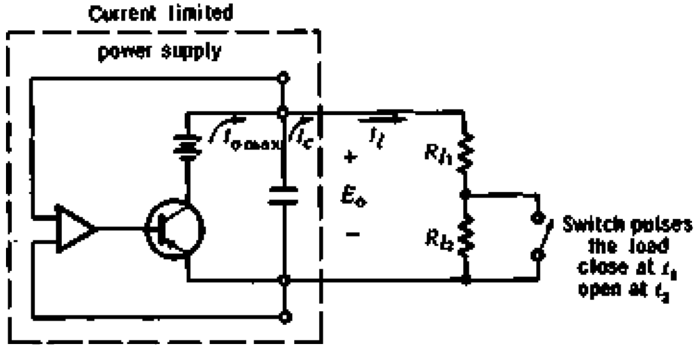


Fig. 9.9 Distribution of currents during overload. For $I_l > I_{o\,max}$, the capacitor makes up the difference with I_c. $I_l = I_{o\,max} + I_c$.

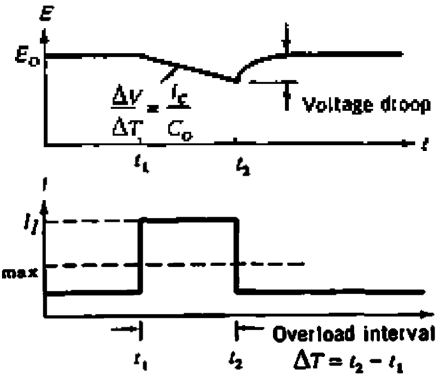


Fig. 9.10 Output waveform during overload.

A numeral illustration will make the calculation clear.

Example: We have a 10 ampere power supply in which $C_o = 30{,}000$ uf. Find the voltage droop ΔV during a 100 microsecond, 50 ampere overload.

Solution:

$$\frac{I_c}{C_o} = \frac{50A - 10A}{.03 \text{ farads}} = \frac{40}{.03} = 1333 \text{ V/S}$$

$$\frac{\Delta V}{\Delta t} = 1333$$

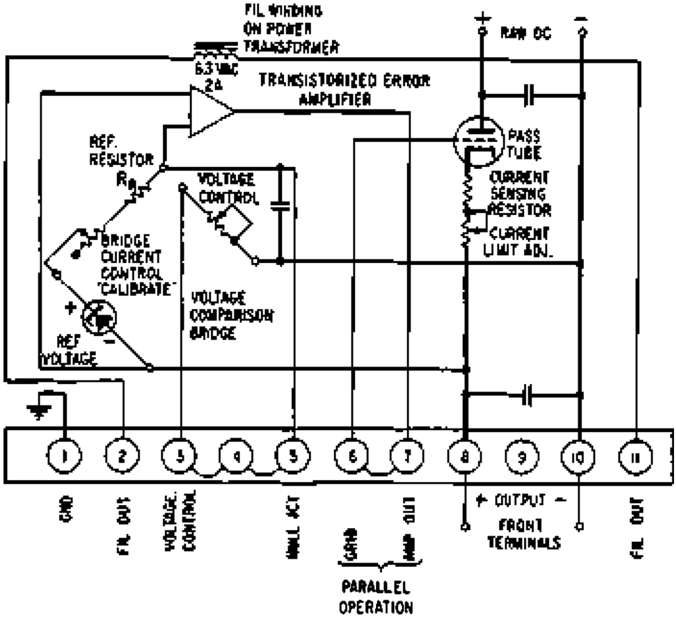
Therefore $\Delta V = 133(100) \times 16^{-6}$ ($\Delta t = 100\ \mu\text{sec} = 100 \times 10^{-6}$ sec)

$$\Delta V = 0.13 \text{ volts}$$

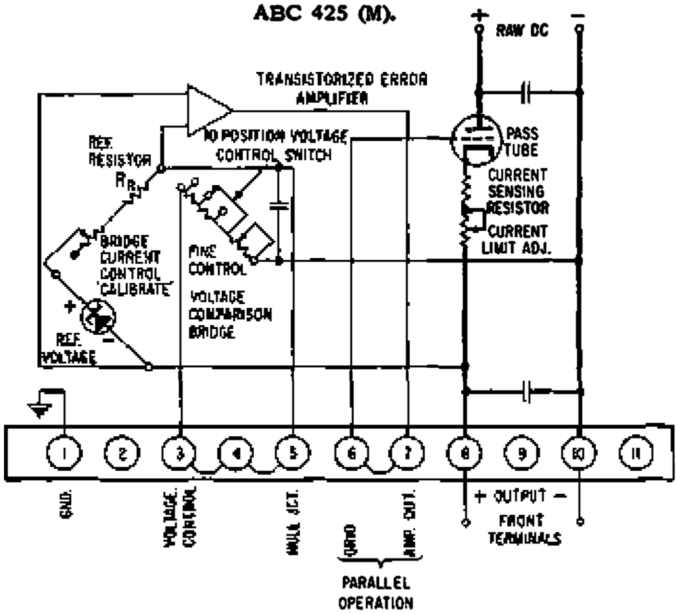
The output voltage droop can be limited to any given value by simply paralleling additional capacitance across the output terminals of the power supply. How much capacitance to add is determined by reversing the calculation.

