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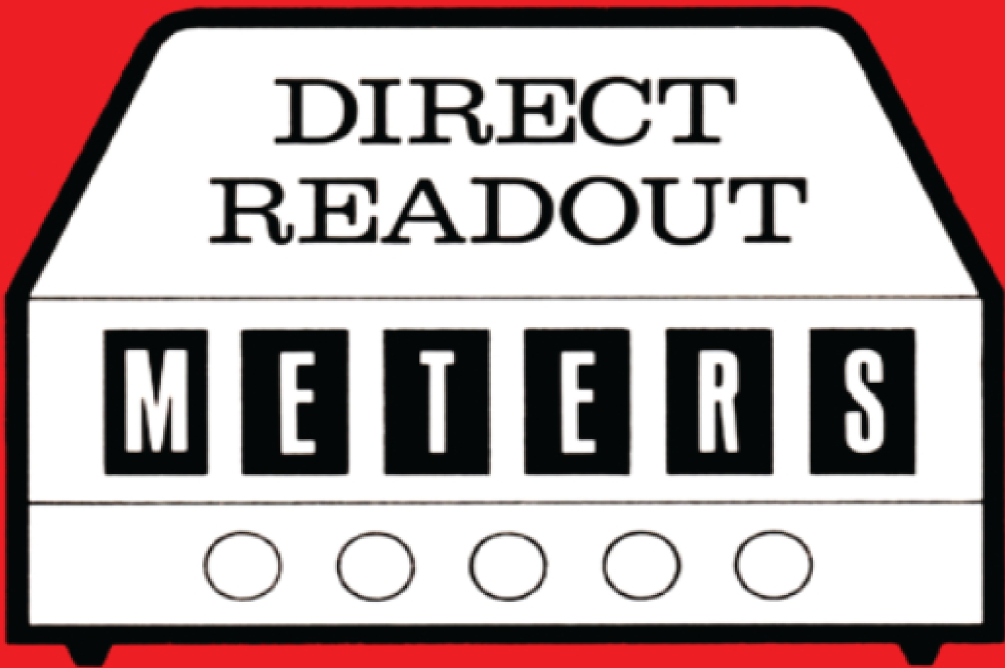


by John D. Lenk

Discusses digital, differential, and special-purpose meters in technicians language—covers operating principles, applications,

# DIRECT READOUT METERS

*by*

John D. Lenk

HOWARD W. SAMS & CO., INC.
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# DIRECT READOUT METERS



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# Preface

Today's laboratory technician must master a host of new-generation meters. While the conventional vom and vtvm fill the need for most shop technicians and home experimenters, industrial electronic measurements require the precision and convenience offered by digital, differential, and special-purpose analog meters. These new-generation meters have been designed to provide the convenience and precision required in industrial electronic measurements today. It therefore behooves the technician to have the information in this book so as to better qualify himself as a laboratory meter specialist, whether the problem is one of testing, servicing, or simply understanding the equipment.

The text treats the subject of laboratory meters in logical progression from the less sophisticated digital-type meters to the more complex special analog types. This approach helps to develop the complexity of the circuits as the reader becomes more familiar with the terms and gains confidence in the subject. The signal sequence is followed whenever possible through each group of circuits. Each network is preceded by a discussion of the basic circuit considerations or problems.

No attempt has been made to include design data or the mathematical equations associated with design problems, except where absolutely necessary to understand particular circuits. However, typical operating limits are provided for each type of meter described.

This introductory book is intended to provide a basic background for modern digital, differential, and analog meters. It explains, in straightforward language, what they are, how they operate, and what they can be expected to do.

John D. Lenk

# Contents

## CHAPTER 4



## CHAPTER 5



## CHAPTER 6



## CHAPTER 7

# 1

# Introduction to Digital Meters

Digital voltmeters or *dvm*'s display measurement in discrete numerals, rather than as a pointer deflection on a continuous scale as commonly used in analog devices. A typical digital voltmeter is shown in Fig. 1-1.

Digital voltmeters have several advantages over analog meters. The direct numeral readout in dvm's reduces human error and the tedious measurement process. Dvm's also eliminate parallax error and increase reading speed. Automatic polarity and range-changing features reduce operator training, measurement error, and possible instrument damage through overload or reversed polarity. Measuring devices for a-c voltages, d-c currents, and resistance are available.

Courtesy Non-Linear Systems, Inc.

**Fig. 1-1. A typical digital voltmeter.**

Permanent recording instruments are available with printers, card and tape punches, and magnetic tape equipment. Data in digital form may be processed with no loss of accuracy.

## BASIC OPERATING PRINCIPLES

The heart of a digital voltmeter is the circuitry which converts analog voltage to a digital form. This is known as analog-to-digital conversion or *adc*. Most digital voltmeters on the market today fall into one of five categories:

1. Successive approximation
2. Continuous balance
3. Ramp (voltage-to-time interval)
4. Integrating
5. Integrating and potentiometric

The basic operating principles of each of the five basic types of digital meters will be given in this chapter. Additional circuit descriptions are covered in later chapters.

## SUCCESSIVE-APPROXIMATION DVM

The successive-approximation type of digital voltmeter converts the input voltage into digital form by a series of approximations and decisions. This type of voltmeter consists of a digital storage register (digital accumulator), a digital-to-analog converter, a comparison network (error detector), a precision voltage reference, and control circuitry. The input voltage is compared first with the most significant bit. The actual comparisons are made successively in binary form. If the input voltage is less than the most significant bit of the reference, the most significant bit of the register is cleared, and the next lower bit is switched in for comparison. The process of switching to the next lower significant bit is continued until a decision is made on all digits.

## CONTINUOUS-BALANCE DVM

The continuous-balance type of dvm performs a digital measurement by comparing the unknown voltage with a voltage derived from a stable reference supply. At the beginning of a measurement the unknown voltage is compared with the "full-scale" reference. If null (balance condition) is not reached, a voltage derived from the reference is reduced by an incremental value representing a unit of the least significant digit. This is done by automatic switching of the precision resistors. This process continues until null is reached.

## RAMP DVM

The ramp (voltage-to-time conversion) category forms the economy class of digital voltmeters. The operating principle of the ramp digital voltmeter is to measure the length of time it takes for a linear ramp of voltage to become equal to the unknown input voltage, starting

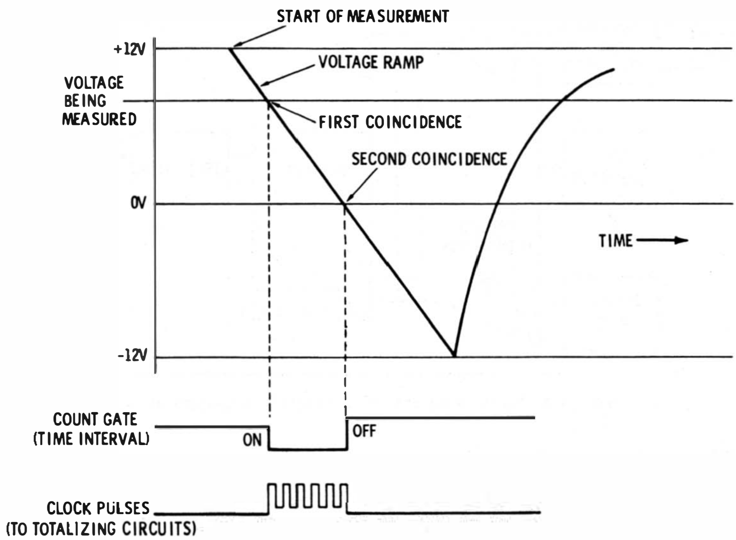



**Fig. 1-2. Timing diagram of ramp-voltage generation.**

from a known level. This time period is measured with an electronic time-interval counter and displayed on "in-line" indicating tubes. Two advantages of this instrument are low price and simplicity. However, it requires an input noise-filter if superimposed noise is present.

Conversion of a voltage to a time interval is illustrated by the timing diagram of Fig. 1-2. At the start of a measurement cycle a ramp voltage is generated. The ramp is compared continuously with the voltage being measured. At the instant the two voltages become equal, a coincidence circuit generates a pulse which opens a gate. The ramp continues until a second comparator circuit senses that the ramp has reached zero volts. The output pulse of this comparator closes the gate. If the input is a negative voltage, coincidence with it would occur before zero coincidence. The circuitry senses which coincidence occurs first and switches the polarity indicator accordingly.

The time between the gate opening and closing is proportional to the input voltage. The gate allows clock pulses to pass into the totalizing circuits, the number of pulses counted during the gating interval being a measure of the voltage. Choice of ramp slope and clock rate enables the totalizing circuit readout to read directly in millivolts (for example, a slope of 300 volts/second and a clock rate of 300 kc).

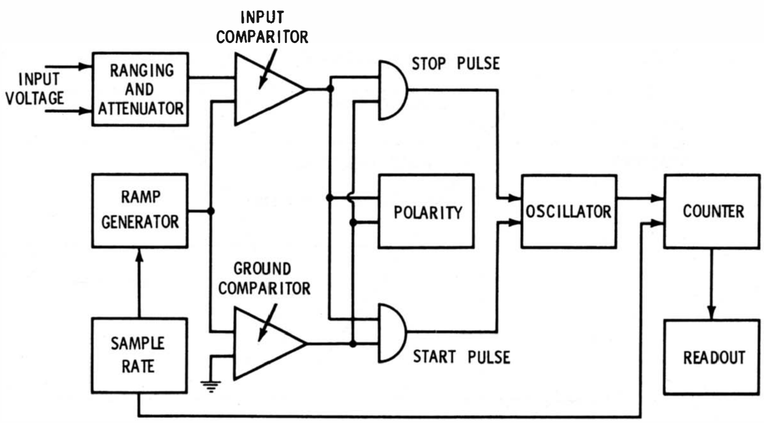



**Fig. 1-3. Block diagram of a typical ramp-type dvm.**

The main advantage of voltage-to-time conversion as a technique for dvm's is its simplicity. Slowly varying input voltages do not disturb the operation of the voltmeter, as often happens with null-seeking voltmeters which may continually hunt for, but never reach, a balance.

A block diagram of a typical ramp-type dvm is shown in Fig. 1-3. Here, a voltage ramp is generated and compared with the unknown voltage, and with zero voltage. Coincidence with either voltage starts the oscillator, and the electronic counter counts the cycles. Coincidence with the second comparator stops the oscillator. The elapsed time is proportional to the time which the ramp takes to travel between the unknown voltage and zero volts, or vice versa. The order in which the pulses come from the two comparators indicates the polarity of the unknown voltage. The accumulated reading in the counter can be used to control ranging circuits.

## INTEGRATING DVM

The integrating (voltage-to-frequency conversion) dvm's represent an advancement over the ramp-type units. The integrating digital voltmeter measures the true average input voltage over a fixed encoding time, instead of measuring the voltage at the end of the encoding time as do ramp-type, successive-approximation, or continuous-balance units. Measurement at the end of the encoding time could easily coincide with a burst of noise, creating an inaccuracy in the ramp-type units.

Conversion of a voltage to a frequency is illustrated by the diagram of Fig. 1-4. The circuitry functions as a feedback control system which governs the rate of pulse generation, making the average voltage of the rectangular pulse train equal to the d-c input voltage.

A positive voltage at the input results in a negative-going ramp at the output of the integrator. The ramp continues until it reaches a voltage level that fires the level detector which triggers a pulse generator. The pulse generator produces a rectangular pulse with closely controlled width and amplitude, just sufficient to draw enough charge from a capacitor to bring the input of the integrator back to the starting level. The cycle then repeats.

The ramp slope is proportional to the input voltage. A higher voltage at the input would cause a steeper slope, resulting in a shorter time duration for the ramp. Consequently, the pulse repetition rate would be higher. Since the pulse repetition rate is proportional to the input voltage, the pulses can be counted during a known time interval

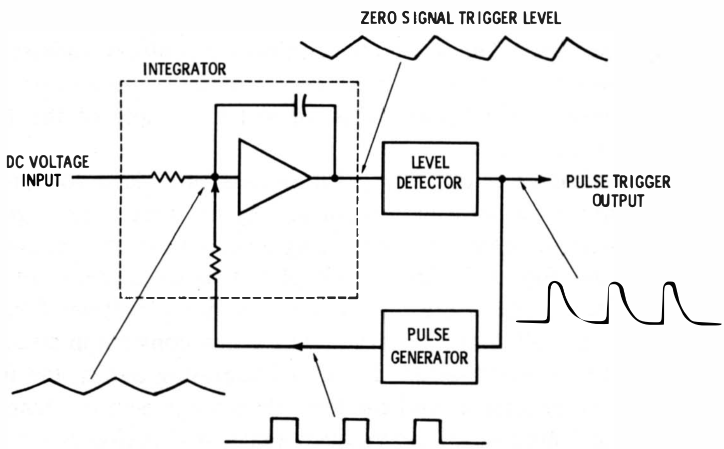



**Fig. 1-4. Network for converting voltage to frequency.**

to obtain a digital measure of the input voltage. Although a voltage ramp is generated in this type of dvm, the amplitude is only a fraction of a volt, and the accuracy of the analog-to-digital conversion is determined not only by the characteristics of the ramp but also by the area of the feedback pulse.

The main advantage of this type of analog-to-digital conversion is that the input is "integrated" over the sampling interval and the reading represents the true average of the input voltage. The pulse repetition frequency "tracks" a slowly varying voltage so closely that changes in the input voltage are accurately reflected as changes in the pulse repetition rate. The total pulse count during a sampling interval, therefore, represents the average frequency and thus the average

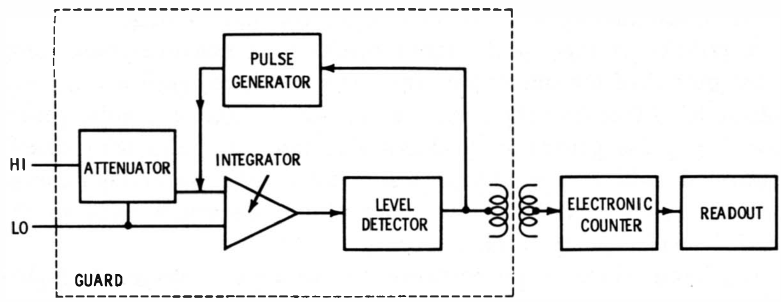



**Fig. 1-5. Simplified block diagram of the guard-circuit technique.**

voltage. This is important when noisy signals are encountered during measurement. The noise is thereby averaged out during the measurement without requiring input filters that would slow down the voltmeter response time. The voltmeter achieves essentially infinite rejection of power-line hum—the most prevalent source of signal noise—when the measurement interval is an exact multiple of the hum waveform period.

Another advantage is that the pulse circuits provide a convenient means of coupling the information out of a guard circuit. Such guard circuits are used to isolate the measuring circuit from the remaining readout circuits. Fig. 1-5 shows a simplified block diagram of the guard-circuit technique. Here the integrating digital voltmeter has a floating input, and all of the voltage-to-frequency conversion circuitry is housed within a guard shield. In other integrating dvm's, the integrator, the pulse generator, and the level detector generate a train of pulses. The total number of pulses over a specified period is directly proportional to the integral of the input signal over this same period.

This arrangement makes it possible to transformer-couple the signal to the digital circuits outside the guard, allowing complete isolation of the measuring circuit itself.

## POTENTIOMETRIC INTEGRATING DVM

The potentiometric integrating dvm combines the features of an integrating dvm and a form of differential voltmeter. (Differential voltmeters are discussed in later chapters.) A conventional integrating dvm measures the true average of the input voltage over a fixed encoding time. A conventional differential voltmeter relies primarily on resistance ratios and a stable reference voltage to ensure accuracy.

It is possible to combine these two basic features, as shown in the block diagram of Fig. 1-6. Such a voltmeter is divided into three sections: a voltage-to-frequency (v/f) converter, a counter, and a digital-to-analog (d/a) converter. Readings are taken in two steps. First, the v/f converter generates a pulse train with a rate proportional to the

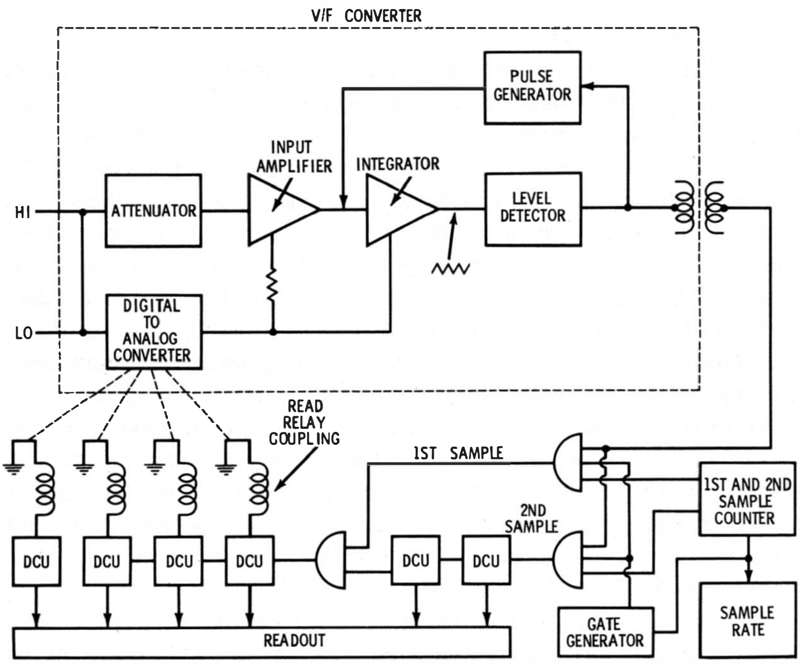



**Fig. 1-6. Block diagram combining features of resistance ratio and stable reference voltage.**

input voltage. This pulse train is gated for a precise time interval and is fed to the first four places in a six-digit counter. The stored (undisplayed) count is transferred to the d/a converter, which produces a highly accurate d-c voltage proportional to the stored count. This voltage is subtracted from the unknown voltage at the input to the v/f converter.

The next step in operation occurs when the pulse train from the v/f converter is again gated a second time going to the last two places in the six-digit counter. At the end of the second gate period the total count is transferred to the six display tubes. The counter display is indicative of the integral of the input voltage.

## ELECTROMECHANICAL DIGITAL METERS

In addition to the all-electronic digital meters, there are a number of electromechanical units in current use. These can be classified into four general types—stepping switch, relay, analog-servo, and stroboscoping.

### Stepping-Switch Type

The stepping-switch instruments are of the comparison type in which a stepping-switch–operated voltage divider creates a feedback voltage equal to the input voltage (see Fig. 1-7). The two major differences between the various stepping-switch digital voltmeters are the sequence (logic) in which they develop the feedback voltage and whether the stepping switches are run dry or kept continuously lubricated by immersion in a bath of oil. The difference is extremely important. The proper type of logic ("scan" logic rather than "tracking" logic; see Glossary) and oil immersion can greatly increase instrument life. The "scan" logic also increases measuring speed and eliminates hunting, which is a disadvantage of the "tracking" instruments.

More stepping-switch digital voltmeters have been produced than any other kind. They are low priced with respect to their accuracy, which is on the order of ±0.01 percent of full scale. They are simpler in construction than any other instrument in this accuracy class and are relatively simple to maintain, providing such features as plug-in stepping switches are incorporated.

The average time per measurement of a four-digit instrument with tracking logic is 1.0 second compared with 1.9 seconds maximum for a four-digit instrument with scan logic. An advantage of the stepping-switch type of instrument is that electrically isolated digital outputs

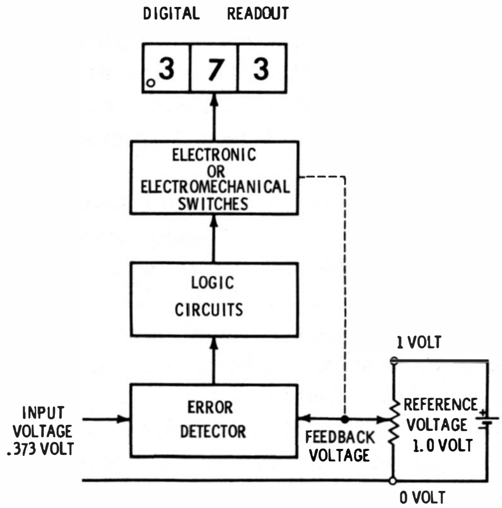



**Fig. 1-7 Voltage divider for creating a feedback voltage equal to the input voltage.**

in contact-closure form are obtained by adding another deck to each stepping switch.

## Relay Type

The relay-type instruments are comparison units in which a relay-operated voltage divider creates a feedback voltage equal to the input (Fig. 1-7). Relay-type digital voltmeters are capable of higher speed than the stepping-switch type since the relays can be operated at a higher speed (100 steps per second rather than 30 steps per second). They can be operated in any sequence (in a stepping switch, the switching sequence is fixed). Some relay-operated instruments make three measurements per second, which compares favorably with the speed of some "all-electronic" instruments having automatic ranging.

## Analog-Servo Type

In the analog-servo type of instrument a feedback voltage is created by a motor-driven variable resistance (potentiometer). Its functional diagram (Fig. 1-8) resembles that of a servo system used for positioning an output shaft (position servo). That is, a balance detector compares the input and feedback voltages and issues commands which are amplified to drive the motor. The motor drives the variable resistance in the proper direction to make the feedback voltage equal to the input. When the feedback and input voltages are equal, the balance-detector output voltage ideally drops to zero, and the motor stops. Voltage readout is in terms of motor shaft position and is displayed by a calibrated drum or several numbered drums, each geared to the adjacent drum by a mechanical transmission having a 10:1 drive ratio.

Analog-servo digital voltmeters are in the same speed class as the slowest stepping-switch instruments. Usually, initial accuracy is about 0.1 percent to 0.5 percent of full scale. Initial accuracies on the order of ±0.01 percent can be achieved at high cost by using a very precise, motor-driven variable resistance. Digital output, for driving printers and other output accessories, can be obtained if an analog-to-digital shaft-position converter is coupled to the drive motor; this will, however, increase overall instrument cost significantly. Automatic ranging is not normally available with analog-servo type instruments.

### Stroboscopic Type

This stroboscopic-type device is somewhat similar to the ramp-type all-electronic instrument. It generates a ramp voltage by continuously rotating the wiper of a variable resistance (potentiometer) which is energized by a d-c reference voltage (Fig. 1-9). The ramp voltage

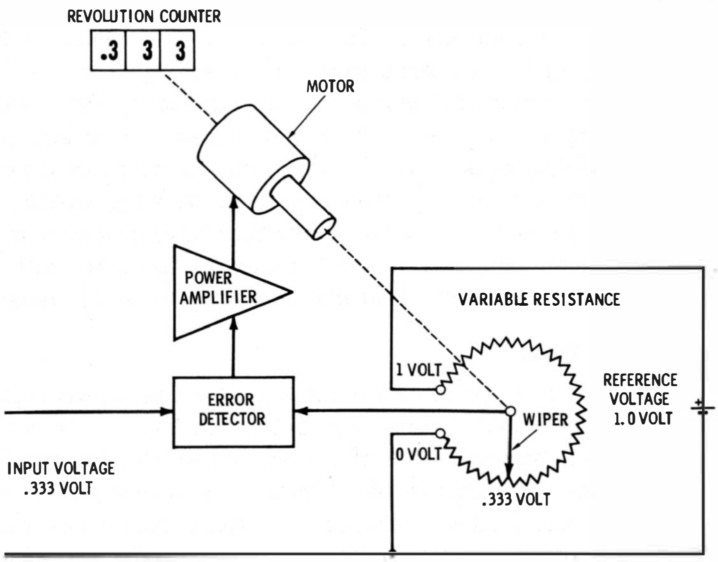



**Fig. 1-8. Functional diagram of the analog-servo type of instrument.**

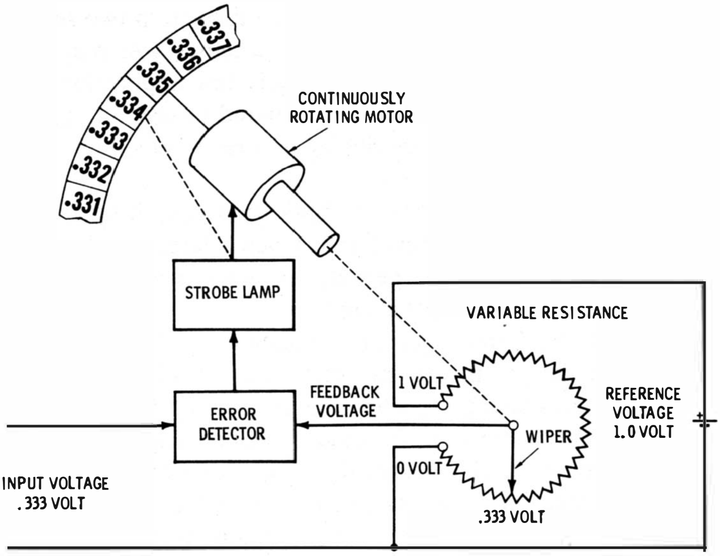



**Fig. 1-9. Generation of the ramp voltage.**

is compared continuously with the input signal so that each time the two voltages are equal a lamp is flashed, illuminating a numbered drum coupled to the shaft of the rotating pot. The numbers aligned with a small window in front of the drum are made visible by the flashing lamp at the time of voltage equality. The drum speed is high enough (1500 rpm) to produce numbers which appear to be stationary.

The price of stroboscopic instruments is relatively low. Readout numeral size is limited by the overall instrument size. Initial accuracy is in the order of 0.5 percent of full scale. However, the high-speed rotation of the variable resistance wiper poses a life problem, and digital output and automatic range changing are not offered in these instruments. A relatively low input impedance (1 megohm) is provided.

## SELECTING A DIGITAL METER

There are many factors that should be considered when selecting a digital meter. The following is a summary of these factors.

### Measuring Speed

Of course, all digital voltmeters are very fast compared with pointer meters and manually balanced potentiometers. But speed varies among the basic types, ranging in balancing time from two seconds to several microseconds. Consider carefully whether you require more than one reading per second since relatively few applications require such balancing speed. Unless you are prepared to spend thousands of dollars, you will generally be sacrificing accuracy, reliability, or versatility to achieve the higher speed.

There are two general types of applications requiring high-speed measurements. One such application is when a large number of readings must be made in a short period; the other is when one reading must be made at a definite instant in time.

Examples of the former type of application are defining the waveshape of transients and measuring many channels of data quickly before test conditions change or when the quantity of data is so large as to require an inordinate time for data collection (applications which involve sampling 1000 channels of data require more than 17 minutes per complete cycle for a one-reading-per-second digital meter).

Examples of the latter type of application include measuring a varying voltage at a definite instant in time so that voltage data can be correlated with other data such as time, distance, etc., as in radar tracking stations. For this type of application, a special digital voltmeter known as a *sample-and-hold* type should be considered. Sample-and-hold models are designed to "follow" input variations until commanded to read, whereupon they hold the input voltage constant (internally) for that instant while it is measured. Such digital voltmeters can provide much greater accuracy in this application than very high speed digital meters can without the sample-and-hold feature.

Reducing overall speed requirements can sharply reduce the cost and complexity of data logging systems and reduce procurement lead time, too. For example, a data printer capable of five recordings per second costs approximately $1500, while one capable of 300 recordings per second costs from $30,000 to $50,000. The cost of high-speed input scanners and other data logging input and output devices rises sharply with speed.

Consider the fact that the digital voltmeters with tracking logic have a very long balancing time when the input signal varies, unless they are desensitized–in which case accuracy is reduced.

### Accuracy

Obviously, the manufacturer's statement of accuracy is not always sufficient. Basic questions include the following: How often must you correct for long term drift in calibration or zero point? When subjected to varying input signals, will the instrument display a value that the input signal actually had during the balancing cycle? Instruments with scan logic will; instruments with tracking logic may not.

When selecting the number of digits needed in a digital voltmeter, consider the limitations of digital resolution. In a four-digit digital meter, a reading at 20 percent of full scale (2000 volts, 20.00 volts, 200 volts) can only be accurate to $\pm 1$ digit, which is one part in 2000 or 0.05 percent of the value read. A reading at 10 percent of full scale (1000 volts, 10.00 volts, 100.0 volts) can also be accurate to $\pm 1$ digit, which is one part in 1000, or 0.1 percent of the value read. Therefore, it is advisable to select a five-digit meter if close to 0.01-percent accuracy is required over most of the measuring range. Select a three-digit meter only if one-percent accuracy is acceptable.

### Reliability

Reliability cannot always be judged on the basis of price alone. Rather, look closely at the type of construction and the more subtle clues—quality of components and derating of components. For example, does the stepping-switch type meter have stock telephone switches or the type specially designed for digital voltmeter use? Do switches have phenolic insulation or diallyl phthalate, which protects the instrument accuracy against the effects of moisture? Are there excessive solder joints and cable connections? Are long-life, mercury-wetted contact relays used? Often it is wise to either check with present users about the reliability which they have observed with the instrument or request all manufacturers to submit a sample for test.

### Versatility

Suppose that you require voltage-ratio and resistance measurements, in addition to voltage measurements, and also require data recorder operation. A multipurpose instrument with built-in circuits designed to perform all of these functions may be lower in cost, more reliable, and much more convenient to use. One which was designed as a digital voltmeter and requires wiring changes or additional modules to convert it for voltage-ratio and resistance measurements, or to convert it for use with data recorders, may be cumbersome.

### Ease of Operation

An inherent advantage of digital voltmeters is the ease of operation; however, some are easier to operate than others. For example, some instruments do not require desensitizing if the input signal varies during measurement—their logic always produces a reading as accurate as the voltage is stable.

### Ease of Servicing

The ease of servicing is an often overlooked factor which deserves major consideration. Some questions to ask here are:

How long does it take to replace a stepping switch?
How much of the construction is plug-in?
How fast can a readout bulb be replaced?
Are spare parts and assemblies readily available?
Does the manufacturer offer maintenance training locally, or must you travel across the nation to get it?
Does the manufacturer offer servicing at regional offices as well as at the main plant?

## DIGITAL VOLTMETER LOGIC

Digital meter logic is the sequence in which the digital voltage-divider switches or circuits and the range and polarity circuits operate in making a measurement. Logic is important because it affects meter speed, life, and response. In digital ratiometers reaction to reference-voltage changes is also important.

The two forms of logic used in most digital meters are *successive trial* or *scan* logic and *tracking* or *digital servo* logic.

### Scan Logic

The distinguishing feature of feedback-voltage creation in scan-logic meters is that the output voltage of each decade is scanned, or sampled, in a definite sequence, and the decade switches are operated once per measurement, starting with the leftmost digit and progressing to the right. Some digital meter manufacturers (such as Non-Linear Systems, Inc.) designate this type of logic as "no-needless-nines" logic when used in stepping-switch meters because it eliminates needless cycling of the switches through their nine and zero positions. Scan logic is also frequently used in relay and all-electronic digital meters.

### Double Scan Logic

This form of scan logic (also called *double duty no-needless-nines logic* or *tracking scan* logic—not to be confused with "tracking" logic) is of particular benefit in stepping switch meters. It affords such meters all the benefits of ordinary scan logic, plus the one operating advantage that tracking logic meters have. This is the capability of displaying a one-digit increase in voltage very quickly (in ordinary scan logic, the positions of the polarity and range switches or circuits, and all leftmost decade switches must be sampled—but not changed—before the rightmost decade switch position can be changed). In starting a measurement with a double scan-logic meter, the logic first determines (without moving stepping switches) which decade or decades must be changed. Then, numerical changes are made, starting with the leftmost decade, which requires a change and progressing to the right. Each stepping switch cycles no more than once per reading; if a switch does not have to change, it does not cycle.

For example, consider a change from 9.326 to 9.327. In double scan logic the 6 can be changed to a 7 almost immediately because in double scan logic the rightmost decade is sampled before all others. In ordinary scan logic, the polarity switch position, range switch positions, and the three leftmost decades must all be sampled (but not changed) before the 6 can be changed to a 7. However, the double scan logic still has all the advantages of ordinary scan logic: no decade switch will cycle more than once per reading; only those decade switches that must be changed to create new numbers will step; the instrument will not hunt when the input varies during measurement, and the meter automatically adapts to a wide range of reference voltage for voltage-ratio measurements. Best of all, no adjustments are necessary.

### Tracking Logic

Despite the many advantages of the scan logic, tracking logic is still used in low-cost digital meters. The most characteristic feature of tracking logic is that decade switches are always free to follow any input signal change that occurs (unless the decade switches are turned off) and that all of the readings are then computed starting with the least significant (rightmost) decade and progressing from the right to the left.

If the input voltage is greater than the feedback voltage, the meter error detector issues "up" pulses to stepping switches which operate each decade of the feedback divider. If the feedback exceeds the input, the error detector issues "down" pulses to these same stepping switches which are on a separate logic-pulse line. (See Figs. 1-10 and 1-11.)

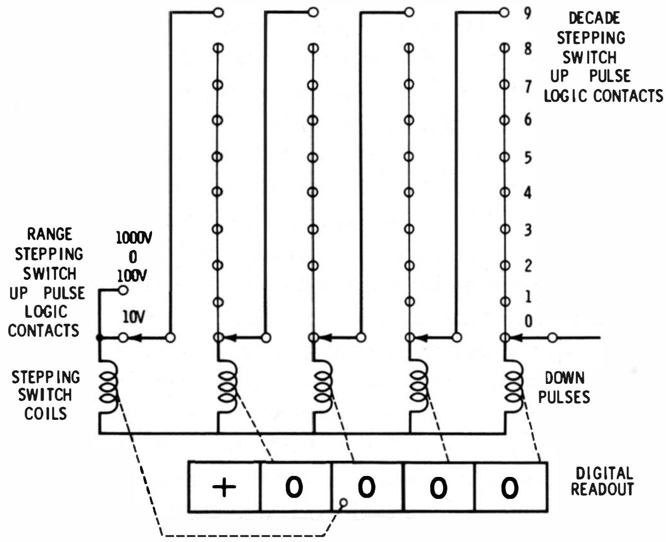



Fig. 1-10. Routing of "up" pulses.

## DIGITAL VOLTMETER ACCURACY STATEMENTS

The accuracy statements of digital voltmeters are somewhat different from those of conventional analog meters. "Percent of full scale" is usually synonymous with "percent of full scale of range in

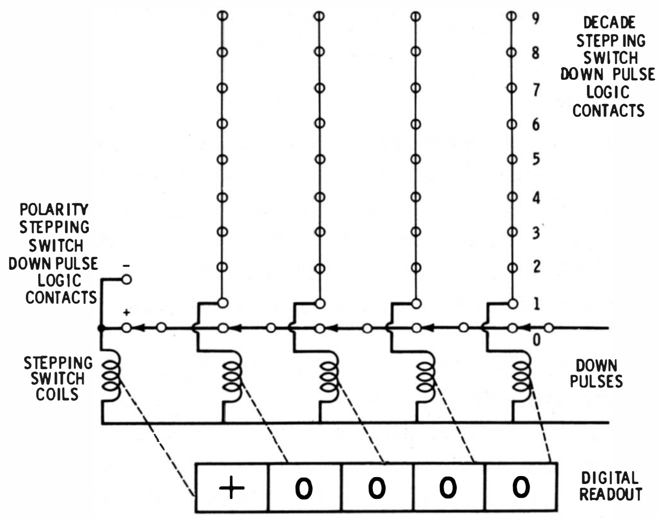



Fig. 1-11. Routing of "down" pulses.

use." These terms indicate an error that is of fixed amplitude over each range. An error of "0.01 percent of full scale of range in use" is an error of 0.001 volt for any reading on the 9.999-volt range, 0.01 volt on the 99.99-volt range, etc.

An error expressed as "percent of reading" indicates an error whose absolute value is governed by the value of the measurement. An error of "0.01 percent of reading" means an error of 0.0009 volt in a 9-volt measurement, 0.00009 volt in a 0.9-volt measurement, etc.

An error of "one digit" also called "one least count" or "one count" means an error the size of the smallest increment displayable by the digital instrument. A one-digit error means an error of 0.001 on the 9.999 range, 0.01 on the 99.99 range, etc.

Errors must sometimes be expressed in mixed quantities. To visualize the total size of the error, convert to some common quantity, such as volts. For example, in measuring 5.000 volts with a four-digit meter whose accuracy is "0.01 percent of reading plus one digit," the maximum error is 0.01 percent of 5 volts, plus 0.001 volt, or a total of 0.0015 volts.

# 2

# Introduction to Differential Meters

The basic function of differential voltmeters or differential voltage measuring is to apply an unknown voltage against one that is accurately known, and to measure the difference between the two on an indicating device. If the known voltage is adjusted to the exact potential of the unknown voltage, one can determine the unknown quantity being measured as accurately as the known voltage (or reference standard).

A typical differential voltage measurement is shown in Fig. 2-1. The null meter ($M_x$) indicates when the voltage ($E_{ac}$) at the potentiometer is equal to the unknown voltage, $E_x$. By considering the ratio of resistances $R_{ab}$ and $R_{ac}$ in the potentiometer, the ratio of $E_{ac}$ to the known reference voltage ($E_{ref}$) may be determined precisely.