$$C = \frac{I_c \Delta t}{\Delta V}$$

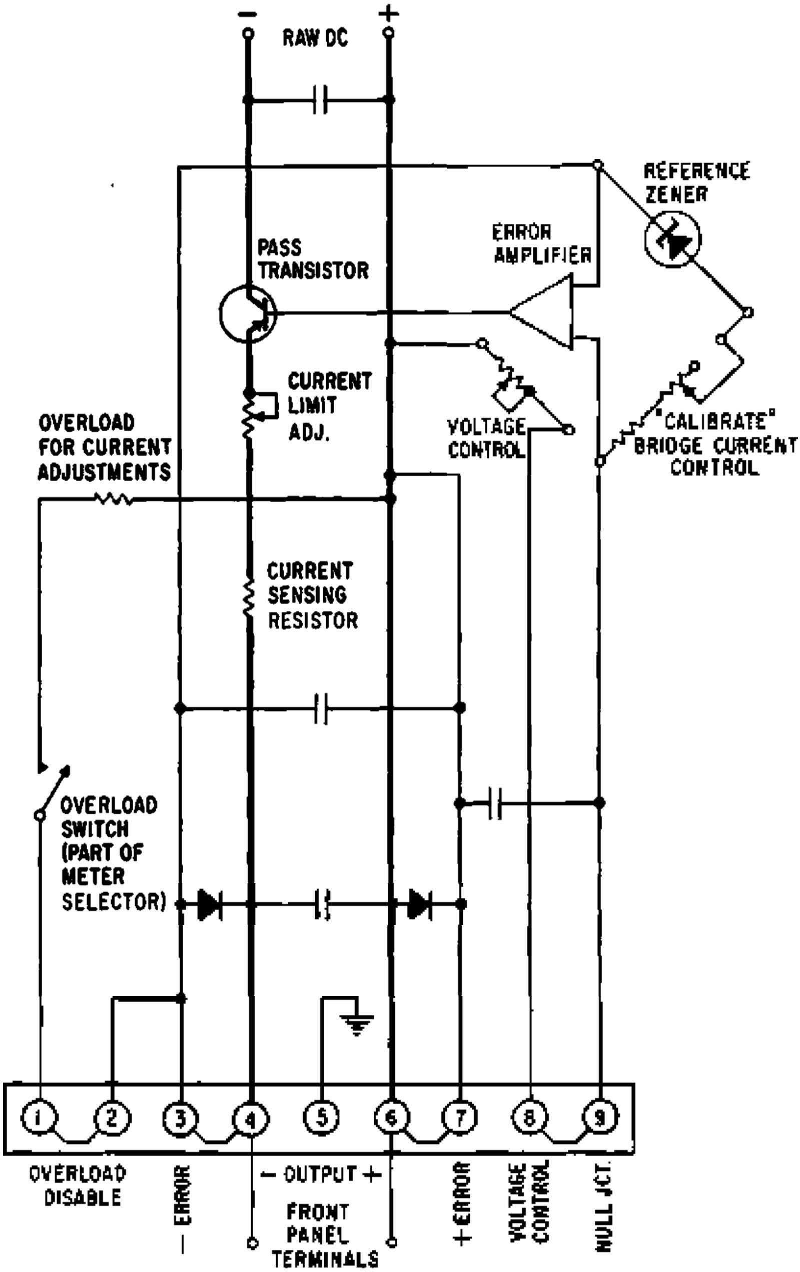
APPENDIX



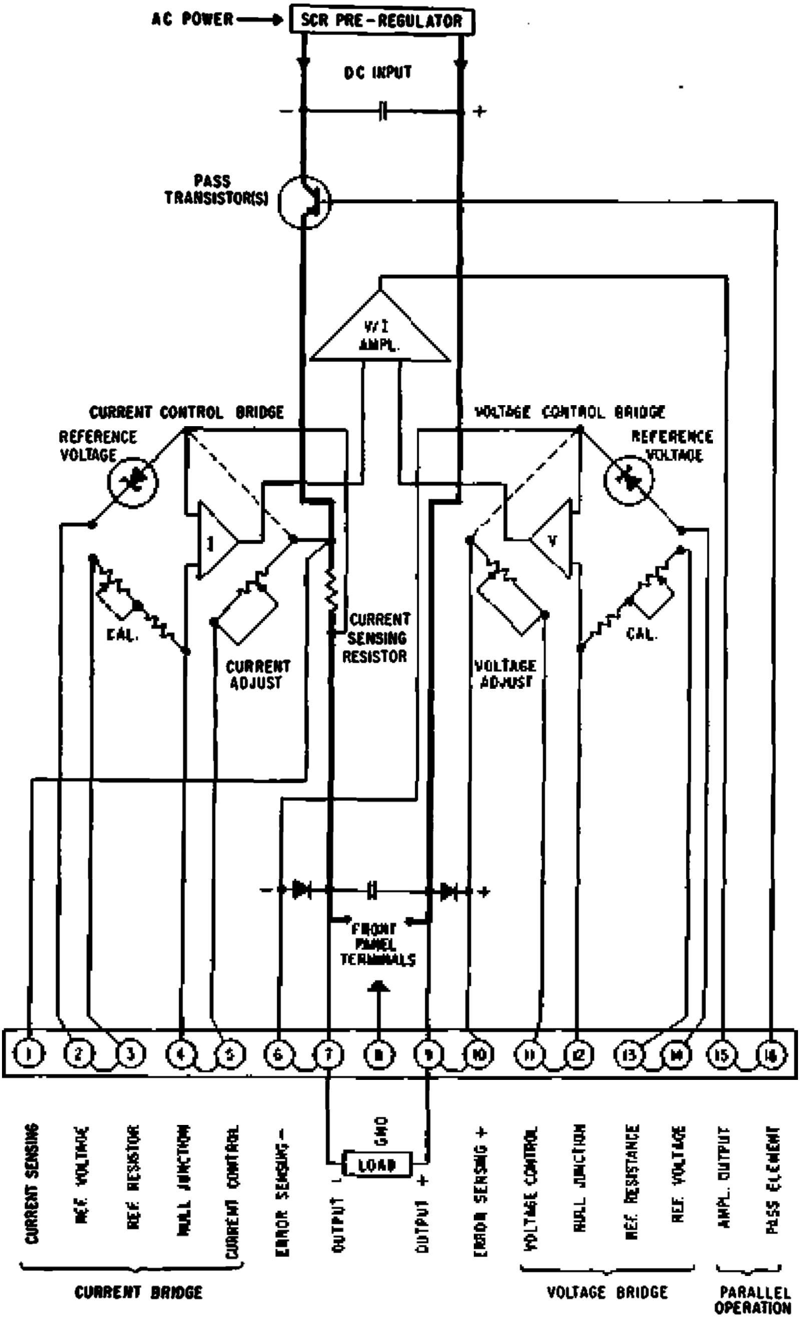
A.1 Simplified diagram and connection guide for Models ABC 200 (M) and ABC 425 (M).