The potentiometric method of voltage measurement is highly accurate since precise resistance ratios can be determined. No current is drawn from either the standard cell or the unknown voltage source at null. The source impedance, $R_x$, therefore does not affect measurement. The divider output tap (b) is adjusted to null $E_{ref}$.

The potentiometric method of voltage measurement (see Fig. 2-1) can be represented by the equation

$$E_x = E_{ref} \frac{R_{ac}}{R_{ab}}$$

where,

$E_x$ is the unknown voltage,
$E_{ref}$ is the reference voltage,
$R_{ac}$ is the full potentiometer resistance,
$R_{ab}$ is the potentiometer resistance from slider to one end.

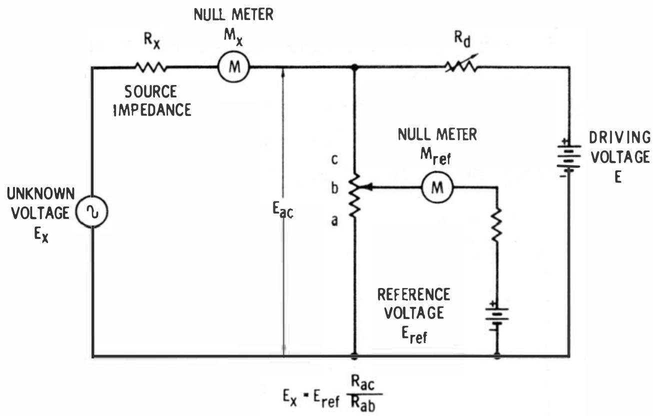



**Fig. 2-1. Potentiometer method of measuring unknown voltages.**

For example, if the reference voltage were 5 volts, the total potentiometer resistance 1000 ohms, and the slider resistance for null 200 ohms, then

$$E_x = 5 \frac{200}{1000} \quad \text{or} \quad 5 \times \frac{1}{5} = 1 \text{ volt}$$

## BASIC OPERATING PRINCIPLES

Fig. 2-2 illustrates a simplified, conventional differential voltmeter in which the potentiometer slider has been replaced by a Kelvin-Varley divider, and the driving or source voltage E has been replaced by an accurate supply, $E_s$, referenced to a standard voltage $E_{ref}$. In practice, the null device is a solid-state voltmeter rather than a galvanometer.

Using the method shown in Fig. 2-2, a high-voltage standard is necessary to measure high voltages. This need may be overcome by inserting a voltage divider between the source and the null-meter circuit. However, this practice provides a relatively low input impedance for voltages that are higher than the reference standard. This low input impedance is undesirable; accurate measurements can not be obtained if current is drawn from the source that is being measured. Most differential voltmeters used today offer an impedance approaching infinity only at null condition and then only if an input voltage divider is not used.

One way to eliminate this problem is to isolate the measuring circuits from the input by an amplifier. Such an arrangement provides a high input impedance which eliminates loading the source, resulting in a differential voltmeter whose input impedance is independent of both null condition and range.

Differential voltmeters are often combined with d-c voltage standards. This approach produces one package that provides several benefits for making precision voltage measurements.

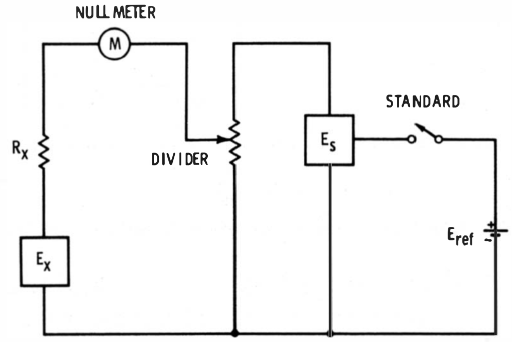



**Fig. 2-2. Simplified conventional differential voltmeter.**

## TYPICAL DIFFERENTIAL VOLTMETER OPERATION

Fig. 2-3 is a block diagram of a differential voltmeter that has been combined with a d-c voltage standard and includes an isolating amplifier at its input. The circuit also has an a-c/d-c converter that permits it to make differential measurements on a-c voltages (by precise conversion of the unknown a-c voltage to an equivalent d-c voltage) and to serve as a precision a-c voltmeter.

When a voltage to be measured is applied to the input terminals, the amplifier responds by re-creating this voltage at the summing point inherently balancing the system to achieve a constant, high input impedance. The amplifier output is converted to a 1-volt level by means of a precision range divider switch for direct comparison with a 1-volt internal reference. The range switch performs two operations. First, it changes the overall feedback factor and thus the overall amplifier gain. Second, the range switch selects the potentiometer tap on the range stick. Consequently, the choice of the proper range enables any input voltage between 0 and 1000 volts to be represented by a proportional voltage between 0 and 1-volt at the tap connecting to the potentiometer.

The isolation stage consists of a series-connected pair of amplifiers, a low-level chopper-stabilized amplifier, and a high-voltage amplifier.

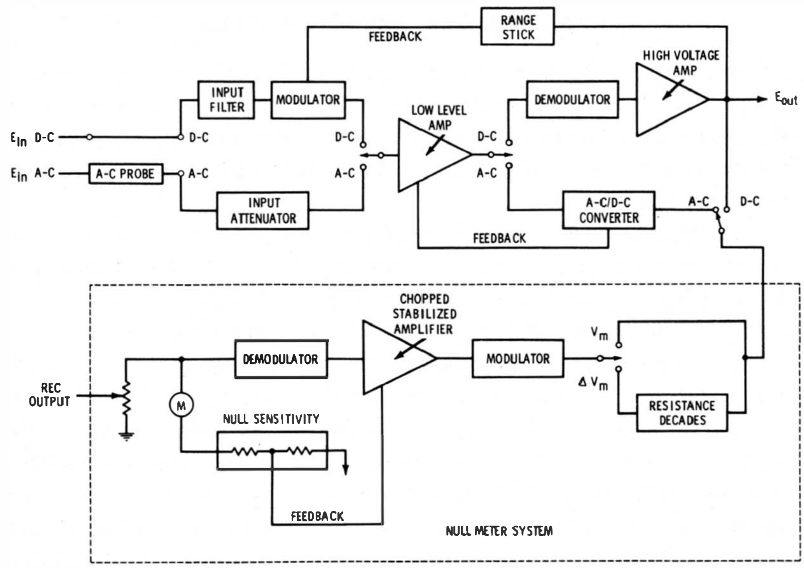



**Fig. 2-3. Block diagram of an a-c/d-c differential voltmeter.**

The high level of feedback (over 100 db) ensures gain accuracy and produces a high input impedance (over 1000 megohms).

The a-c probe is used with a-c measurements. The unknown a-c signal is fed from the precision attenuator to the high-impedance, low-level amplifier. It is then amplified and applied to a diode bridge whose half-wave signal is averaged to produce a d-c output. The d-c output of the converter is measured in the same manner as the d-c different mode—by connecting the null-meter decade divider system. Consequently, the a-c measurement is made with a high degree of accuracy, since the a-c signal is converted to d-c and measured with the same precision dividers and reference standard used in d-c operation. With this method, accuracy can approach that of the d-c mode.

The differential voltmeter can be converted into a d-c standard by means of a front panel control. The same feedback amplifier and range stick are used. However, the internal reference supply now becomes the input to this amplifier. The input is automatically nulled against a feedback voltage which is determined by the range-switch setting and the output voltage at the sensing terminals. A range switch on the decade divider is mechanically linked to a voltage divider connected to the input of the null meter. This vernier provides the fifth and sixth digits of resolution.

## COMMON-MODE SIGNAL PROBLEMS

One of the problems in any voltmeter—especially in those designed for precision measurements such as differential voltmeters—is an effect known as *common-mode insertion*. Therefore, in any precision laboratory-type meter, the common-mode rejection factor becomes an important characteristic.

To understand the common mode problem, assume that we have an instrument with two input terminals, 1 and 2. The instrument ideally responds to the voltage between terminals 1 and 2. This voltage is called the *differential voltage*. If terminal 1 is at +0.1 volts and terminal 2 is at −0.1 volts, the differential voltage applied to the instrument is 0.2 volts. Again, if terminal 1 is at +10.0 volts and terminal 2 is at +10.0 volts, the differential voltage applied to the instrument is zero, and the instrument should respond accordingly. In certain situations, however, when equal +10-volt input signals are applied, the instrument will not respond as if the differential voltage were zero. That is, the instrument's output will indicate that a differential voltage is present. It is apparent that the application of +10 volts to both input terminals may in some way adversely affect operation of the instrument. The +10-volt signal is called the *common-mode voltage* since is it common to both input terminals in terms of amplitude, polarity, and time.

While d-c voltages have been used in the preceding discussion, the same treatment applies for alternating current. Since any voltmeter should ideally respond only to differential voltages, it should ignore the common-mode voltage. The means by which a common-mode voltage may affect an instrument is best shown by means of an example. A typical instrumentation set (simplified) is shown in Fig. 2-4. A transducer (typically a welded thermocouple) is connected to a remotely located instrument (containing a d-c amplifier.)

As shown, a signal voltage causes a signal current to flow in the input resistance via the transducer internal resistances (R1, R2). The instrument has an impedance (Z) to local ground (earth or rack ground). A common-mode voltage exists between local and transducer grounds. The common-mode voltage exists between local and transducer grounds. The common-mode voltage causes a common-mode current to flow through the transducer internal resistances and then through the impedance of the instrument to local ground via the signal leads. It can be seen that the common-mode current is intermingled with the signal current and thus causes a spurious signal

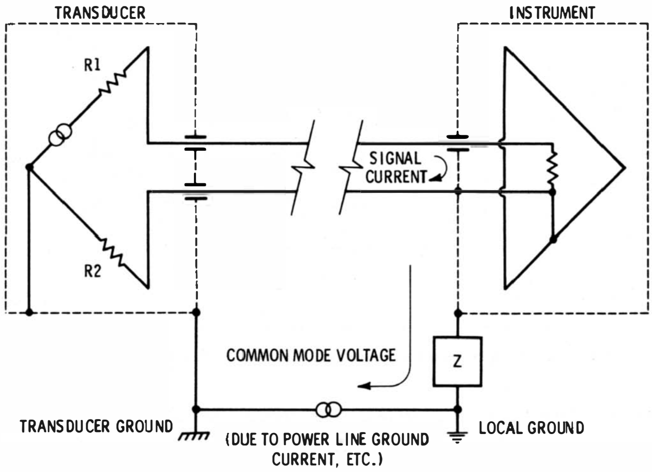



**Fig. 2-4. Common-mode signal insertion.**

current to flow through the instrument's input resistance. This spurious signal current is a source of error.

Capacity of an instrument to reject a common-mode signal and thus reduce the spurious signal current is called *common-mode rejection* (cmr). Cmr is usually specified in db at some frequency or range of frequencies, for example, 130 db minimum at 60 cps. This means that the spurious signal effect of a given common-mode signal is reduced by 130 db or more. For example, the effect of 100-volt, 60-cps, common-mode signal is reduced to that of a 33-$\mu$v maximum equivalent signal. It may be said that a 100-volt, 60-cps, common-mode signal results in a 33-$\mu$v maximum differential signal.

One of the major sources of common-mode signals is induced ground currents, usually at the a-c power-line frequency. These signals can generate a potential of several volts between the signal source ground and the chassis ground. Unless shunted, these currents will cause a voltage to appear at the input or output which can be larger than the signal itself, resulting in an erroneous reading or output voltage. In differential voltmeters and many digital meters (especially those having a differential or comparison circuit) the input and output terminals are often "floating." That is, they are mounted separately from the main chassis. Any a-c voltages present at these terminals can be rejected by means of a circuit ground.

Such a circuit is shown in Fig. 2-5. This circuit is used by Hewlett-Packard in the Model 740A, which is both a d-c standard and a

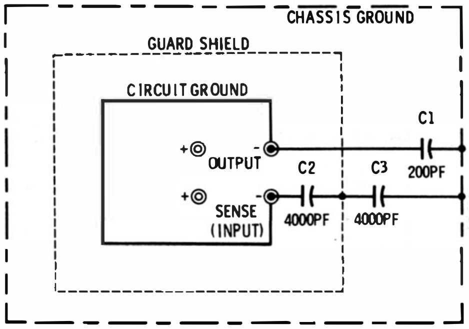



Fig. 2-5. Guard system used in Hewlett-Packard Model 740C.

differential voltmeter. As shown, the negative terminals of both input and output are connected together. The output negative terminal is connected to the chassis through capacitor C1. The input negative terminal is connected to the guard shield (a metal enclosure around the floating terminals) through capacitor C2. In turn, the guard shield and chassis are connected by capacitor C3. This shunts any a-c potentials that may exist between the terminals and chassis, terminals and guard, or guard and chassis.

Another guard circuit used by Vidar, Inc. in their Model 510 is shown in Fig. 2-6. A comparison of this setup with the arrangement

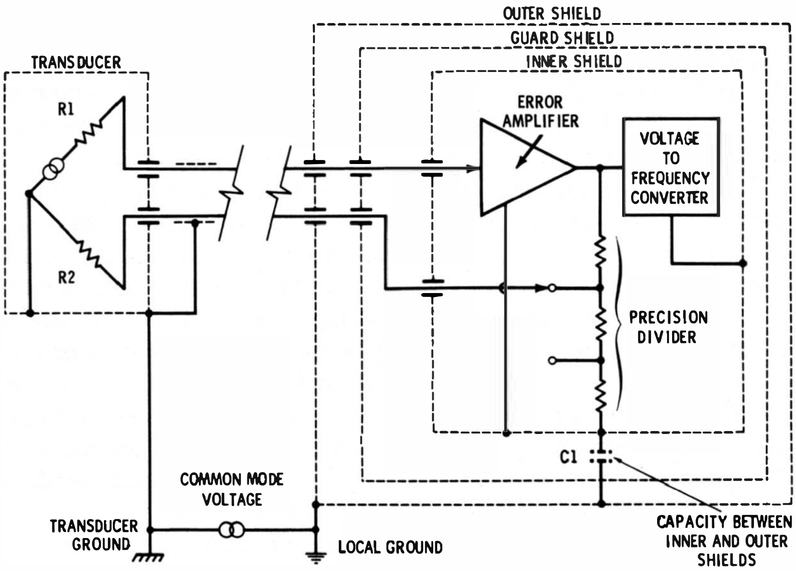



Fig. 2-6. Guard system used in Vidar Model 510.

shown in Fig. 2-4 reveals that the common-mode path has been broken by means of the partitioning between the inner and outer shields (the instrument proper is enclosed by the guard shield.) In this way the instrument-to-local-ground impedance (shown as Z in Fig. 2-4) has been raised to a very high value. The effect of raising the instrument to local ground impedance is to reduce the spurious signal current due to the common-mode voltage. The higher the instrument-to-local-ground impedance is, the lower is the spurious signal current.

It should be noted that the instrument-to-local-ground impedance consists primarily of the small (but not negligible) capacitance between the inner and outer shields. This means that the instrument-to-local-ground impedance will be a function of frequency and that the common-mode rejection will be a function of frequency—the common mode rejection decreasing by 6 db per octave as the common-mode signal frequency increases.

In most cases, this effect does not pose a problem, since the predominant common-mode signal usually occurs at line frequency (50/60 cps).

# 3

# Introduction to Analog Meters

The conventional shop-type vom and vtvm are analog meters. That is, they use rectifiers, amplifiers, and other circuits to generate a current proportional to the quantity being measured. This current, in turn, drives a meter movement. Laboratory-type analog meters also operate in this manner. However, laboratory meters include many circuit refinements to improve their accuracy and stability. Many of the special-purpose analog meter circuits are unique to laboratory equipment. This chapter describes the basic principles of analog meters and shows how these basic principles are adapted to laboratory use.

## METER MOVEMENTS AND SCALES

The meter movements in most shop-type equipment consist of a pointer attached to a coil supported by pivots and jewels. In high-accuracy meters, a taut-band suspension is substituted in place of pivots and jewels. The moving coil in the taut-band meter mechanism is suspended on a platinum-alloy ribbon, eliminating friction and problems concerning repeated measurements. In order to eliminate tracking error on mass-produced meters, the meter faces are custom-calibrated and photographically printed to match exactly the linearity characteristics of each individual meter movement at all points. By combining taut-band suspension with custom-calibrated scales, the chance for mechanical error is kept to an absolute minimum.

## BASIC D-C MEASUREMENTS

The d-c voltmeter represents a straightforward application of electronics to measuring instruments. In laboratory work, d-c voltages are usually measured by a vtvm rather than a vom. Most vtvm's have a direct-coupled amplifier preceding the meter movement. Fig. 3-1 is a basic vtvm circuit for measurement of d-c voltage. The amplifier performs two important functions. First, it increases the input im-

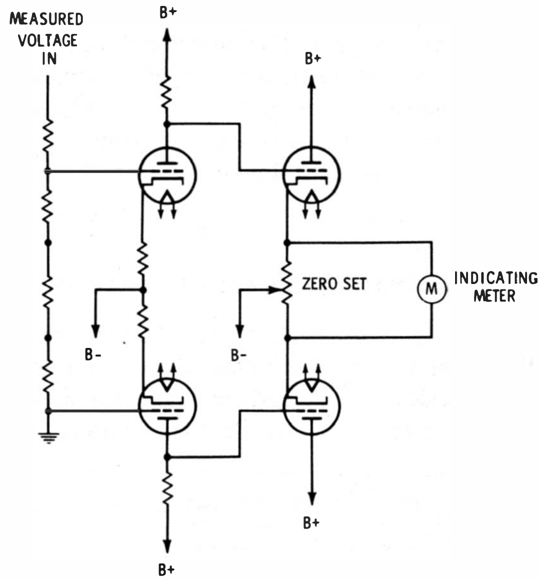



Fig. 3-1. Basic d-c vacuum-tube voltmeter circuit.

pedance of the meter so that the instrument draws negligible current from the circuit under test. Because of the amplifier the electronic voltmeter is a voltage-operated device, whereas the simple meter movement (vom) is a current-operated device. This distinction is important, since the voltage in many circuits can be altered by the current required for operating a meter movement alone.

A second amplifier function is to increase the effective sensitivity of the meter movement. An amplifier changes the measured quantity into a current of sufficient magnitude to deflect the meter movement. An amplifier also limits the maximum current applied to the meter movement. Therefore, there is little danger that unexpected overloads will burn out the meter movement.

## ELIMINATING DRIFT

One of the problems common to the basic vtvm circuit is drift. This is especially aggravated on low-level measurements. A widely used technique for eliminating drift in such low-level measurements is to convert the d-c signal into a comparable a-c signal by alternately applying and removing the d-c signal. The resulting "chopped" signal is amplified in a-c amplifiers and then synchronously rectified at a high level for operating the meter movement. Overall d-c feedback ensures accuracy of the d-c gain. Therefore, d-c drift is limited to a value set by the input chopper. The photoconductive chopper is often used in precision laboratory-type electronic voltmeters. In such a circuit the d-c signal is converted to a comparable a-c signal by illuminating a group of photoconductive or photosensitive resistors periodically. This results in a low-noise, high-impedance chopper action.

## AUTOMATIC RANGING CIRCUITS

Automatic ranging is usually limited to digital meters. The great majority of analog meters (including most laboratory equipment) require that the scales be changed manually for measurements that vary widely. However, it is possible to combine the touch-and-

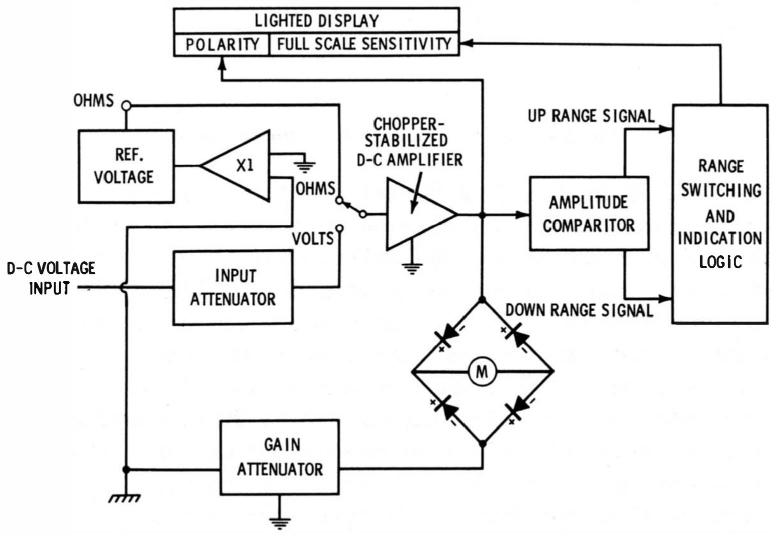



**Fig. 3-2. Block diagram of Hewlett-Packard Model 414A automatic ranging voltmeter.**

read convenience of a digital meter with the economy of an analog instrument. With such a circuit, both range changing and polarity selection are automatic. This provides rapid "hands free" measurement of both voltage and resistance. Such a circuit is shown in Fig. 3-2. A chopper-stabilizer d-c amplifier, input attenuator, gain attenuator, and metering circuit form the basic circuit. Range-changing decisions are indicated by means of a lighted display and are based on two signal levels, one near full scale on a given range and the other at one-fourth of full scale. An amplitude comparator produces an "up-ranging signal" whenever the input voltage tends to rise above the level which is near full scale and a "down-ranging signal" whenever the input voltage tends to fall below the level near one-fourth of full scale. Range switching and indicating logic are a set of four multivibrators which define the twelve ranges of the instrument.

## BASIC D-C CURRENT MEASUREMENTS

In the basic vom the meter movement (by itself) serves the purpose of measuring appreciable amounts of direct current. In these cases the meter coil requires relatively few turns to generate sufficient magnetic flux for deflecting the meter pointer. For lower current measurements the sensitivity of the meter movement must be increased, usually by using more turns in the coil. These added turns increase the resistance of the current path. This can be troublesome in low-impedance circuits. The laboratory-type electronic current-measuring instruments overcome this difficulty by measuring the small voltage drop across a low-value resistance placed in series with the current to be measured. Most laboratory meters are equipped with internal, calibrated, shunt resistors for reading direct currents, in this way, without accessory equipment.

## CLIP-ON METERS

Current measurements using a series resistor have the obvious disadvantage of interrupting the circuit under test. In many applications, insertion of a resistance in the current path may alter the current being measured or even alter the circuit operation. To overcome this difficulty, clip-on milliammeters use current probes which simply clip around the current-carrying wire and measure direct currents. A typical clip-on meter will measure currents from 0.1 milliampere to 10 amperes.

These probes use the second-harmonic flux-gate principle to sense magnetic flux around the wire. (See Fig. 3-3.) The probe encircles the wire with a magnetic core which is saturated periodically by a 20-kc driving current. Saturation interrupts the magnetic circuit, effectively "gating" any flux induced in the core by current in the wire. This gated a-c flux couples with sensing coils on the core, inducing a 40-kc voltage proportional to the current in the wire. The circuitry amplifies the coil voltage and drives the indicating meter accordingly. High linearity is achieved by using negative d-c feedback current, balancing the input ampere-turns against the feedback ampere-turns.

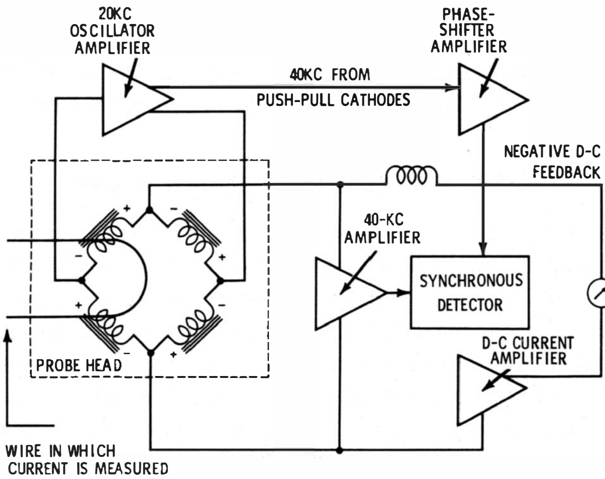



**Fig. 3-3. Block diagram of the Hewlett-Packard Model 428B clip-on type probe.**

The clip-on probes are finding wide use in solid-state circuit measurements where current has to be monitored carefully. Sensitivity is such that base current can be measured. There are a variety of other uses, such as measuring the current in ground loops, where the impedance is too low for the series-resistance technique to be applied.

A unique feature of these probes is that the sums and differences of currents in several wires can be determined by running the wires through the probe at the same time. This technique is useful for balancing push-pull amplifier stages: run the two plate leads in opposite directions through the probe and then adjust for a null on the current meter.

The probes allow current measurements in large-dimension conductors such as pipes, multiconductor cables, lead-sheathed cables, or microwave waveguides. With large aperture probes, difficult-to-measure quantities, such as corrosion current in small structural members and circulating direct current and low-frequency alternating current in ground straps and waveguides, can easily be determined.

## RESISTANCE MEASUREMENTS

Resistance is usually determined through the familiar form of Ohm's law, $E = IR$. In simple vom's this is done by applying a known voltage (E) to the unknown resistance (R) and then measuring the current (I) passing through it. With E and I known, R can be computed. In actual practice, computation is unnecessary since the resistance scale of the meter is precomputed.

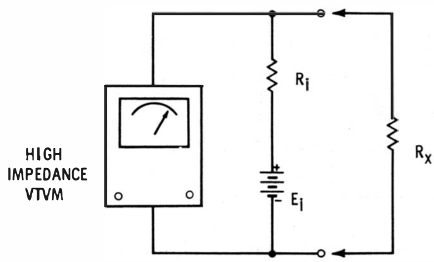



**Fig. 3-4. Circuit for making nominal resistance measurements with an electronic voltmeter.**

Laboratory-type electronic voltmeters use a modified procedure for resistance measurement. As shown in Fig. 3-4, the current in the circuit depends on the series combination of the unknown resistor ($R_x$) and the internal resistor ($R_i$). This means that both the voltage and current in the external circuit will change according to the value of the unknown. The resistance scales of the meter are calibrated for the measurement of this unknown resistance.

If $R_x$ were infinite, the meter would read the full battery voltage ($E_i$). Full-scale deflection, therefore, would correspond to a resistance of infinity. If $R_x$ were zero (short circuit), the meter would read zero. The mid-scale range occurs when $R_x$ equals $R_i$.

The resistance $R_i$, included as part of the ohmmeter circuit, provides a convenient means of changing the range of the instrument. When values of low resistance are being measured, the finite resistance of the ohmmeter leads—included in the total resistance measurement—can contribute considerable error. To meet this problem the circuit can be altered so that the resistance of the current-carrying leads is calibrated as part of $R_i$, while the resistance in the voltmeter leads is insignificant compared with the high input impedance of the metering circuit. Such a circuit is shown in Fig. 3-5.

An external power source is often used in laboratory work where it is necessary to measure very high or very low resistances. For very high resistances a high voltage is applied to the unknown, and the current is measured on a sensitive current meter. High-resistance measurements can be disturbed by the impedance of the measuring voltmeter when this impedance is comparable to the resistance being

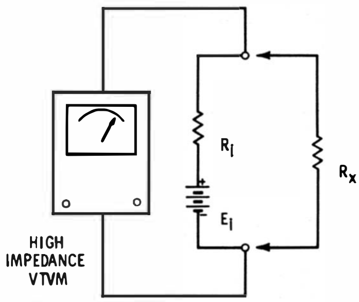



**Fig. 3-5. Circuit for measuring low-value resistance.**

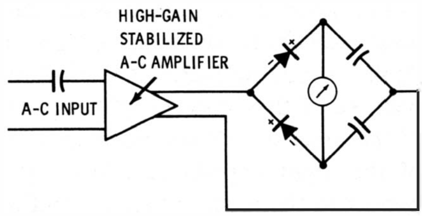
measured. Many laboratory meters account for this by adjusting the value of $R_i$ on the high-resistance ranges to compensate for the voltmeter input impedance. For example, on a 100-megohm scale the value of $R_i$ is actually 200 megohms. The parallel combination of the 200-megohm $R_i$ and the 200-megohm input impedance of the meter gives an effective internal impedance of 100 megohms.

To measure extremely low resistances such as those found in short lengths of large wire or in relay contacts, a constant-current source may be used to supply a fixed amount of current through the unknown resistance. A sensitive voltmeter is then used to measure the voltage drop across the resistance being measured. With this combination, resistance measurement as low as one microhm may be made.

## BASIC A-C VOLTAGE MEASUREMENTS

Electronic instruments for measuring a-c voltages also use an amplifier with the meter movement, but add a rectifier circuit to convert the alternating current to direct current. Most shop-type vom and vtvm units are rms-reading instruments. Mathematically, the root-mean-square (rms) value of any complex quantity is obtained by

Fig. 3-6. Circuit for average-responding voltmeter.



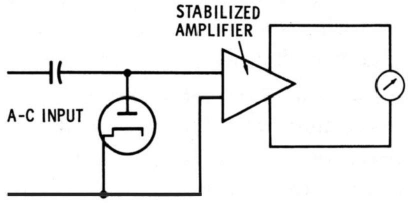

adding the squares of each component and then taking the square root of this sum. Laboratory-type meters, however, fall into three broad categories: average-responding, peak-responding, and rms-responding.

The circuit principle of the average-responding meter is shown in Fig. 3-6. Here, the a-c signal is amplified in a gain-stabilized a-c amplifier and then is rectified by the diodes. The resulting current pulses drive the meter. The meter deflection is proportional to the average value of the waveform being measured.

The peak-responding voltmeter shown in Fig. 3-7 places the rectifier in the input circuit where it charges the small input capacitor to the peak value of the input signal. This voltage is passed to a d-c amplifier which drives the meter. Meter deflection is proportional to the peak amplitude of the input waveform.

Both of these meters (average-responding and peak-responding) have scales calibrated such that the rms value of a sine-wave input voltage is indicated, since the meters are used primarily for sine-wave measurements. Therefore, the average-responding type reads 1.11 times higher than the average voltage, while the peak-responding type indicates 0.707 of the peak value. Consequently, both meters may be in error if the measured signal is not a pure sine wave. The amplitude and phase of the harmonics present affect the peak and average values of the waveform, upsetting the rms calibration. The average-reading voltmeter is not affected by distortion as much as the peak-reading type. However, if highly complex waveforms are measured, then a true rms-responding voltmeter is recommended.

Fig. 3-7. Circuit for peak-responding voltmeter.



For general laboratory work the average-responding voltmeter is used extensively, since it is less affected by distortion of the waveform. Peak-responding meters are used for higher-frequency measurements because of their lower input capacitance. The capacitance to ground of the input circuit and probe of a voltmeter must be included as part of the input impedance. This capacitance acts as a high-frequency bypass to the input resistance and limits the frequency range of most a-c voltmeters.

Since the diode rectifier of peak-responding voltmeters is placed in the probe tip preceding the amplifier, shortening the signal path, the a-c capacitance is low. Input capacitances of one to three picofarads are characteristics of those instruments. The extension of this technique into the low-voltage (millivolt) range is impractical because of the nonlinear response of diodes at low signal levels. A variation of the rectifying technique is required to eliminate the diode nonlinearity. Usually, this is accomplished by using two diodes.

## TRUE RMS-RESPONDING VOLTMETERS

Complex waveforms are measured most accurately by an rms-responding voltmeter. In laboratory instruments this is performed by sensing the heating power of the waveform which is proportional to $(E_{rms})^2$. The indicating circuitry responds to the square root of the heating power. Heating power is measured by feeding an amplified version of an input waveform to the heater of a thermocouple, the voltage output of which is proportional to the heating power of the waveform.

One of the major problems in this technique is the nonlinear behavior of the thermocouple, as well as slow response and possible burnout of the thermocouple. These factors complicate calibration of the indicating meter. This difficulty can be overcome by the use of two thermocouples mounted in the same thermal environment. Nonlinear effects in the measuring thermocouple are cancelled by similar nonlinear operations of the second thermocouple.

As shown in Fig. 3-8, the amplified input signal is applied to the measuring thermocouple, and a d-c feedback voltage is fed to the balancing thermocouple. The d-c voltage is derived from the voltage output difference between the thermocouples. The circuitry may be looked upon as a feedback control system which matches the heating power of the d-c feedback voltage to the input waveform heating power. Meter deflection is proportional to the d-c feedback voltage

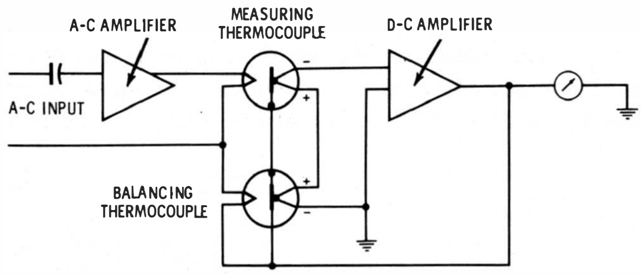


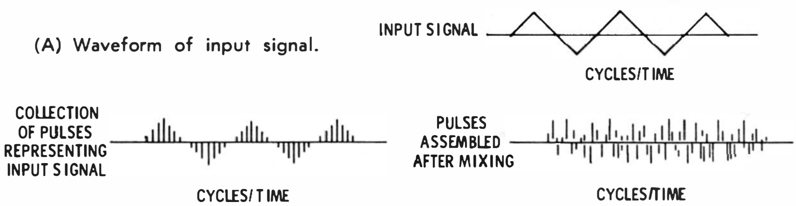
**Fig. 3-8. Circuit for true rms-responding voltmeter.**

which, in turn, is proportional to the rms of the input signal. Therefore, the meter indication is linear.

## THE SAMPLING VOLTMETER

The sampling voltmeter is used to measure complex or nonsinusoidal signals. There are several types of sampling voltmeters; one of the most effective uses the incoherent sampling technique. This is particularly useful where the waveforms have large crests such as sawtooth or triangular waves.

The incoherent sampling technique is illustrated in Fig. 3-9. The technique is best explained by representing each sample with a pulse



(A) Waveform of input signal.

(B) Waveform of a collection of signals.

(C) Waveform of signals after mixing.

**Fig. 3-9. Incoherent sampling technique.**

whose height is proportional to the amplitude of the input signal at the instant the sample is taken. The average, rms, or peak value of the collection of pulses in Fig. 3-9B differs from the input signal (Fig. 3-9A) only by a scale factor. Suppose that these pulses are collected and scrambled. The order in which the pulses appear after being scrambled together results in a waveform similar to that of Fig. 3-9C. Since the pulses are the same, the average, rms, and peak values of this rearranged waveform are identical to the average, rms, and peak values of the waveform in Fig. 3-9B.

Fig. 3-10 is a block diagram of such a voltmeter. Samples taken in the probe are pulsed by the incoherent pulse generator in the sampler. These pulsed samples are fed through the attenuators and amplifiers to the "boxcar" (zero-order hold) circuit. The hold circuit

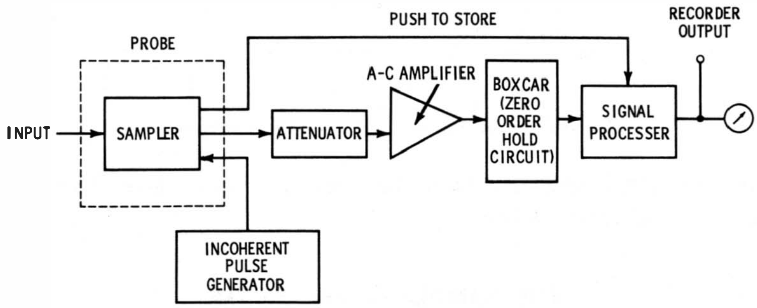



**Fig. 3-10. Block diagram of a Hewlett-Packard Model 3406A sampling voltmeter.**

stores each sample until the next sample is taken. The output of the boxcar circuit is available directly. This output may be used to obtain true rms measurements when used with an rms voltmeter or for peak measurements when used with a peak-reading voltmeter. The output of the boxcar circuit is also fed to a special signal processor which contains noise-suppression circuits for low signal ranges. Low-strength signals mixed with noise of equal or near-equal strength will produce inaccurate readings.

# 4

# Typical Digital Meter Circuits

This chapter is devoted to typical digital meter circuits (the integrating digital meter is discussed further in a later chapter). Several instruments have been selected for discussion. These represent a cross section of digtal meter circuits in current use.

## LOGIC CIRCUITS AND BINARY COUNTING MACHINES

To fully understand the operation of digital meters it is necessary to understand logic circuits and the various binary counting systems. Thus, before going into the details of digital meter circuits, these subjects will be summarized in the following paragraphs.

### Binary Counting Systems

The binary counting system is used in digital computers, pulse-code modulation telemetry, and in various other electronic equipment. In the binary system each number is made up using only zeros and ones, rather than zero through nine as in the familiar decimal systems. Consequently, instead of requiring ten different values to represent one digit the various forms of digital equipment using the binary method need only two values for each digit. In digital equipment the values are easily indicated by the presence or absence of a signal, or by positive and negative signals, or perhaps by two different levels.