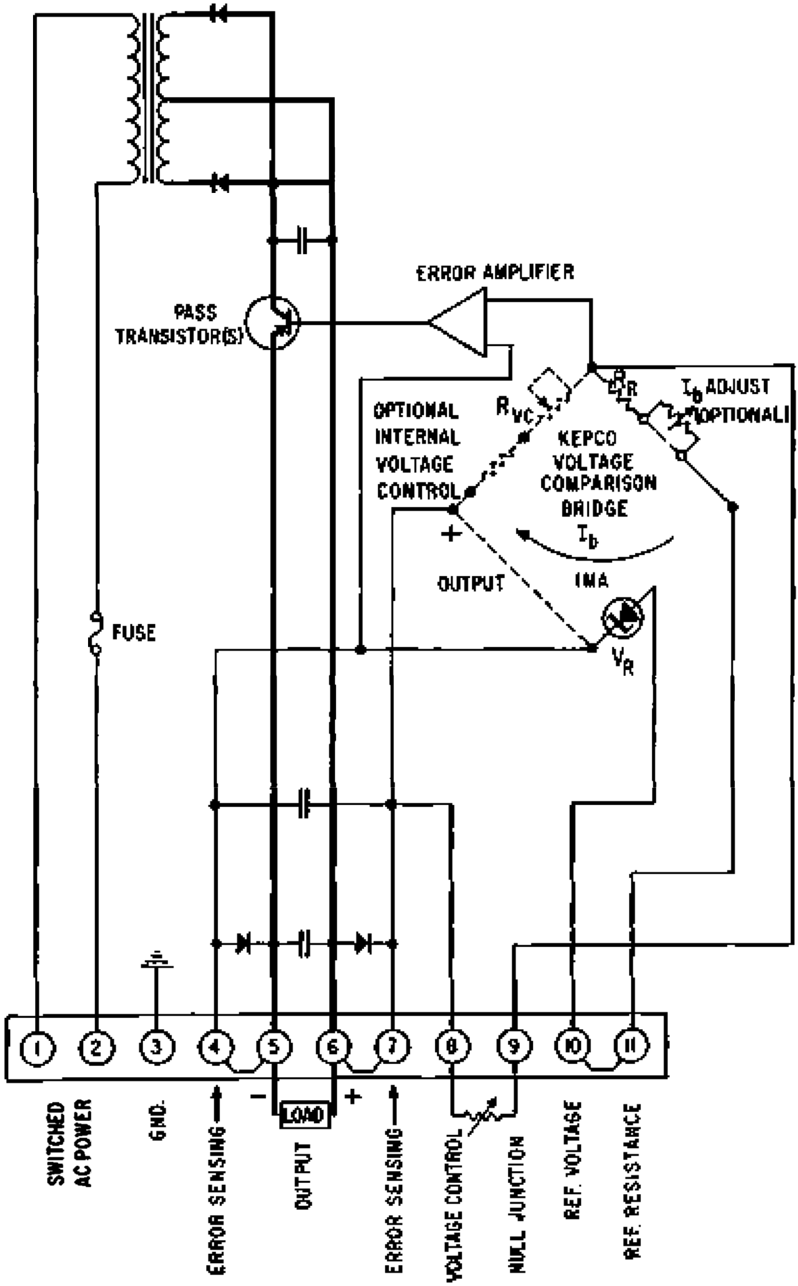
A.2 Simplified diagram and connection guide for Models ABC 1000, ABC 1500 and ABC 2500.



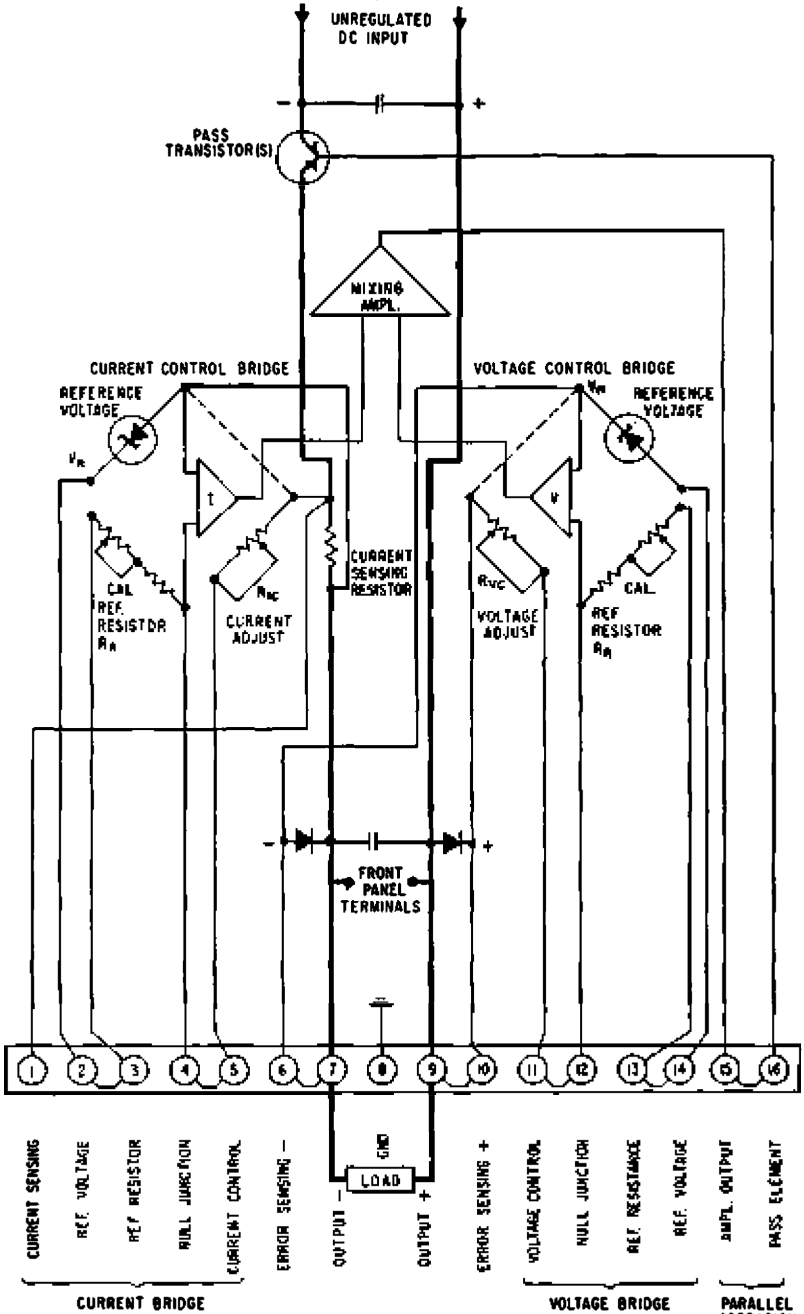
A.3 Simplified diagram and connection guide for ABC all-transistor models.



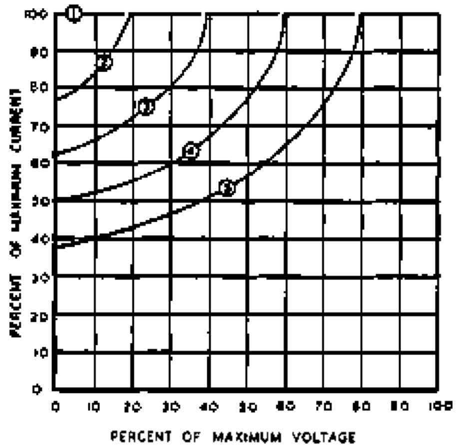
A.4 Simplified diagram for Model KS auto-crossover power supply with rear terminal connection detail.



A.5 Simplified diagram for Model PAX power supply with rear terminal connection detail.



A.6 Simplified diagram of Model CK auto-crossover power supply with rear connection detail.



This graph is provided to help compute the over-ranging capabilities of HB Power Supplies when externally programmed outside of the normal range of the voltage band switch. The graph is particularly useful for determining the voltage compliance range of the power supplies in their constant current mode of operation. To use the graph, plot the desired current as a horizontal line; the intercept that this line makes with each numbered locus defines the minimum output voltage permissible for that particular range switch position. For example: with Model HB 4AM operating in current regulator mode at 60% of rated current, the voltage compliance limits are 30% to 80% when the voltage range switch is in position (4). This corresponds to operation at 240 ma in band position 4 with 160 volts of compliance between 100 V DC and 260 V DC.

Fig. A.7 Load current derating graph for HB models.

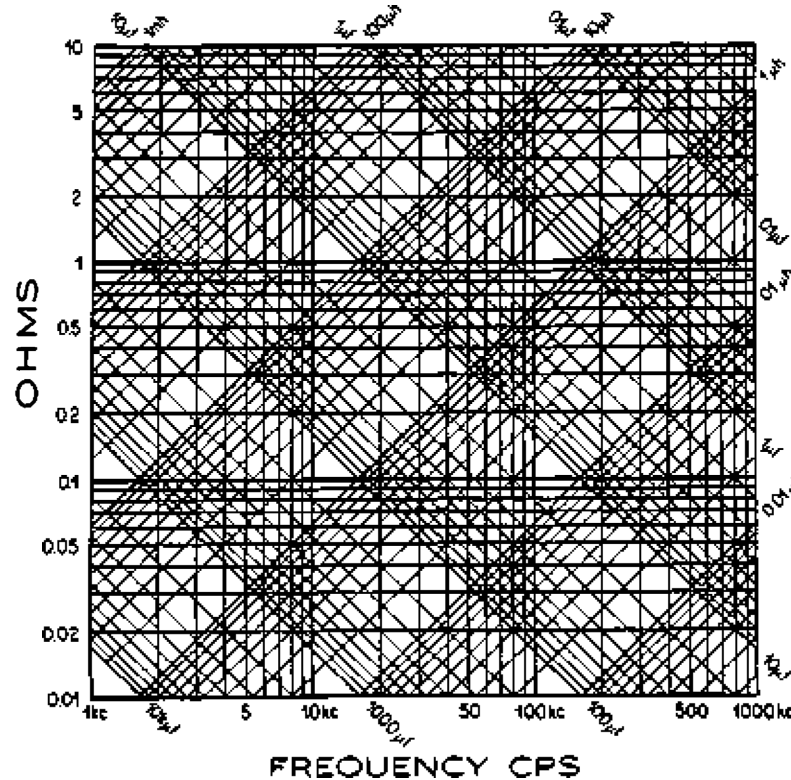


Fig. A.8 Reactance chart.

THIS NOMOGRAPH CAN BE USED TO FIND

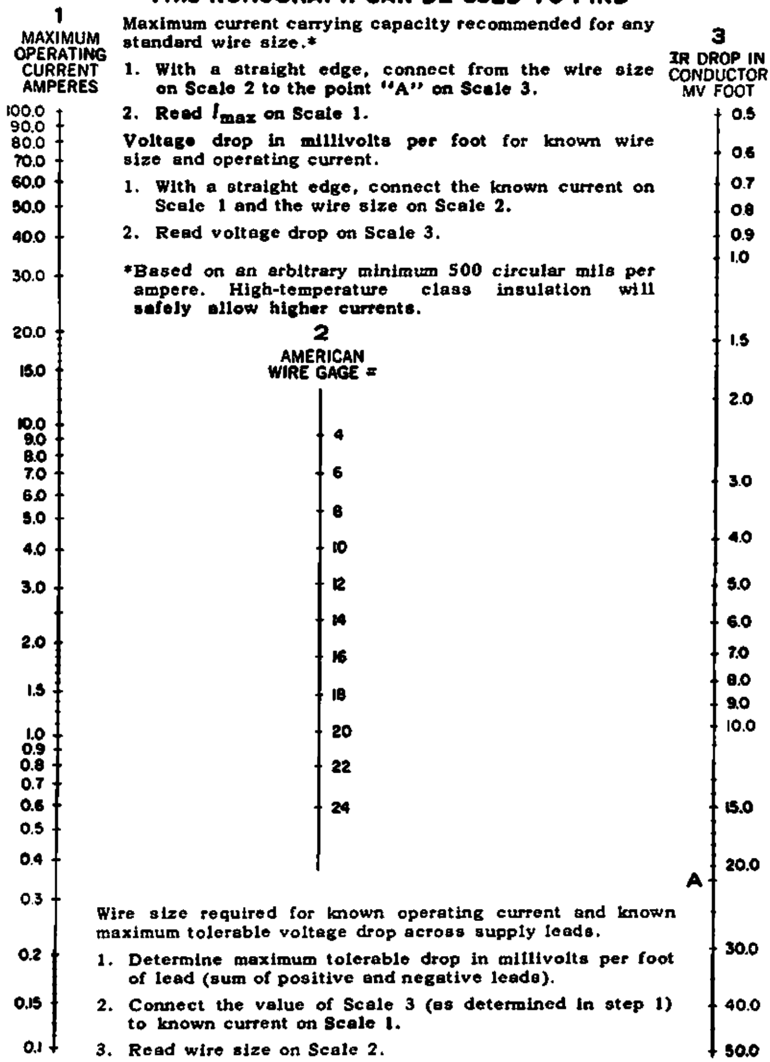
Maximum current carrying capacity recommended for any standard wire size.*

1. With a straight edge, connect from the wire size on Scale 2 to the point "A" on Scale 3.
2. Read I_{max} on Scale 1.

Voltage drop in millivolts per foot for known wire size and operating current.

1. With a straight edge, connect the known current on Scale 1 and the wire size on Scale 2.
2. Read voltage drop on Scale 3.

*Based on an arbitrary minimum 500 circular mils per ampere. High-temperature class insulation will safely allow higher currents.



Wire size required for known operating current and known maximum tolerable voltage drop across supply leads.

1. Determine maximum tolerable drop in millivolts per foot of lead (sum of positive and negative leads).
2. Connect the value of Scale 3 (as determined in step 1) to known current on Scale 1.
3. Read wire size on Scale 2.

NOTE: A voltage regulated Power Supply controls the voltage across its output terminals. Hence the wire conductors used to connect the load must be considered as part of the load. At high load currents the voltage drop across the supply leads may appreciably degrade regulation at the load. Kepco models equipped with the *remote error sensing* feature can automatically compensate for voltage drops of up to 500 mv across each load supply lead.

A.9 Nomograph of voltage drop across load supply leads (as a function of wire size and load current).