One of the easiest ways to understand the binary counting system is to compare it with the more familiar decimal system. In the decimal system the value of each digit is based on ten and the powers of ten. When you read these (decimal) numbers, you automatically add all of the ones, tens, hundreds, thousands, etc. Take the number 3733, for example. You could read this as "three seven three three," but more than likely you would read it as "three-thousand, seven-hundred and thirty-three." The decimal number 3733 means:

$$\begin{array}{l} 3 \times 1000 = 3000 \\ 7 \times \ \ 100 = 700 \\ 3 \times \ \ \ \ 10 = 30 \\ 3 \times \ \ \ \ \ \ 1 = \underline{3} \\ \qquad\qquad\ \ 3733 \end{array} \qquad \text{or} \qquad \begin{array}{ccccccccc} 10^3 & & 10^2 & & 10^1 & & 10^0 & & \\ 1000 & & 100 & & 10 & & 1 & & \\ \underline{\times 3} & & \underline{\times 7} & & \underline{\times 3} & & \underline{\times 3} & & \\ 3000 & + & 700 & + & 30 & + & 3 & = & 3733 \end{array}$$

In other words, you can say that the extreme right-hand digit is multiplied by one, the digit second from the right is multiplied by 10, the digit third from the right is multiplied by 100, and the digit fourth from the right is multiplied by 1000.

In the binary counting system the value of each digit is based on two and the powers of two. This can be displayed as follows:

| $2^8$ | $2^7$ | $2^6$ | $2^5$ | $2^4$ | $2^3$ | $2^2$ | $2^1$ | $2^0$ |
|---|---|---|---|---|---|---|---|---|
| 256 | 128 | 64 | 32 | 16 | 8 | 4 | 2 | 1 |

In the binary system, if the digit is a zero, its value is zero. If the digit is one, its value is determined by its position from the right. For example, to represent the number 77 in binary numbers, the following combination of zeros and ones must be used:

| 256 | 128 | 64 | 32 | 16 | 8 | 4 | 2 | 1 |
|---|---|---|---|---|---|---|---|---|
| | | 1 | 0 | 0 | 1 | 1 | 0 | 1 |
| | | 64 + | 0 + | 0 + | 8 + | 4 + | 0 + | 1 = 77 |

This means that binary number 1001101 equals the decimal number 77. Most digital voltmeters derive their binary-coded decimal (bcd) electrical output from a decade binary counter. Each decade consists of a counter containing flip-flop circuits connected in series. A typical system will have four flip-flops which count up to nine using either a 8421 binary code or a 2*421 (2 star 421) binary code. Tables 4-1 and 4-2 illustrate the difference in the two codes.

**Table 4-1. 8421 Binary Code**

| Decimal | 8 | 4 | 2 | 1 | 8421 BCD Counting Sequence |
|---|---|---|---|---|---|
| 0 | 0 | 0 | 0 | 0 | All zeros equal 0 |
| 1 | 0 | 0 | 0 | 1 | 1 equals 1 |
| 2 | 0 | 0 | 1 | 0 | 2 equals 2 |
| 3 | 0 | 0 | 1 | 1 | 2 plus 1 equals 3 |
| 4 | 0 | 1 | 0 | 0 | 4 equals 4 |
| 5 | 0 | 1 | 0 | 1 | 4 plus 1 equals 5 |
| 6 | 0 | 1 | 1 | 0 | 4 plus 2 equals 6 |
| 7 | 0 | 1 | 1 | 1 | 4 plus 2 plus 1 equals 7 |
| 8 | 1 | 0 | 0 | 0 | 8 equals 8 |
| 9 | 1 | 0 | 0 | 1 | 8 plus 1 equals 9 |

The 2*421 binary code derives its name from the numerical weighting assigned to each of the four "bit" positions in the code. The 2* (two star) bit is called the *real* two and has a numerical weighting of two. The 2 bit also has a weighting of two, but is used only as an added value. The 2 bit is never used by itself, but only in combination with other bits in the code. The 2* bit is included in all combinations from two through nine.

**Table 4-2. 2*421 Binary Code**

| Decimal | 2* | 4 | 2 | 1 | 2*421 BCD Counting Sequence |
|---|---|---|---|---|---|
| 0 | 0 | 0 | 0 | 0 | All zeros equal 0 |
| 1 | 0 | 0 | 0 | 1 | 1 equals 1 |
| 2 | 1 | 0 | 0 | 0 | Real 2 equals 2 |
| 3 | 1 | 0 | 0 | 1 | Real 2 plus 1 equals 3 |
| 4 | 1 | 0 | 1 | 0 | Real 2 plus 2 equals 4 |
| 5 | 1 | 0 | 1 | 1 | Real 2 plus 2 plus 1 equals 5 |
| 6 | 1 | 1 | 0 | 0 | Real 2 plus 4 equals 6 |
| 7 | 1 | 1 | 0 | 1 | Real 2 plus 4 plus 1 equals 7 |
| 8 | 1 | 1 | 1 | 0 | Real 2 plus 4 plus 2 equals 8 |
| 9 | 1 | 1 | 1 | 1 | Real 2 plus 4 plus 2 plus 1 equals 9 |

### Logic and Logic Symbols

Fig. 4-1 is a schematic of an AND gate. If $E_1$ and $E_2$ are positive, $E_0$ (output) will also be positive. If $E_1$ or $E_2$ is negative, $E_0$ will be negative. In a typical system, $E_1$ and $E_2$ will be a nominal +12 volts or −12 volts, and E will be 16 volts. Electron flow for such an AND gate is shown in Fig. 4-2, while the corresponding logic symbol is shown in Fig. 4-3.

A schematic of an OR gate is shown in Fig. 4-4. This gate is similar to an AND gate, except the diodes are reversed and E is a negative voltage. If either $E_1$ or $E_2$ is positive, output $E_0$ will also be positive.

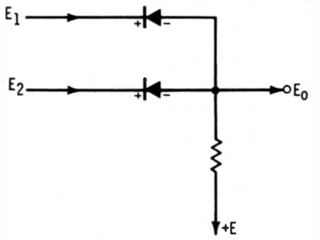



**Fig. 4-1. Network for the positive AND gate.**

If both $E_1$ and $E_2$ are negative, $E_0$ will also be negative. Electron flow for this OR gate is shown in Fig. 4-5. The symbol for a positive-true OR gate with three inputs is shown in Fig. 4-6. NOTE: A negative-true

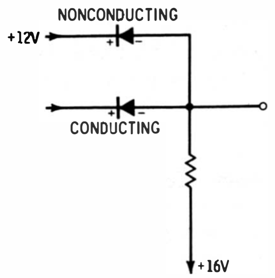



**Fig. 4-2. Diagram of electron flow in an AND-gate network.**

OR gate is shown in Fig. 4-7, and a negative-true AND gate is shown in Fig. 4-8.

In the dynamic AND gate shown in Fig. 4-9, voltage E can be either positive or negative and will inhibit or enable the gate, depending on the polarity of the applied voltage. If E is positive and a

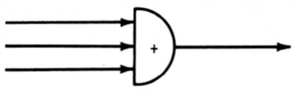



**Fig. 4-3. Symbol for an AND gate with three inputs.**

positive pulse appears at $E_1$, the pulse will forward-bias the diode, and the output $E_0$ will go positive. However, if E is negative, $E_0$ will not go positive even though a positive pulse is at $E_1$. The symbol for a dynamic AND gate is shown in Fig. 4-10. The dot is used to designate the enable leg (E in Fig. 4-9).

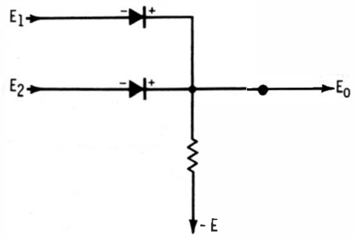



**Fig. 4-4. Network for positive OR gate.**

Fig. 4-11 shows the symbol for an inverter. Inverters are used to convert either a negative-going pulse to a positive pulse or a positive-going pulse to a negative pulse.

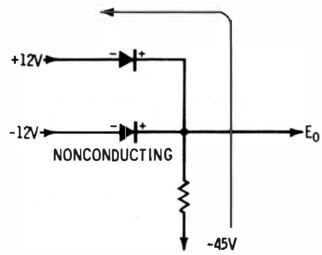



**Fig. 4-5. Diagram of electron flow in an OR-gate network.**

Fig. 4-12 shows a circuit combining an OR gate with an inverter. A positive pulse (12 volts) applied to one or more of the diodes in the OR gate will cause the output of the gate and $E_1$ to switch to the same level. $E_2$ will then become +15 volts. This is sufficient to stop the

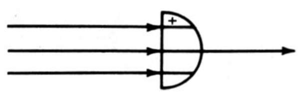
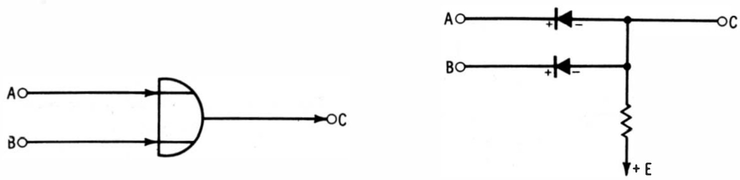


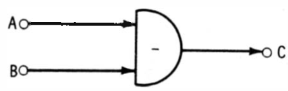

**Fig. 4-6. Symbol for an OR gate with three inputs.**

transistor from conducting and cause its output $E_0$ to switch from plus 12 volts (the emitter potential) to minus 16 volts (the collector supply potential). Thus, for a positive 12-volt pulse input, the output will be a negative 16-volt pulse.

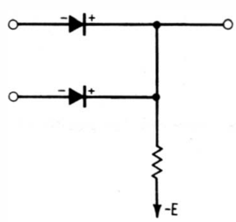



(A) Symbol. (B) Schematic.

**Fig. 4-7. Negative-true logic OR gate.**




(A) Symbol.




(B) Schematic.

**Fig. 4-8. Negative-true logic AND gate.**

Dynamic AND gates are often used in conjunction with flip-flops (bistable multivibrators) in digital circuits. Such an arrangement is shown in Fig. 4-13. When d-c power is first applied to the flip-flop,

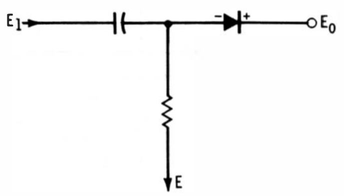
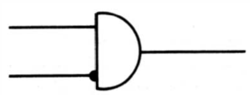



**Fig. 4-9. Diagram for dynamic AND-gate network.**

it immediately assumes a random state where one transistor or the other conducts and the second one is cut off. Which transistor will conduct is determined by inherent differences in components. If tran-



DOT INDICATES THE ENABLE LEG

**Fig. 4-10. Symbol for a dynamic AND gate.**

sistor $Q_2$ is conducting and its collector is positive, the flip-flop is defined as having a "0." bit in it. Until there is a change in the flip-flop state, the circuit "stores" the 0 bit. When transistor $Q_1$ is con-

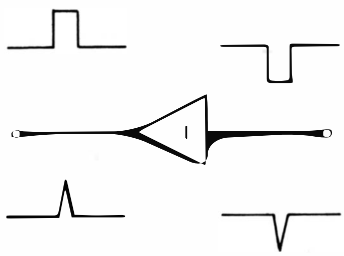



**Fig. 4-11. Symbol and waveshapes for an inverter.**

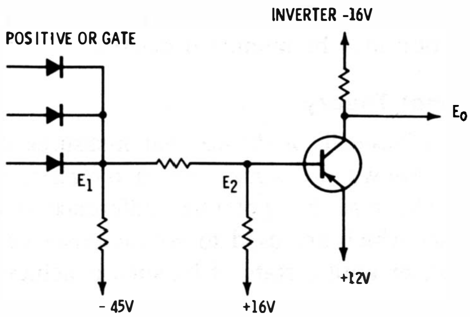



**Fig. 4-12. Network for combination of OR gate and inverter.**

ducting and transistor $Q^2$ is cut off, the flip-flop is defined as having a "1" bit in it.

## MODEL 880 DIGITAL MULTIMETER

The first instrument selected for discussion is manufactured by Electro Instruments, Inc., San Diego, California; it is designated as their Model 880. The instrument is designed to measure d-c voltages and ratios. Display is by means of Burroughs Nixie tubes mounted in a three-position, tiltable readout and consists of five digits, decimal point, and appropriate symbols for all measurements. A photograph of the unit is shown in Fig. 4-14. Electrical outputs are also provided for remote operation of the instrument and for operation of an ex-

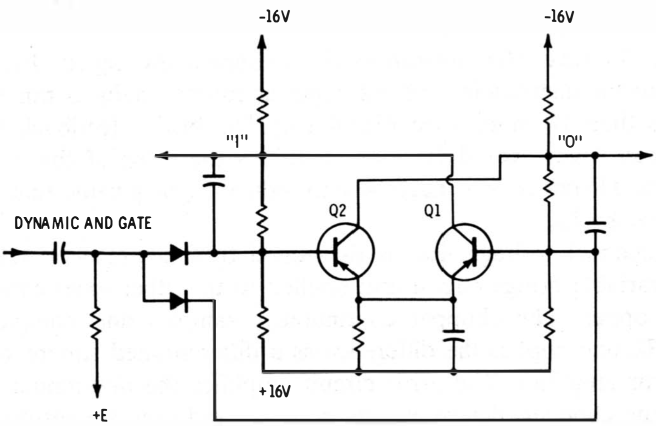
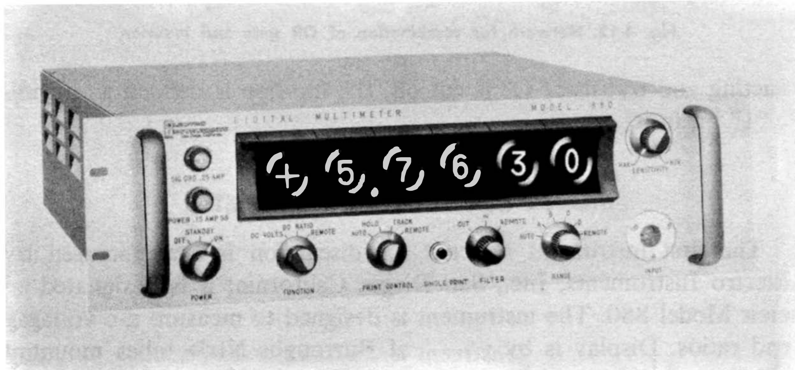



**Fig. 4-13. Operational diagram for a flip-flop with a dynamic AND-gate input.**

ternal printer or recorder. Operation is completely automatic. However, range selection may be manual if desired.

### Basic Operational Theory

The meter is basically a device that measures d-c voltages by comparing the unknown d-c signal with a reference voltage from a digital register. The resulting potential difference is converted into information pulses which are used to continuously vary the contents of the digital register until a state of balance is achieved.



Courtesy Electro Instruments, Inc.

**Fig. 4-14. Electro Instruments Model 880 digital multimeter.**

The overall operational theory for the Model 880 can be divided into three parts: d-c voltage measurement, d-c ratio measurement, and print control operation.

*D-C Voltage Measurement*—The unknown d-c signal, $E_x$, is applied to the instrument, and the input attenuator reduces this voltage to less than 10 volts. (See Fig. 4-15.) The bridge feedback voltage ($E_{fb}$) can never exceed 10 volts, as this is the value of the reference voltage. Therefore, it is necessary to reduce $E_x$ to a value that can be equalled by $E_{fb}$.

A feedback voltage, $E_{fb}$, is developed from a precision reference by a variable bridge circuit and applied to the other fixed contact on the chopper. The chopper continuously samples and compares $E_x$ with $E_{fb}$ and applies the difference as a differentiated square wave to an error amplifier. The error circuit amplifies the differential output from the chopper, determines its polarity, and uses the resultant signals to generate further pulses. The polarity of the square wave, with respect to the position of the moving contact, determines whether these

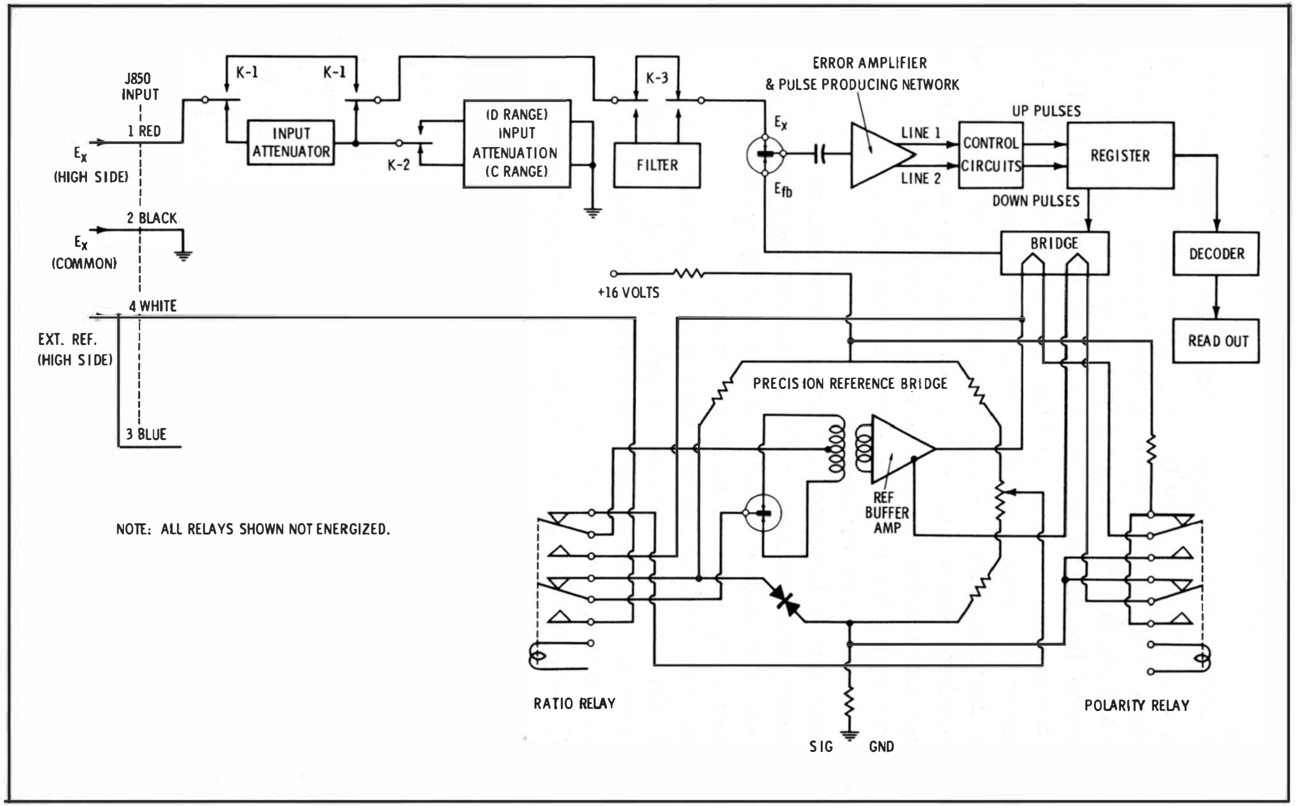



Fig. 4-15. Block diagram of the Electro Instruments Model 880.

pulses will be directed on line 1 or line 2 to subsequent control circuits.