GLOSSARY OF POWER SUPPLY TERMS

This glossary is published to assist the engineer in the proper application and understanding of Kepco Regulated Power Supplies. The glossary gives the precise definition for every term that is used in describing and specifying Kepco Power Supplies.

ACCURACY

Used as a specification for the output voltage power supplies, accuracy refers to the absolute voltage tolerance with respect to the stated nominal output.

AMBIENT OPERATING TEMPERATURE (Range):

The range of environmental temperatures in which a power supply can be safely operated. For units with forced air cooling, the temperature is measured at the air intake.

BIPOLAR:

Having two poles, polarities or directions. Applied to amplifiers or power supplies, it means that the output may vary in either polarity from zero; as a symmetrical program, it need not contain a d-c component. *(See* Unipolar)

BRIDGE CURRENT:

The circulating control current in the comparison bridge. Bridge current equals the reference voltage divided by the reference resistor. Typical values are 1 ma and 10 ma, corresponding to control ratios of 1000 ohms/volt and 100 ohms/volt, respectively.

CALIBRATION, PROGRAMMING:

Calibration with reference to power supply programming describes the adjustment of the control bridge current to calibrate the programming ratio in ohms per volt. Many programmable supplies incorporate a "calibrate" control as part of the reference resistor which performs this adjustment.

CLOSED-LOOP GAIN (Operational Gain):

The gain, measured with feedback, is the ratio of the voltage appearing across the output terminal pair to the causative voltage required at the input resistor. The closed-loop (operational) gain is denoted by the symbol G in diagrams and equations. If the open-loop gain A is sufficiently large, the closed-loop gain can be satisfactorily approximated by the ratio of the feedback resistor R_f to the input resistor R_1. *(See* Open-loop, Loop gain)

COMMAND REFERENCE:

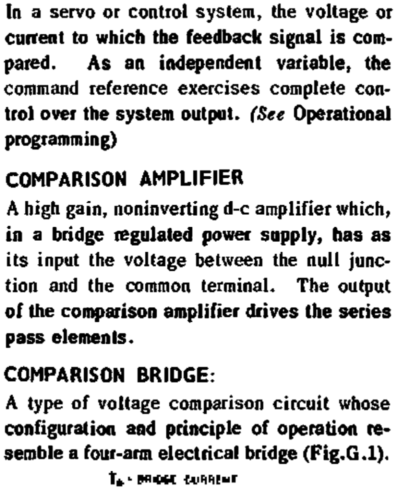
In a servo or control system, the voltage or current to which the feedback signal is compared. As an independent variable, the command reference exercises complete control over the system output. *(See* Operational programming)

COMPARISON AMPLIFIER

A high gain, noninverting d-c amplifier which, in a bridge regulated power supply, has as its input the voltage between the null junction and the common terminal. The output of the comparison amplifier drives the series pass elements.

COMPARISON BRIDGE:

A type of voltage comparison circuit whose configuration and principle of operation resemble a four-arm electrical bridge (Fig.G.1).

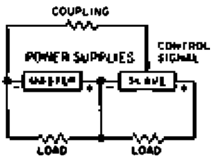


G.1 Kepco comparison bridge connected as a voltage regulator.

The elements are so arranged that, assuming a balance exists in the circuit, a virtual zero error signal is derived. Any tendency for the output voltage to change in relation to the reference voltage creates a corresponding error signal, which, by means of negative feedback, is used to correct the output in the direction toward restoring bridge balance (*See* Error signal).

COMPLEMENTARY TRACKING:

A system of interconnection of two regulated supplies in which one (the master) is operated to control the other (the slave). The slave supply voltage is made equal (or proportional) to the master supply voltage and of opposite polarity with respect to a common point (*See* Fig. G.2).



G.2 Complementary tracking

COMPLIANCE EXTENSION:

A form of master/slave interconnection of two or more current regulated power supplies to increase their compliance voltage range through series connection.

COMPLIANCE VOLTAGE:

The output voltage of a d-c power supply operating in constant current mode. The compliance range is the range of voltages needed to sustain a given value of constant current throughout a range of load resistances.

CONSTANT CURRENT POWER SUPPLY

(Current Regulator):

A power supply capable of maintaining a preset current through a variable load resistance. This is achieved by automatically varying the load voltage in order to maintain the ratio E_{load}/R_{load} constant.

CONSTANT VOLTAGE POWER SUPPLY

(Voltage Regulator):

A power supply that is capable of maintaining a preset voltage across a variable load resistance. This is achieved by automatically varying the output current in order to maintain the product of load current times load resistance constant.

CONTROL RATIO:

The required change in control resistance to produce a one volt change in the output voltage. The control ratio is expressed in ohms per volt and is reciprocal of the bridge current.

COOLING:

In power supplies, the cooling of regulator elements refers to the method used for removing heat generated in the regulating process. Methods include radiation, convection, and conduction or combinations thereof.

COOLING, CONVECTION:

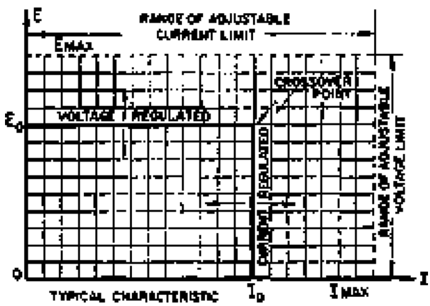
A method of heat transfer which uses the natural upward motion of air warmed by the heat dissipators.

COOLING, LATERAL FORCED AIR:

An efficient method of heat transfer by means of side-to-side circulation which employs blower movement of air through or across the heat dissipators.

CROSSOVER (AUTOMATIC) VOLTAGE/ CURRENT:

The characteristic of a power supply that automatically changes the method of regulation from constant voltage to constant current (or vice versa) as dictated by varying load conditions (Fig. G.3). The constant

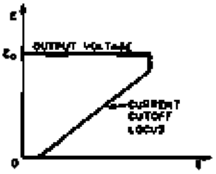


G.3 Automatic voltage/current crossover.

voltage and constant current levels can be independently adjusted within the specified voltage and current limits of the power supply. The intersection of constant voltage and constant current lines is called the crossover point *E*, *I* and may be located anywhere within the volt-ampere range of the power supply.

CURRENT CUTOFF:

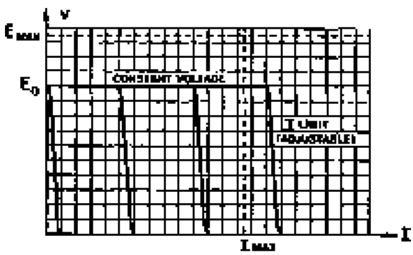
An overload protective mechanism designed into certain regulated power supplies to automatically *reduce* the load current as the load resistance is *reduced*. This "negative resistance" characteristic reduces overload dissipation to negligible proportions and protects sensitive loads. See Fig. G.4 for the *E, I* characteristic of a power supply equipped with a current cutoff overload protector.



G.4 Output characteristic of a power supply equipped with a *current cut-off* over-load protector.

CURRENT LIMITING (Automatic):

An overload protection mechanism which limits the maximum output current to a preset value, and automatically restores the output when the overload is removed. *(See* Short circuit protection and Fig. G.5).



G.5 Plot of typical current limiting curves.

CURRENT SENSING RESISTOR:

A resistor placed in series with the load to develop a voltage proportional to load current. A current regulated d-c power supply regulates the current in the load by regulating the voltage across the sensing resistor.

"DELTA," MINIMUM:

A qualifier, often appended to a percentage specification to describe that specification when the parameter in question is a variable, and particularly when that variable may approach zero. The qualifier is often known as the "minimum delta *V*," or minimum delta *I*," as the case may be.

DRIFT:

See Stability.

ERROR SIGNAL:

The error signal is the difference between the output voltage and a fixed reference voltage compared in ratio by the two resistors at the null junction of the comparison bridge; i.e., $\epsilon = E_o - E_r\,[R_{vc}/R_r]$ *(see* Fig. G.1). The error signal is amplified to drive the pass elements and correct the output.

FILTERS:

Filters are RC or LC networks arranged as low pass devices to attenuate the varying component that remains when a-c voltage is rectified. In power supplies without subsequent active series regulators, the filters determine the amount of ripple that will remain in the d-c output. In supplies with active feedback series regulators, the regulator mainly controls the ripple with output filtering serving chiefly for phase-gain control as a *lag* element.

FLUX-O-TRAN:®

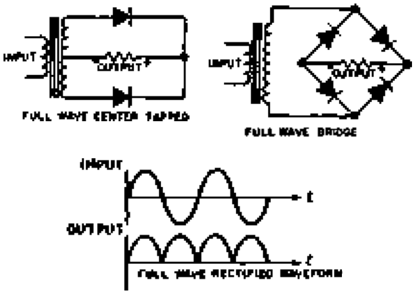
A registered trademark of Kepco, Inc., applied to ferro-resonant voltage regulating transformers of a special design, which are used in many proprietary designs. The Flux-O-Tran, with its resonating capacitor provides a squarewave output (for high rectifier and and filter efficiency) whose magnitude is largely independent of the primary voltage amplitude.

FREQUENCY RESPONSE:

The measure of an amplifier or power supply's ability to respond to a sinusoidal program. The frequency response measures the maximum frequency for full-output voltage excursion. This frequency is a function of the slewing rate and unity gain bandwidth.

FULL-WAVE RECTIFICATION:

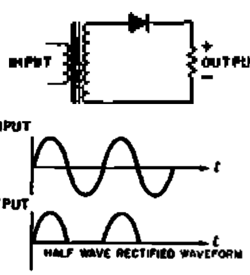
In the rectifying process, full-wave rectification inverts the negative half-cycle of the input sinusoid so that the output contains two half-sine pulses for each input cycle. A pair of rectifiers arranged as shown with a centertapped transformer, or a bridge arrangement of four rectifiers and no centertap are both methods of obtaining full-wave rectification. *(See* Fig. G.6).



G.6 Full wave rectification.

HALF-WAVE RECTIFICATION:

In the rectifying process, half-wave rectification passes only one-half of each incoming sinusoid, and does not pass the opposite half-cycle. The output contains a single half-sine pulse for each input cycle. A single rectifier, arranged as in Fig. G.7, provides half-wave rectification. Because of its poorer efficiency and larger a-c component, half-wave rectification is usually employed in noncritical low current circumstances.



G.7 Half wave rectification.

HIGH SPEED REGULATOR:

A power supply regulator circuit which, by the elimination of its output capacitor, has been made capable of much higher slewing rates than are normally possible. High speed (HS) regulators are used where rapid step programming is needed, or as current regulators for which they are ideally suited. (*See* Slewing rate.)

HYBRID:

A combination of disparate elements to form a common circuit. In power supplies, the combination of vacuum tubes and transistors in the regulating circuitry.

INVERTING AMPLIFIER:

An amplifier whose output polarity is reversed as compared to its input. Such an amplifier obtains its negative feedback by a connection from output to input, and with high gain is widely used as an operational amplifier. An operational d-c power supply can also be described as a high gain inverting amplifier.

ISOLATION VOLTAGE:

A rating for a power supply which specifies the amount of *external* voltage that can be connected between any output terminal and ground (the chassis). This rating is important when power supplies are connected in series.

LAG NETWORKS:

Resistance-reactive components, arranged to control phase-gain rolloff versus frequency. Used to assure the dynamic stability of a power supply's comparison amplifier. The main effect of a lag network is a reduction of gain at relatively low frequencies so that the slope of the remaining rolloff can be relatively more gentle.

LEAD NETWORKS:

Resistive-reactive components arranged to control phase-gain rolloff versus frequency. Used to assure the dynamic stability of a power supply's comparison amplifier. The main effect of a lead network is to introduce a phase lead at the higher frequencies, near the unity gain frequency.

LINEARITY, PROGRAMMING:

The linearity of a programming function refers to the correspondence between incremental changes in the input signal (resistance, voltage or current) and the consequent incremental changes in power supply output. Direct programming functions are inherently linear for the Kepco bridge regulator, and are accurate to within a percentage equal to the supply's regulating ability.

LINE REGULATION:

The maximum steady-state amount that the output voltage or current will change as the result of a specified change in line voltage (usually for a step change between 105-125 or 210-250 volts, unless otherwise specified). Regulation is given either as a percentage of the output voltage or current, and/or as an absolute change, ΔE or ΔI.

LOOP (LEAKAGE) CURRENT:

A d-c current flowing in the feedback loop (voltage control) independent of the control current generated by the reference zener diode source and reference resistor. The loop (leakage) current remains when the reference current is made zero. It may be compensated for, or nulled in special applications to achieve a very high impedance (zero current) at the feedback (voltage control) terminals.

LOOP GAIN:

A measure of the feedback in a closed-loop system, being equal to the ratio of the open-loop to the closed-loop gains, in db, A – G. The magnitude of the loop gain determines the error attenuation and, therefore, the performance of an amplifier used as a voltage regulator. *(See* Open-loop and Closed-loop gain.)

MASTER/SLAVE OPERATION:

A system of interconnection of two regulated power supplies in which one (the master) operates to control the other (the slave). Specialized forms of the master/slave configuration are used in a) *Complementary tracking* (plus and minus tracking around a common point), b) *Parallel operation* to obtain increased current output for voltage regulation, c) *Compliance extension* to obtain increased voltage output for current regulation.