The control circuits consist of front panel switches and the following circuit boards: polarity, range gates, range flip-flops, and d-c ratio logic. The basic functions of these circuits are to convert the pulses arriving on line 1 or line 2 into up or down pulses and to provide control for ranging and polarity. Up pulses cause a digital register to increase both its count and the value of $E_{fb}$; down pulses cause the register to decrease both its count and the value of $E_{fb}$.

The digital register is a counter made up of five decades (units, tens, hundreds, thousands, and ten thousands) and is used to control a bridge and decoder circuit. The output from the bridge is $E_{fb}$, the reference signal that is applied during the measuring cycle to one of the fixed contacts on the chopper.

Output signals from the register are converted by the decoders from binary-coded decimal (bcd) into decimal form for readout presentation. The readout visually displays the digital equivalent of the unknown voltage after the meter has balanced.

A precision reference supply, working in conjunction with a reference buffer amplifier, provides a highly regulated $\pm 10$ volts d-c to the bridge. The reference signal can be positive or negative, depending on the state of a polarity relay.

*D-C Ratio Measurement*—For d-c ratio measurements (see Fig. 4-15), the d-c voltage signals are applied to the instrument using three terminals on the input connector. Pin 4 is connected to the high side of the external 10-volt reference source, pin 1 to the high side of the unknown signal $E_x$, and pin 2 to the common of both the external reference and $E_x$.

With the Function switch set to the Ratio position, the ratio relay is energized and the polarity relay de-energized. The ratio relay disconnects the internal reference source and allows the external reference to be used as the voltage supply for the reference bridge. Instrument operation is the same as during d-c voltage measurements, but the meter now registers the ratio of the unknown d-c signal, $E_x$, to the external reference voltage.

*Print-Control Operation*—The meter is equipped with electrical output connectors and can be used to operate high input-impedance digital recorders, requiring ground on their selected lines. (See Fig. 4-16.)

Print-control circuitry in the meter will provide a closed-loop timing system which operates independently of the recorder. On each measurement, after balance is achieved, a null-detector circuit sets a print control flip-flop. The output of this flip-flop is used to perform the following functions:

1. Provide a print command that can be used by a recorder.
2. Place a hold on the error amplifier so the meter will retain its last reading.
3. Trigger a scanner advance mono (monostable multivibrator) which provides a suitable delay time for a print cycle to take place.
4. Reset a noisy channel mono.

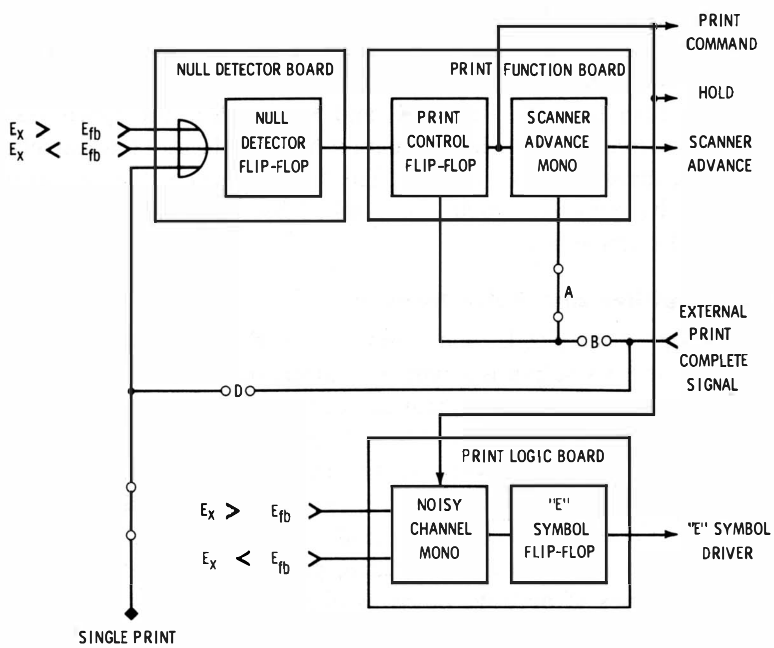



**Fig. 4-16. Simplified block diagram of print control.**

As the scanner advance mono returns to its rest state, it resets the print-control flip-flop so that the instrument will be able to perform its next measurement.

If, however, an external-print complete signal from a recorder or other external device is used to release the meter, the scanner advance mono will serve only as a scanner advance signal generator. For reliable operation, the external-print complete signal should be a 15-volt positive pulse with a maximum rise time of three microseconds and a minimum duration of five microseconds.

### Circuit Analysis

The following paragraphs provide a detailed analysis of the Model 880 circuits. Because of the complex nature of the equipment, the circuits will be described in simplified form. Simplified schematics will be used in place of complete schematic diagrams wherever possible.

### Signal Input Attenuation and Filtering Circuits

These circuits shown in Fig. 4-15 are straightforward and do not require detailed analysis. The high side of the unknown voltage is routed to $K_1$ and/or $K_2$ of the d-c voltage input circuits. There is no attenuation of the input signal $E_x$ in the B range. In the C and D ranges, input attenuation circuits reduce the unknown d-c input signal to less than 10 volts. Relays $K_1$ and $K_2$ switch in the appropriate attenuation circuits. A 66-db attenuation filter is placed between relay $K_1$ and the error-amplifier circuits. The filter is switched in or out of the circuit by relay $K_3$ which is controlled by the positioning of the front panel Filter switch.

### Error Amplifier and Pulse Network

The chopper samples $E_x$ and compares it with the feedback voltage (see Fig. 4-15). Its output is a symmetrical square wave with an amplitude representing the proportional difference between $E_x$ and $E_{fb}$ and having a maximum amplitude 0.8 volts (waveform B in Fig. 4-17). This signal is applied to a differentiating circuit. If $E_x$ is more negative than $E_{fb}$, a series of positive pulses is produced by the differentiator (waveform C in Fig. 4-17). A smaller negative pulse (which is clamped out later) follows each positive pulse. If $E_x$ is more positive than $E_{fb}$, a series of negative pulses is obtained in the same manner. The pulses from the differentiator are amplified by cascaded error and threshold amplifier circuits. The amplifier signals are then applied to a clamper for removal of the unwanted spikes produced by the differentiator circuits (waveform D of Fig. 3-17).

The clamping circuit (Fig. 4-18) is controlled by the same signal that drives the chopper (waveform A of Fig. 4-17). The square-wave output signal from the chopper lags the chopper drive signal by approximately 90°. This phase relationship enables the clamper to remove the pulse produced by the leading edge of the square wave and pass the pulse produced by the trailing edge (waveform E of Fig.

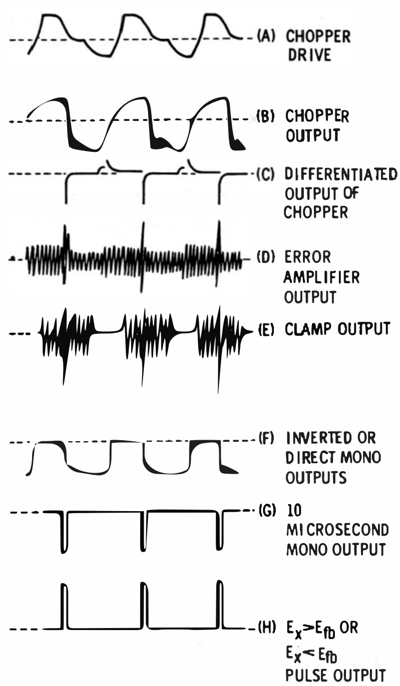



**Fig. 4-17. Output waveforms of the pulse-producing network.**

4-17). The pulse remaining at the output of the clamp circuit will be positive if $E_x$ is positive with respect to $E_{fb}$, and it will be negative if $E_x$ is negative with respect to $E_{fb}$. The clamp circuit output is used to drive a phase-splitter.

The function of the phase-splitter (Fig. 4-19) is to obtain two output signals of opposite phases from a single input. The two outputs of

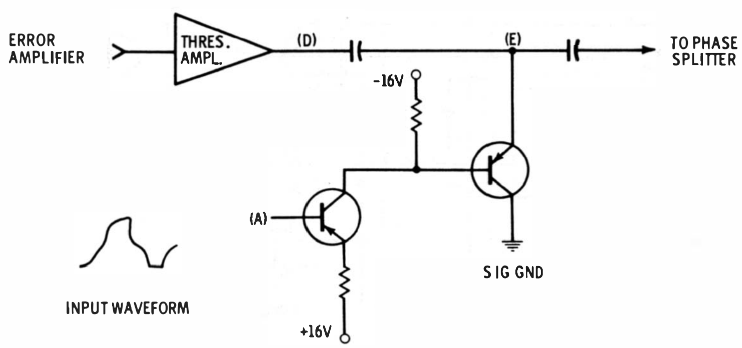



**Fig. 4-18. Diagram of the operation of the clamp circuit.**

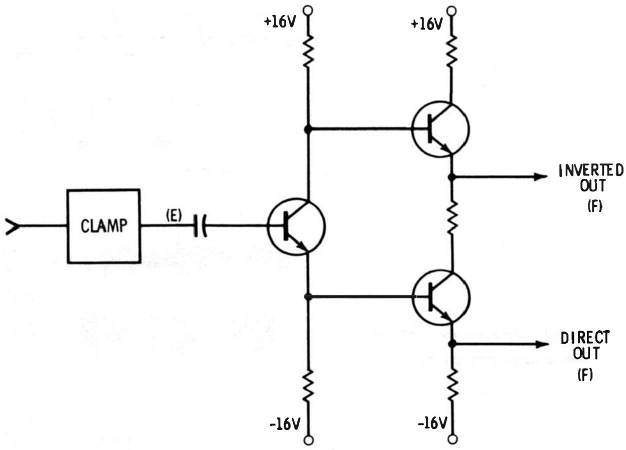



**Fig. 4-19. Diagram of the operation of a phase-splitter circuit.**

the phase splitter, one inverted and one direct, are coupled to the inverted and direct monos by dynamic AND gates 1 and 2. Both the inverted and direct monos are monostable multivibrators which require a positive pulse to trigger them.

When triggered, either the direct mono or the inverted mono then triggers a 10-microsecond mono which produces negative output pulses of 10-microsecond duration (Fig. 4-20). The output of the three mono circuits and the output from a polarity mono (to be discussed later) are connected to the input terminals of negative OR gates in such a manner that OR gate 5 presents a negative 10-microsecond pulse at its output, when triggered, and OR gate 4 presents a 10-microsecond negative pulse when the direct mono is triggered.

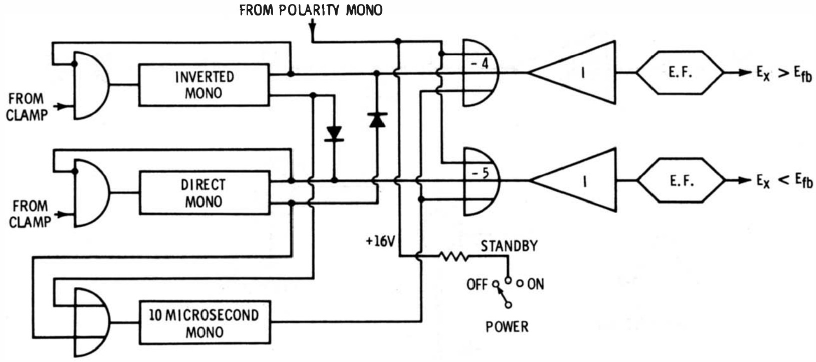



**Fig. 4-20. Diagram of the operation of the mono circuits.**

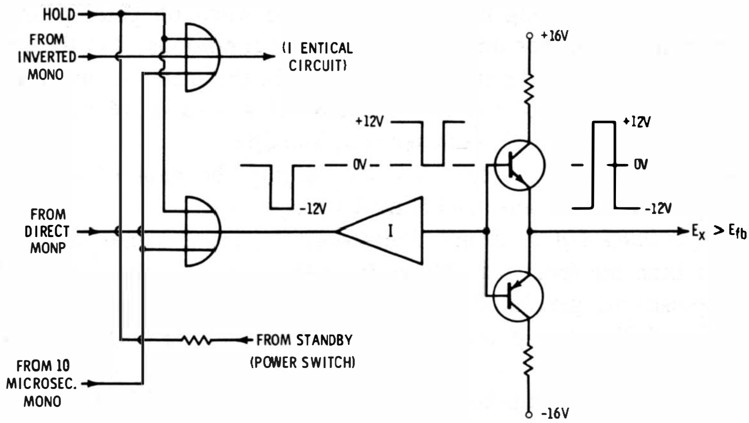



**Fig. 4-21. Diagram of the operation of the emitter-follower circuits.**

Negative output pulses from OR gate 4 or OR gate 5 are inverted in polarity and amplified by complemented emitter-follower circuits (Fig. 4-21). The amplified pulses designated as $E_x > E_{fb}$ from OR gate 4 are

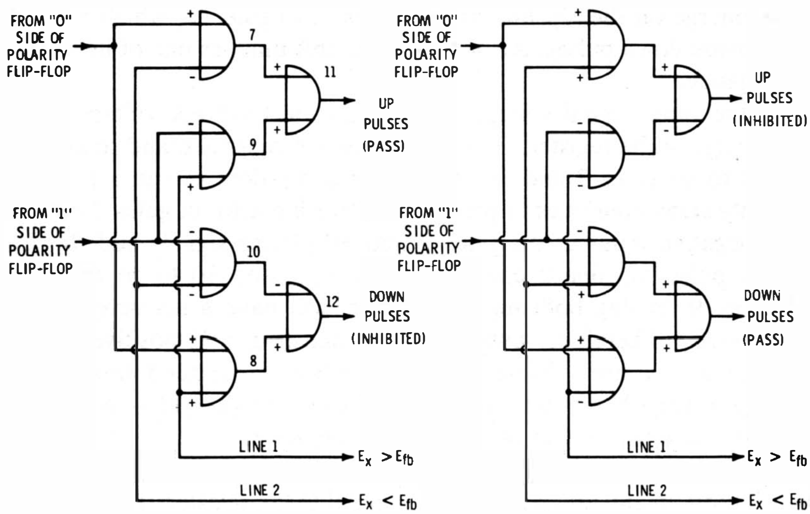



**Fig. 4-22. Diagram of up-pulse or down-pulse determination.**

then routed to OR gate 6 (part of the print-command circuit null-detector) and to OR gates 8 and 9 (Fig. 4-22). The amplified pulses designated as $E_x < E_{fb}$ from OR gate 5 are also routed to OR gates 7 and 10.

The pulse-producing network is disabled when OR gates 4 and 5 are inhibited by either one of the following conditions: (1) the positive pulse output of the polarity mono when the meter is undergoing a range change or polarity reversal and (2) a positive 16 volts from the Power switch when the meter is in Standby.

Meter sensitivity is controlled by varying the gain of the error amplifier with the front-panel Sensitivity control.

*Determining Up or Down Pulses*—If the input signal voltage is greater than the feedback voltage from the bridge of the register, $E_x > E_{fb}$ pulses are generated and routed by line 1 to OR gates 8 and 9. (See Fig. 4-22.) The polarity flip-flop (which will be discussed later) has a zero in it when in a positive polarity state. (The polarity flip-flop uses the basic flip-flop circuit previously discussed, Fig. 4-13). With the polarity flip-flop in the zero condition, positive input is provided to OR gates 7 and 8, and negative input to OR gates 9 and 10. These signals and the $E_x > E_{fb}$ pulses on line 1 cause OR gates 7, 8, and 9 to go positive. Having both inputs negative, OR gate 10 will have a negative output. Therefore, AND gate 11 is satisfied and positive pulses appear at its output. These are amplified by an emitter follower and emerge on the Up line as up pulses. AND gate 12, which is used to provide down pulses, is not satisfied at this time, as one of its inputs is negative.

If the input signal voltage is less than the feedback voltage from the bridge of the register, $E_x < E_{fb}$ pulses are generated and routed by line 2 to OR gates 7 and 10. The polarity flip-flop, still in a positive polarity state, continues to provide positive input to OR gates 7 and 8, and negative input to OR gates 9 and 10. These signals and the $E_x < E_{fb}$ pulses on line 2 cause OR gates 7, 8, and 10 to go positive. OR gate 9, having both inputs negative, will have a negative output (Fig. 4-22). Therefore, AND gate 12 is satisfied, and positive pulses appear at its output. These are amplified by an emitter follower and emerge on the Down line as down pulses. AND gate 11 is not satisfied at this time, as one of its inputs is negative.

*Effect of Up and Down Pulses on the Register*—As shown in Fig. 4-15, both up and down pulses are applied to the register. This register is essentially a counter circuit that counts the pulses (in binary form) and holds the count for decoding (to decimal form) and readout (on the Nixie tubes). The register is made up of five decades (units, tens, hundreds, thousands, and ten thousands). In the standard Model 880, each decade contains flip-flop circuits that will count from zero to nine in the 2*421 binary code previously described. (The Model 880 can also be obtained with the binary count made in the 8421 form.) In either event the decade flip-flops are similar to that shown in Fig. 4-13.

The up and down pulses are applied to the decades through the dynamic AND gate so that each successive pulse changes the count by 1. For example, if all of the counter flip-flops are at the 0 state, the first up pulse will switch the units decade flip-flop to 1. The flip-flop will remain in that condition until the next pulse. This action is continued until all of the flip-flops are set to the appropriate count (in binary form).

In addition to counting (and holding) the pulses, the decade flip-flops also control the nulling bridge, which is actually part of the decade and register circuitry. The meter seeks its null by switching resistors in or out of a bridge circuit which uses +10 volts and ground for reference points. The bridge resistors are connected by transistor conduction to either a precision 10-volt reference supply or to ground, depending on the state of the corresponding flip-flops in the decade register.

The simplified resistor wiring for the bridge is shown in Fig. 4-23. A typical bridge switching-circuit is shown in Fig. 4-24. Four such circuits are used in the units, tens, and hundreds decades (Fig. 4-25).

In each decade the voltages across the resistors connected to the 2*, 4, 2, and 1 flip-flops are summed and applied to the chopper (as $E_{fb}$). Attenuation resistors, located between decades, reduce the effective voltage provided by a given decade to a value consistent with its significance in the counting sequence.

In operation (Figs. 4-23, 4-24, and 4-25), if the flip-flop is in its 0 state, the positive output from the 0 side forward-biases $Q_1$ to stop it from conducting. The output from the 1 side of the flip-flop being negative, back-biases a diode to allow $Q_2$ to be driven into conduction. This effectively connects one end of the bridge resistor at the emitter of $Q_2$ to ground. If the flip-flop had been in a 1 state, the polarities of its output would be reversed. $Q_2$ would then be cut off, and $Q_1$ would conduct to connect the bridge resistor to the 10-volt precision reference supply.

As the flip-flops are moved to the appropriate count, the corresponding bridge resistors are switched in or out so that the bridge output equals that of the input voltage being measured. When this occurs, the chopper output is zero, and there are no further up or down pulses. All of the flip-flops and corresponding bridge resistors will remain in this condition unless the input voltage is changed.

Therefore, the flip-flops have counted the pulses until each flip-flop in the corresponding decade is in a state (0 or 1) that represents the value of the input voltage (in binary form). The outputs of each flip-flop in each decade must then be transformed from a binary code format to a decimal form suitable for presentation by the readout. This is accomplished by the decoder and readout circuits.

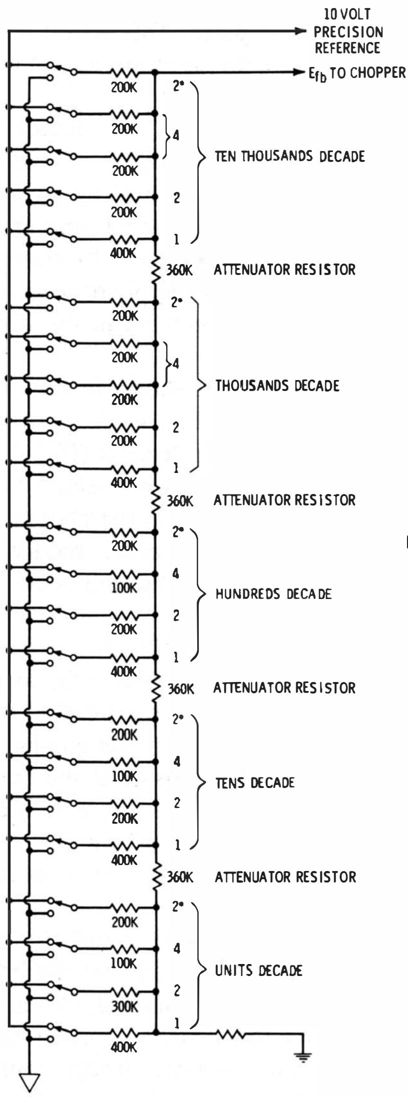



**Fig. 4-23. Simplified diagram of resistor bridge network.**

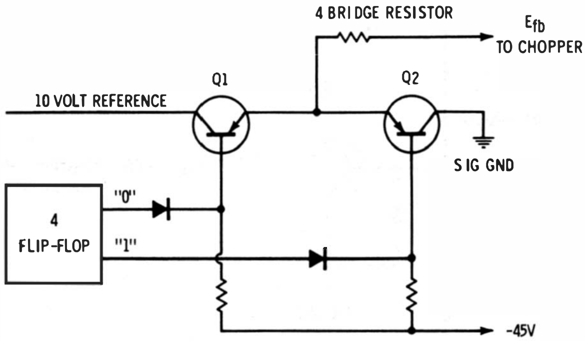



Fig. 4-24. Diagram of typical bridge switch.

## Decoder and Readout Operation

The decoder circuits are primarily special combinations of resistor AND gates. A typical decoder is shown in Fig. 4-26. In operation the transistor conducts when its base is driven positive with respect to its emitter. The design of the circuit is such that the base will never go positive unless positive voltage is applied simultaneously to all four inputs at R1, R2, R3, and R4. Since these resistors are connected to the output terminals of four flip-flops in the decades of the register, the diode will conduct for only one of the nine possible states of the register.

Fig. 4-27 shows how the decoder is connected to the outputs of a typical decade. As can be seen, the collector of the decoder tran-

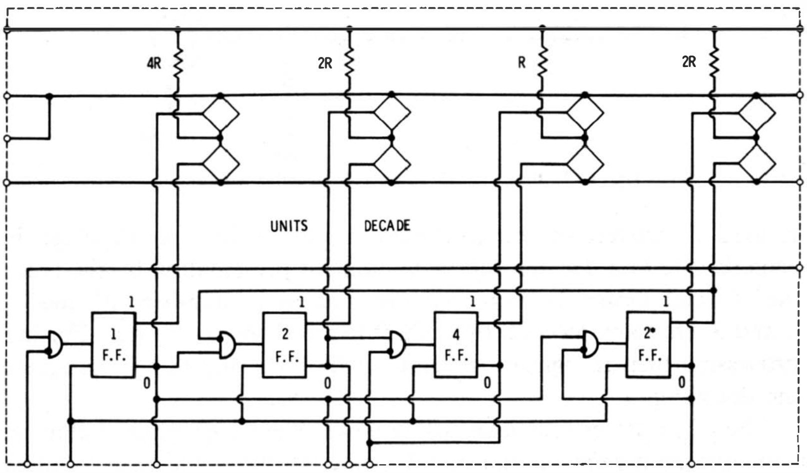



Fig. 4-25. Diagram of precision resistor switching of units, tens, and hundreds decades.

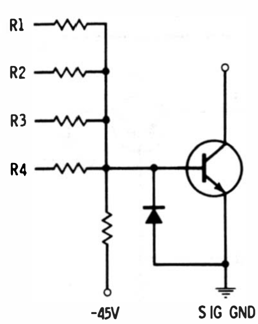



**Fig. 4-26. Diagram of number 3 decoder.**

sistor is connected to the digit-3 Nixie indicator. In this example the decade is at count 3 and is providing positive outputs for a binary 2. With the transistor driven into conduction, the full +200-volt supply is applied to the Nixie, energizing the number 3. The circuit shown in Fig. 4-27 is used only in the ten-thousands decades. All other decades use the circuit shown in Fig. 4-28. Operation of this circuit (known as a *biquinary* decoder) is as follows. As before, the biquinary decoder

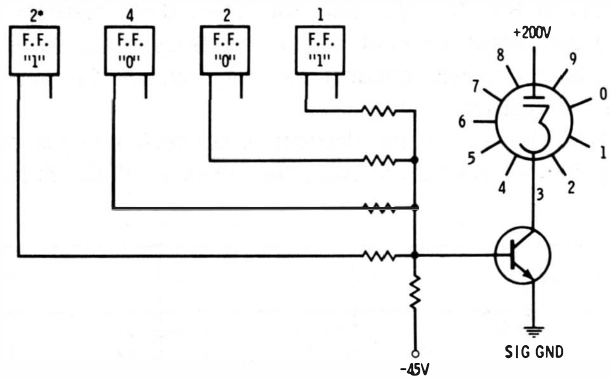



**Fig. 4-27. Diagram of number 3 decoder operation.**

is used to convert the binary-coded output of the four flip-flops in each decade to a decimal representation for presentation by the readout. Circuit design is such that two successive numbers (0 and 1, 2 and 3, etc.) are decoded by each of five resistive AND gates. Further processing then recognizes the state of the 1-bit flip-flop to complete the decoding.

The Nixie driver transistor will conduct when both of the following two conditions exist: 1) the decoder gate output at the base terminal is positive and 2) the emitter line (corresponding to the selected odd or even number) is at a nominal ground potential through the odd and even select transistors ($Q_a$ and $Q_b$). The ten cathodes of the Nixie indicator tube are connected to the collectors of ten corresponding driver transistors for a given decade. Thus, when a particular driver stage is conducting, the selected indicator cathode of the Nixie

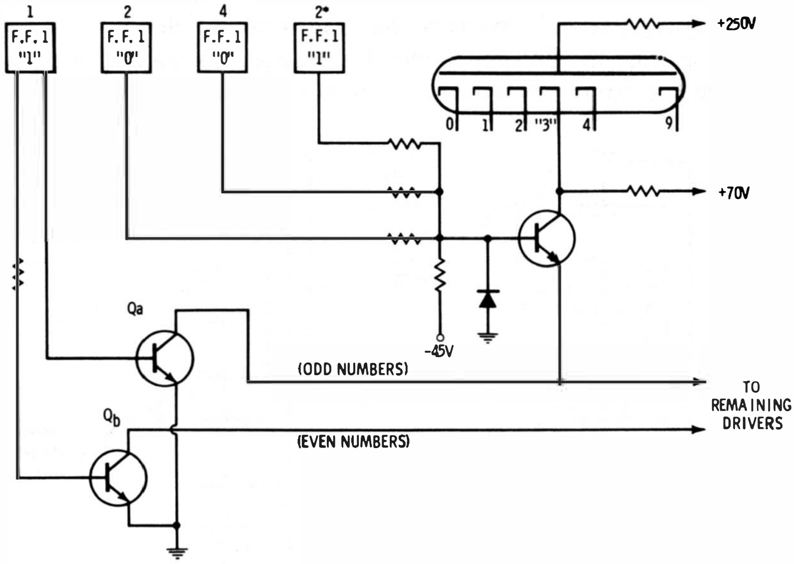



Fig. 4-28. Diagram of typical biquinary decoder stage.

will glow as a result of the application of the full power-supply voltage (+250 volts) less the internal voltage drop of the tube from the anode to the selected cathode.

## Polarity-Changing Operation

During d-c voltage measurements the polarity state existing in the meter is indicated by a plus or a minus symbol in the Nixie readout. The symbol Nixie lamps are energized by Nixie driver circuits similar to those just described. Actual polarity-changing is controlled by the polarity flip-flop discussed earlier.

In operation, if the readout indicates a positive input and the polarity of the input signal is reversed to negative, the meter immediately senses that the input signal ($E_x$) is less than the feedback voltage ($E_{fb}$). Down pulses are generated in the normal manner, and the decades in the readout are driven to +0.0000 with down pulses still being generated.

The down pulses are applied through a series of AND gates to trigger a polarity mono. The output of this polarity mono, which is normally negative when the mono is in its at-rest state, now goes positive and inhibits or places a hold on OR gates 4 and 5 in the pulse-producing network (Fig. 4-20). This prevents pulses that are still being generated by the chopper from passing through the gates and disturbing the digital register while polarity changing is taking place.

A pulse from the polarity mono also changes the state of the polarity flip-flop from 0 to a 1. The 0 output side of this flip-flop then

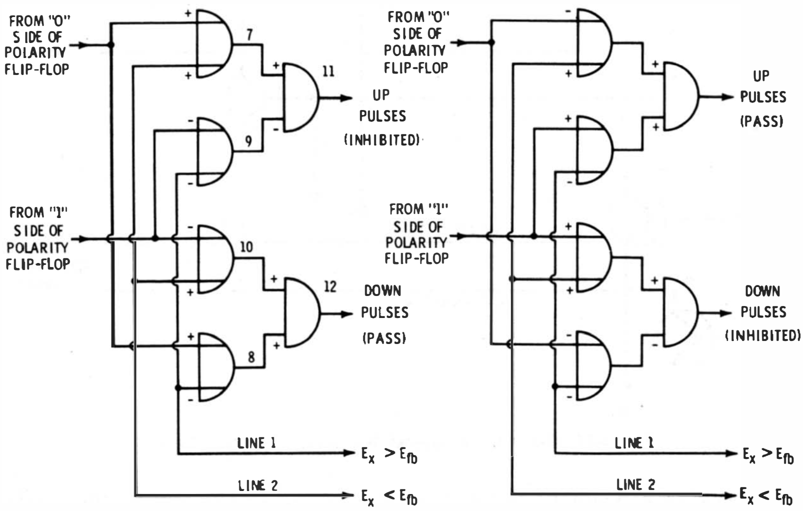



**Fig. 4-29. Diagram of polarity-changing operation.**

goes negative, making the output from its 1 side positive. This reverses the polarity of the drive signals applied to the Nixie drivers and causes the negative-polarity symbol to be displayed. The symbol Nixie circuit is similar to the number Nixie circuits, except that it displays a symbol (+, −, R [ratio], E [voltage]) instead of numbers when the corresponding driver transistor conducts.

Fig. 4-29 shows operation of the gating circuits during a polarity change. When the polarity flip-flop changes state, AND gate 11 is enabled so that $E_x < E_{fb}$ pulses, which are still being generated by the chopper, can pass through the gate to the up-pulse line. At the same time a pulse, generated by the polarity flip-flop as it is changing state, passes through an OR gate that controls a bridge reference supply relay. When energized, this relay reverses the precision reference output of the bridge circuit so that $-10$ volts instead of $+10$ volts is applied to the register bridge.

At this point the readout indicates $-0.0000$. As before, pulses are produced until the meter reaches a null and the readout is equal to the applied voltage. This operation is accomplished in the same manner as previously described for a positive input signal.

## MODEL 111 DIGITAL VOLTMETER

The next instrument selected for discussion is manufactured by Simpson Electric Company, Chicago, Illinois, and is designated as their Model 111. The instrument is designed to measure d-c voltages. Display is by means of seven individually lighted bars and is activated

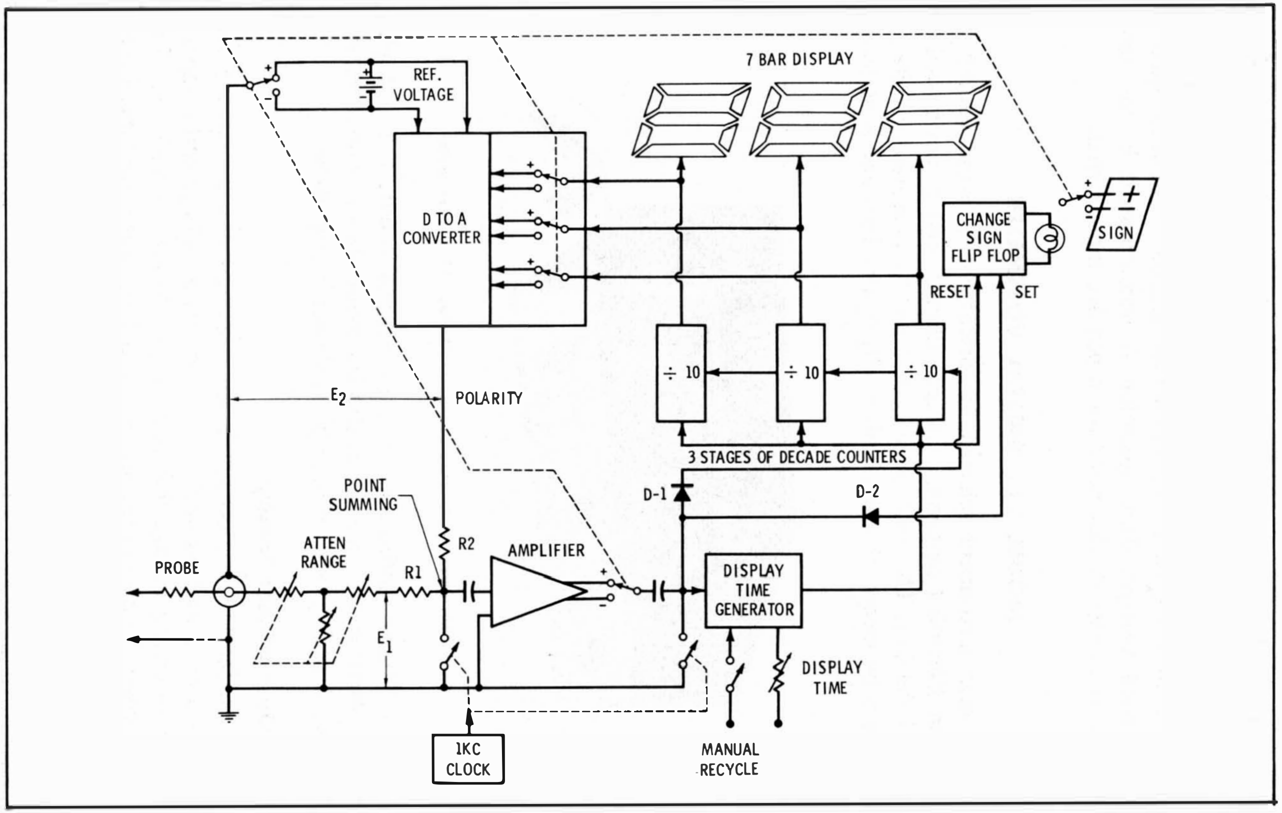

**Fig. 4-30. Simpson Electric Company Model 111 digital meter.**

by 1-2-4-2 binary-coded logic from the internal counters. The display time is variable and can be controlled manually or from a remote location. A photograph of the unit is shown in Fig. 4-30.

### Basic Operational Theory

The Model 111 is basically a null seeking-balancing system that samples a measured portion of the probe input voltage and compares it to an opposing voltage that is being developed internally (see Fig. 4-31). When both voltages are equal and opposite, the unit is said to be at *equilibrium* and a steady-state readout will exist at the counters.

In order to provide numerical readout the development of the internal voltage is divided into 1000 discrete steps (each step corresponding to a number between 000 and 999). As this voltage in-




Courtesy Simpson Electric Company

**Fig. 4-31. Simplified block diagram of Model 111 digital meter.**

creases by its 1/1000 increments until equilibrium of the counters scale is attained.

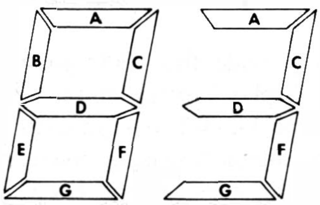
The three stages of decade counters provide the 1000 possible output states of 000 to 999. A seven-bar display is connected to each of the decades to provide the readout of the 1-2-4-2 logic in each of the counters. The counters also drive the digital-to-analog converter which divides a reference voltage into 1000 steps. At the count of 000, the digital-to-analog converter output is zero volts. At the count of 999, the output has increased to approximately five volts. As the counters scale from 000 to 999 (because of added pulses), the output of the digital-to-analog converter (consisting of a staircase waveform) is increased by 1/1000 of the reference voltage (per pulse). Because no more than one second is required to achieve balance, the counter input repetition frequency becomes 1 kc, which in turn becomes the clock rate of the instrument.

The polarity switch reverses the output sense of the digital-to-analog converter. That is, a negative-going staircase is generated when a positive voltage is measured. Conversely, a measured negative voltage results in a positive-going staircase.

When the counters are in the 000 state, the digital-to-analog converter output voltage ($E_2$) is zero. When $E_2$ and a positive input voltage ($E_1$) are applied to the summing network ($R_1$ and $R_2$), a positive voltage representing an out-of-balance signal appears at the summing point.