MODULAR:

The term *modular* is used to describe a type of power supply designed to be built into other equipment, either chassis or rack mount. It is usually distinguished from laboratory bench equipment by a large choice of mounting configurations and by a lack of meters and controls.

MTBF Mean time between (or before) failure:

A measure of reliability giving either the time before first failure or, for repairable equipment, the average time between repairs. MTBF may be approximated or predicted by summing the reciprocal failure rates of individual components in an assembly.

NULL JUNCTION:

That point on the Kepco bridge at which the reference resistor, the voltage control resistance and one side of the comparison amplifier coincide. The null junction is maintained at almost zero potential and is a *virtual ground.* *(See* Summing point.)

OFFSET VOLTAGE:

A d-c potential remaining across the *comparison amplifier's* input terminals (from the null junction to the common terminal) when the output voltage is zero. The polarity of the offset voltage is such as to allow the output to pass through zero and the polarity to be reversed. It is often deliberately introduced into the design of power supplies to reach and even pass zero output volts.

OPEN-LOOP GAIN:

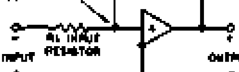
The gain, measured without feedback, is the ratio of the voltage appearing across the output terminal pair to the causative voltage required at the (input) null junction. The open-loop gain is denoted by the symbol A in diagrams and equations. *(See* Closed loop and Loop gain.)

OPERATIONAL POWER SUPPLY:

A power supply whose control amplifier has been optimized for signal processing applications rather than the supply of steady-state power to a load. A self-contained combination of operational amplifier, power amplifier and power supplies for higher level operation applications.

OPERATIONAL PROGRAMMING:

The process of controlling the output voltage of a regulated power supply by means of signals (which may be voltage, current, resistance or conductance) which are *operated*

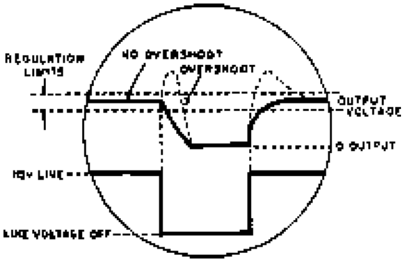


G.8 Operational programming.

on by the power supply in a predetermined fashion. Operations may include algebraic manipulations, multiplication, summing, integration, scaling and differentiation. *(See* Fig. G.8.)

OUTPUT IMPEDANCE:
The effective dynamic output impedance of a power supply is derived from the ratio of the measured peak-to-peak change in output voltage to a measured peak-to-peak change in alternating load current. Output impedance is usually specified throughout the frequency range d-c to 100 kc.

OVERSHOOT:
A transient rise beyond regulated output limits, occurring when the a-c power input is turned on or off, and for line or load step changes. *(See* Figs. G.9, G.11a-b).



G.9 Scope view of turn-off/turn-on effects on a power supply.

OVER-TEMPERATURE PROTECTION:
A thermal relay circuit which turns off the power automatically should an over-temperature condition occur.

PARALLEL OPERATION:
Voltage regulators, connected together so that their individual output currents are added and flow in a common load. Several methods for parallel connection are used: spoiler resistors, master/slave connection, parallel programming and parallel padding. Current regulators can be paralleled without special precaution.

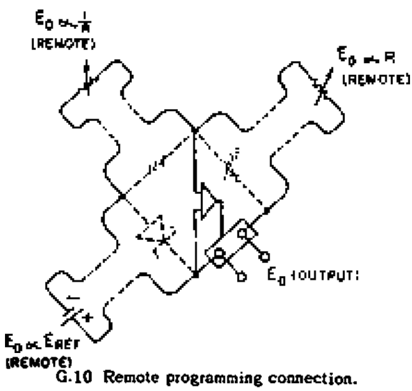
PARALLEL PADDING:
A method of parallel operation for two or more power supplies in which their current limiting or automatic crossover output characteristic is employed so that each supply regulates a portion of the total current, each parallel supply adding to the total and "padding" the output only when the load current demand exceeds the capability – or limit setting – of the first supply.

PARALLEL PROGRAMMING:
A method of parallel operation for two or more power supplies in which their feedback terminals (voltage control terminals) are also paralleled. These terminals are often connected to a separate programming source.

PASS ELEMENT:
A controlled variable resistance device, either a vacuum tube or power transistor, in series with the source of d-c power. The pass element is driven by the amplified error signal to increase its resistance when the output needs to be lowered or to decrease its resistance when the output must be raised. *(See* Series regulator.)

POWER SUPPLY (a-c to d-c):
Generally, a device consisting of a transformer, rectifier and filter for converting a-c to a prescribed d-c voltage or current.

PROGRAMMING:
The control of any power supply functions, such as output voltage or current, by means of an external or remotely located variable control element. Control elements may be variable resistances, conductances, or variable voltage or current sources. *(See* Fig. G.10.)



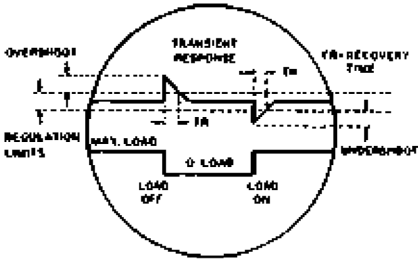
G.10 Remote programming connection.

PROGRAMMING SPEED:
Programming speed describes the time required to change the output voltage of a power supply from one value to another. The output voltage must change across the load and because the supply's filter capacitor forms an RC network with the load and in-

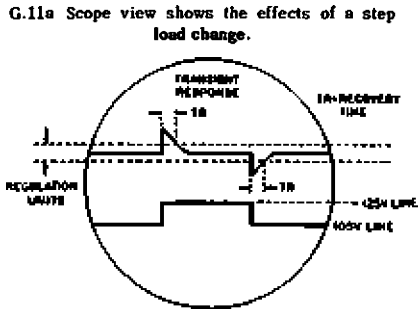
ternal source resistance, programming speed can only be described as a function of load. Programming speed is the same as the "recovery time" specification for *current regulated* operation; it is not related to the recovery time specification for voltage regulated operation.

RECOVERY TIME (Current Regulation):
Specifies the time needed for the output current to return to a value within the regulation specification after a step load or line change. For load change, current will recover at a rate governed by the rate-of-change of the compliance voltage across the load. This is governed by the RC time constant of the output filter capacitance, internal source resistance and load resistance. *(See* Programming speed.)

RECOVERY TIME (Voltage Regulation):
Specifies the time needed for the output voltage to return to a value within the regulation specification after a step load or line change. Recovery time, rather than response time, is the more meaningful and therefore preferred way of specifying power supply performance, since it relates to the regulation specification. *(See* Figs. G.11a-b).



G.11a Scope view shows the effects of a step load change.



G.11b Scope view shows the effects of a step line change.

REGULATED POWER SUPPLY:
A power supply which maintains a constant output voltage (or current) for changes in the line voltage, output load, ambient temperature or time.

REGULATION:
The maximum amount that the output will change as a result of the specified change in line voltage, output load, temperature or time. Line regulation, load regulation, stability, and temperature coefficient are defined and usually specified separately.

REMOTE ERROR SENSING:
A means by which the regulator circuit senses the voltage directly at the load. This connection is used to compensate for voltage drops in the connecting wires.

RESPONSE TIME (Time Constant):
Specifies the time required for a voltage or current excursion to be reduced to 37% of its peak value after a step load or line change. This is not the preferred way of specifying voltage regulator performance. *(See* Recovery time).

RESOLUTION:
The minimum voltage (or current) increment within which the power supply's output can be set using the panel controls. For continuous controls, the minimum increment is taken to be the voltage (or current) change caused by one degree of shaft rotation.

RIPPLE:
Stated either in peak-to-peak or in *rms* value, ripple specifies the maximum a-c component that appears in a d-c output. Unless specified separately, ripple includes unclassified noise.

SERIES OPERATION:
The output of two or more power supplies connected together to obtain a total output voltage equal to the sum of their individual voltages. Load current is equal and common through each supply. The extent of series connection is limited by the maximum specified potential rating between any output terminal and ground. *(See* Isolation voltage.) For series connection of current regulators, master/slave (compliance extension) or automatic crossover is used.

SERIES REGULATOR:

A device placed in series with a source of power that is capable of controlling the voltage or current output by automatically varying its series resistance. *(See* Pass element.)

SHORT CIRCUIT PROTECTION (Automatic):

Any automatic current limiting system which enables a power supply to continue operating at a limited current, and without damage, into any output overload including short circuits. The output voltage must be restored to normal when the overload is removed, as distinguished from a fuse or circuit-breaker system which opens at overload and must be closed to restore power. *(See* Current limiting, Fig. G.5.)

SHUNT REGULATOR:

A device placed across the output, which controls the current through a series dropping resistance to maintain a constant voltage or current output.

SLAVED TRACKING:

A system of interconnection of two or more regulated supplies in which one (the master) operates to control the others (the slaves). The output voltage of the slave units may be equal or proportional to the output voltage of the master unit. (The slave output voltages track the master output voltage in a constant ratio.) *(See* Complementary tracking, Master/slave.)

SLEWING RATE:

A measure of the programming speed or current-regulator response timing. The slewing rate measures the maximum rate-of-change of voltage across the output terminals of a power supply. Slewing rate is normally expressed in volts per second ($\Delta E/\Delta T$) and can be converted to a sinusoidal frequency-amplitude product by the equation $f(E_{pp})$ = slewing rate/π, where E_{pp} is the peak-to-peak sinusoidal volts. Slewing rate = $\pi f(E_{pp})$. *(See* High speed regulator.)

SPOILER RESISTORS:

Resistors used to *spoil* the load regulation of regulated power supplies to permit parallel operation when not otherwise provided for.

STABILITY, LONG TERM (LTS):

The change in output voltage or current as a function of time, at constant line voltage, load and ambient temperature (sometimes referred to as drift).

STEP LINE VOLTAGE CHANGE:

An instantaneous change in line voltage (e.g., 105-125V a-c); for measuring line regulation and recovery time.

STEP LOAD CHANGE:

An instantaneous change in load current (e.g., zero to full load; for measuring the load regulation and recovery time.

SUMMING POINT:

(See Null junction). The null junction is called a summing point because, as the input to a high gain d-c amplifier, operational summing can be performed at this point. As a virtual ground, the summing point decouples all inputs so that they add linearly in the output, without other interaction. *(See* Operational programming.)

TEMPERATURE COEFFICIENT (TC):

The percent change in the output voltage or current as a result of a 1°C change in the ambient operating temperature (% per °C).

TEMPERATURE, OPERATING:

The range of environmental temperatures in which a power supply can be safely operated (typically, -20°C to +50°C). [*See* Ambient operating temperature (Range).]

TEMPERATURE, STORAGE:

The range of environmental temperatures in which a power supply can be safely stored (typically, -40°C to +85°C).

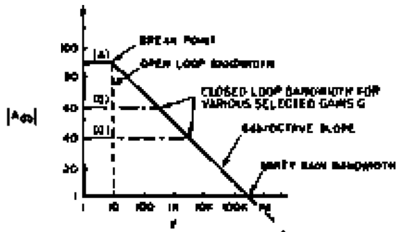
UNIPOLAR:

Having but one pole, polarity or direction. Applied to amplifiers or power supplies, it means that the output can vary in only one polarity from zero and, therefore, must always contain a d-c component. *(See* Bipolar.)

UNITY GAIN BANDWIDTH:

A measure of the gain-frequency product of an amplifier. Unity gain bandwidth is the frequency at which the *open-loop gain be-*

comes unity, based on a 6 db per octave crossing. [*See* Fig. G.12, Typical Gain-Frequency (Bodé) Plot.]



G.12 Gain-frequency (Bodé) plot.

VIP:®

A model designation of Kepco, Inc., applied to a group of load protectors: (V) Voltage, (I) Current, (P) Protectors. The VIP devices provide overvoltage, undervoltage and over/under current sensing and protection circuits.

VIX,® INDICATORS:

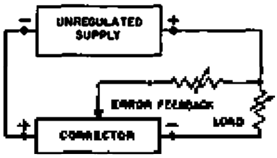
Voltage/Current Crossover Indicators. VIX indicators are a pair of small mode lamps on the front panel of automatic crossover power supplies. One lamp lights during voltage regulated operation of the power supply, the other, during current regulation operation.

VIX SIGNAL:

A keyed voltage, whose polarity is an indication of power supply output voltage/current regulation mode. The polarity abruptly reverses at the crossover point and can be used to actuate external mechanisms such as lamps, alarms, etc.

VOLTAGE CORRECTOR:

An active source of regulated power placed in series with an unregulated supply to sense changes in the output voltage (or current); also to correct for the changes by automatically varying its own output in the opposite direction, thereby maintaining the total output voltage (or current) constant. (*See* Fig. G.13.)



G.13 Circuit used to sense output voltage changes.

VOLTAGE REFERENCE:

A separate, highly regulated voltage source used as a standard to which the output of the power supply is continuously referred.

WARMUP TIME:

The time (after power turn-on) required for the output voltage, or current, to reach an equilibrium value within the stability specification.

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