The clock period of one millisecond is divided into two parts. During the first part (approximately 500 microseconds), grounding switches at the input and output of the amplifier are closed, causing a no-signal condition at the counters and allowing the input and output capacitors to charge to their steady-state values. During the second part of the clock period the switches open, allowing the positive out-of-balance signal to be amplified as a positive-going pulse. The series diode $D_1$ passes this pulse to the counters, causing a count of 001 to appear. This causes the output of the digital-to-analog converter to go negative by 1/1000 of the reference voltage. This completes the action of the first clock-pulse period.

The process is duplicated again in the second clock period. During the first part of the second period the switches are closed to establish 0-volt reference. In the second part of this period, they are opened, and a positive out-of-balance signal condition at the summing point will then initiate another count. The meter will try again for balance.




ALL LAMPS 1764D
(SELECTED)

NUMBER

| LAMP | 1 | 2 | 3 | 4 | 5 | 6 | 7 | 8 | 9 | 0 |
|---|---|---|---|---|---|---|---|---|---|---|
| A | OFF | ON | ON | OFF | ON | ON | ON | ON | ON | ON |
| B | OFF | OFF | OFF | ON | ON | ON | OFF | ON | ON | ON |
| C | ON | ON | ON | ON | OFF | OFF | ON | ON | ON | ON |
| D | OFF | ON | ON | ON | ON | ON | OFF | ON | ON | OFF |
| E | OFF | ON | OFF | OFF | OFF | ON | OFF | ON | OFF | ON |
| F | ON | OFF | ON | ON | ON | ON | ON | ON | ON | ON |
| G | OFF | ON | ON | OFF | ON | ON | OFF | ON | ON | ON |

LAMP DUTY TABLE

**Fig. 4-32. Diagram of seven-bar lamp display arrangement.**

If the voltage to be measured is in range, then at some count between 000 and 999, a null or slightly negative voltage will be obtained at the summing point. At this time, more pulses can pass to the counters through $D_1$, and the meter reaches equilibrium.

At equilibrium, the display time generator starts its functions to hold the final-state displays of the counter for a time, depending on the setting of the display time control and/or manual recycle button. It also (not shown in Fig. 4-31) locks out pulses from the amplifier so that stray noise spikes or removal of the probe will not cause spurious readings. At the end of the display cycle, the counters are reset, and the meter repeats the entire cycle just described.

In the case of a wrong-polarity input signal, a negative-going pulse will appear at the output of the amplifier during the second half-cycle of the first sampling period (counters at 000). $D_1$ prevents them from counting. However, the other diode ($D_2$) passes the negative pulse and triggers the change-sign flip-flop. This lights the Change-Sign lamp. Reversing the polarity switch reverses the sense of the digital-to-analog converter output and the sense of the amplifier output, so that a negative out-of-balance signal at the summing point will initiate a count, and each count will produce a positive step at the digital-to-analog converter output.

Fig. 4-32 shows the seven-bar display arrangement. The seven bars are arranged to form numerals. Any number from zero through nine can be formed with the proper combination of bars. For example, to form the number 3, lamps B and E do not light, but the remaining lamps, A, D, G, C, and F, do light. The lamps are turned on or off by driver transistors which, in turn, are controlled by the decade counter flip-flop circuits. Each of the three numerals is provided with a separate seven-bar display.

# 5

# Typical Differential Meter Circuits

This chapter is devoted to typical differential meter circuits. While the basic operating principles of all differential meters are the same, as discussed in Chapter 2, the circuits used in the various meters differ considerably.

## MODEL 823A

The instrument selected for discussion is manufactured by the John Fluke Mfg. Co., Inc., Seattle, Washington, and is designated as their Model 823A. It is representative of the basic differential-meter principle in that it will measure both alternating current and direct current, as well as high resistances, on a differential basis. A photograph of the unit is shown in Fig. 5-1 and an overall block diagram is shown in Fig. 5-2.

### Basic Theory of Operation

As shown in Fig. 5-2, the circuit consists basically of an a-c to d-c converter, a d-c vacuum-tube voltmeter (vtvm), and a 0- to 500-volt d-c reference. Overall operation of the 823A can be summarized as follows:

To measure the approximate value of a d-c voltage, the unknown voltage is connected directly to the d-c vtvm. The a-c to d-c converter and 0- to 500-volt reference supply are then cut out of the circuit.

Courtesy John Fluke Mfg. Co., Inc.

**Fig. 5-1. John Fluke Model 823A differential meter.**

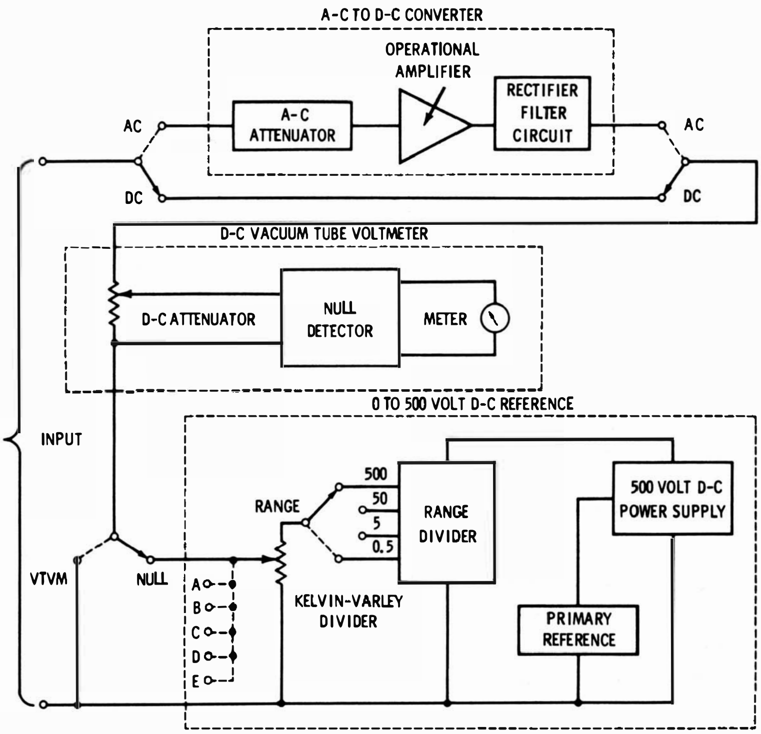



**Fig. 5-2. Block diagram of the Model 823A differential meter.**

To accurately measure a d-c voltage using the differential principle, the unknown voltage is connected across the series combination of the d-c vtvm, and the 0- to 500-volt reference supply. The reference voltage is then adjusted with the five voltage-readout dials until it matches the unknown voltage that is indicated by the null on the d-c vtvm. The unknown voltage is then read from the reference voltage dials.

All a-c measurements are made by first converting the a-c input voltage to a d-c voltage by means of the a-c to d-c converter. The 823A then operates essentially the same as for direct current. The following paragraphs provide a detailed analysis of the Model 823A circuits.

### D-C Vacuum-Tube Voltmeter Circuits

The d-c vtvm is composed of an attenuator, a null detector, and a meter. (See Figs. 5-3, 5-4, and 5-5.) The heart of the d-c vtvm is the null detector in which the d-c signal input is modulated, amplified, rectified, and finally filtered to produce a d-c output. The chopper-amplifier portion of the null detector has a high amount of negative current feedback. This makes the output current approximately equal to the signal voltage divided by the impedance of the feedback network, regardless of the amplifier characteristics. The high negative feedback also makes the amplifier relatively insensitive to the gain changes in individual stages due to aging and replacement. The output current from the null detector is indicated on a meter that has taut-band suspension. This suspension does away with all friction associated wit hmeter pivot stickiness.

*Null-Detector Circuit*—At the input of the null detector (see Fig. 5-3), R201, C201, R202, and C202 form a double section, low-pass filter that reduces any a-c component present on the d-c voltage being measured. The difference between the voltage appearing at the output of the filter and the voltage developed across the feedback network is converted to an alternating voltage by V201 and V202, a series-shunt photochopper. The photochoppers are driven by an astable multivibrator operating at 84 cps to prevent interference from line-frequency pickup. The chopped voltage is amplified by solid-state amplifier Q201 through Q205 before being delivered to demodulator photocell V203. During half the chopper cycle the output of the amplifier is clamped close to the null-detector common potential by V203. During the other half-cycle the output is filtered by R221 and C211 to provide a d-c output.

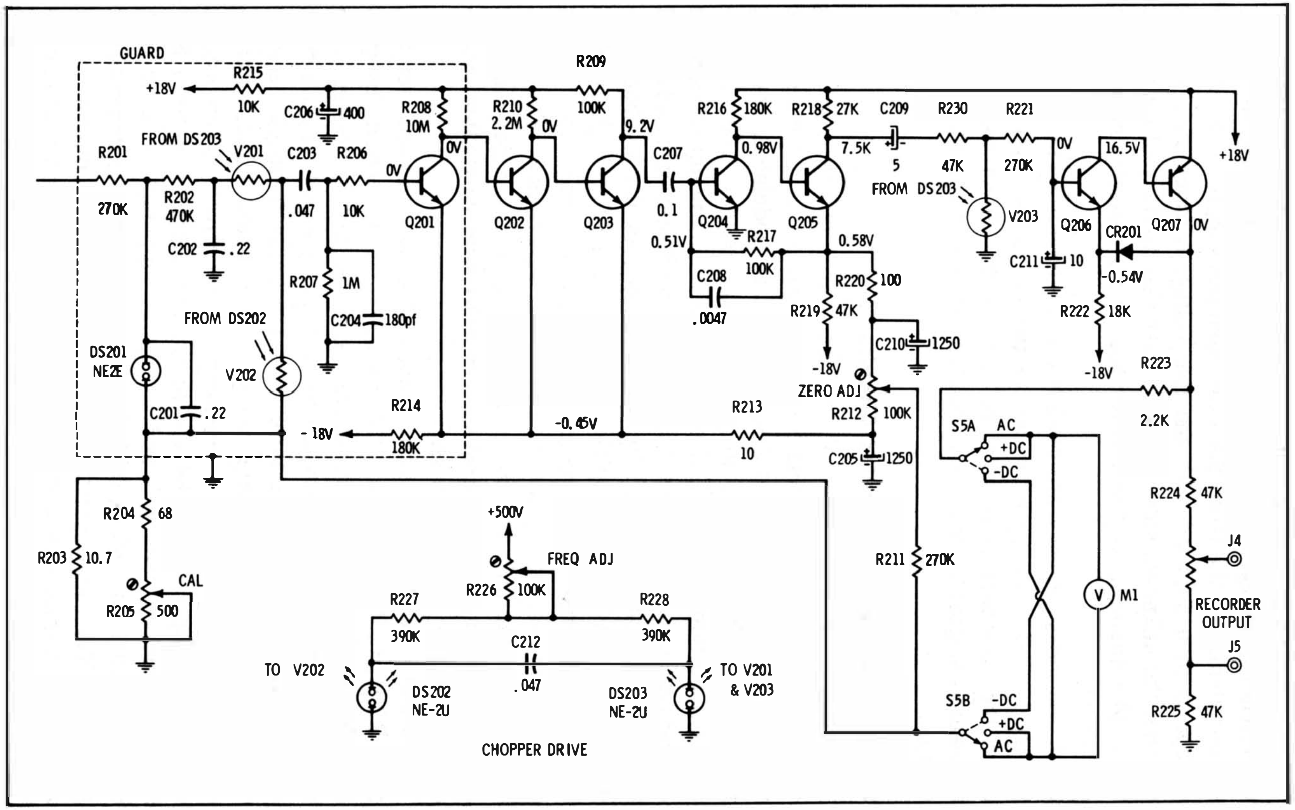



Fig. 5-3. Schematic of a null-detector circuit.

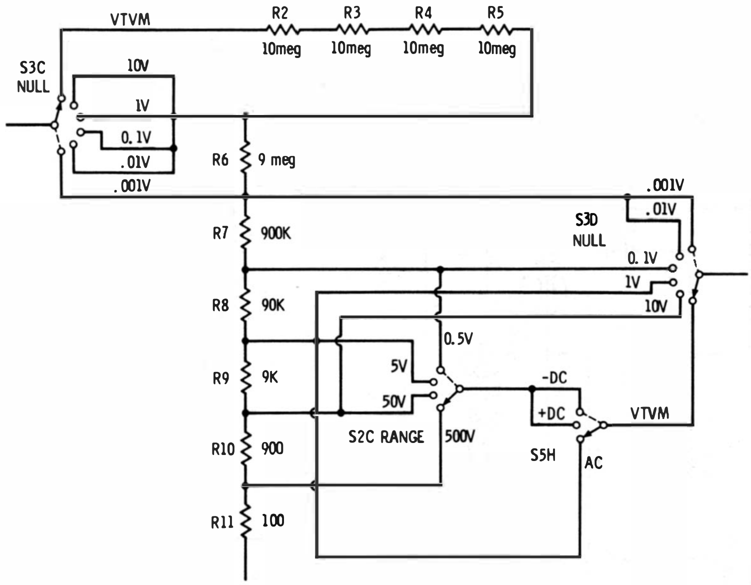


Fig. 5-4. Schematic of an input attenuator circuit.

To prevent loading of the chopper amplifier, especially at high temperatures, the filtered output is fed to unity-gain amplifiers Q206 and Q207 which supply the necessary meter current. The voltage developed across feedback network R203, R204, and R205 is proportional to the meter (output) current. When DS202 illuminates V202, its resistance drops from about 50 megohms to 5 kilohms, and the amplifier input is equal to the feedback voltage. When DS203 illuminates V201, its resistance drops from about 50 megohms to 5 kilohms and the amplifier input is equal to the output voltage from the input filter.

The impedance of the feedback network (R203, R204, and R205) is adjustable between 9.28 and 10.5 ohms. Since the output current is approximately equal to the signal voltage divided by the impedance of the feedback network, a 1-mv signal voltage indicates an output current of 95.1 to 108 $\mu$a. The output current is adjusted to 100 $\mu$a during calibration by means of the feedback network. Thus, current feedback makes the output current essentially proportional to the signal voltage. For full-scale deflection, a 1-mv signal voltage will cause 100 $\mu$a to flow through the meter.

A *recorder output* is picked off divider string R224, R1, and R225. Output level control R1 provides a means of adjusting the output

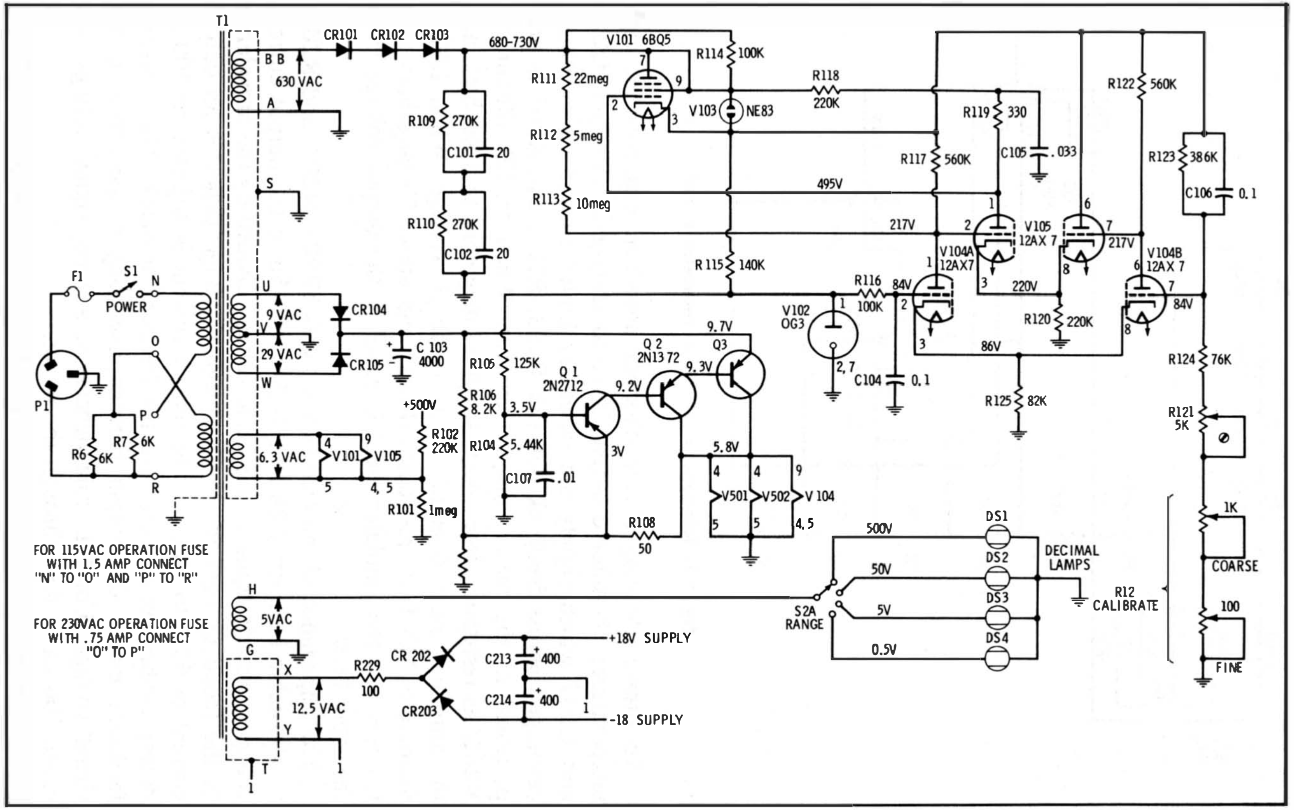



Fig. 5-5. Schematic of a 500-volt power supply.

voltage up to a maximum of at least 18 millivolts at full-scale deflection. The voltage at the output terminals is proportional to the meter reading.

Variable resistor R212 provides a means of adjusting the output current of the amplifier to zero when there is no input signal by utilizing the negative bias voltage on the emitter of Q201 and the positive bias voltage on the emitter of Q205. The gain of the amplifier is adjusted by means of R205 in the feedback circuit.

*Input Attenuator Circuit*—In the d-c vtvm mode, four positions on the vtvm attenuator (Fig. 5-4.) selected by range-switch section S2C provide the necessary reduction of the 500-, 50-, 5-, and 0.5-volt ranges for proper null-detector input. For this mode the resistance of the attenuator, and thus the input resistance of the 823A, is 50 megohms (R2 through R6).

In the d-c differential mode, the voltage difference (unknown voltage minus reference voltage) is reduced by four positions on the vtvm attenuator selected by null switch sections S3C and S3D to give full-scale deflections corresponding to inputs of 10, 1, 0.1, and 0.01 volts. For full-scale deflection corresponding to 0.001 volt the voltage across the attenuator is fed directly to the null detector. Although the resistance of the vtvm attenuator is 10 megohms (R6 through R11) for the 10-, 1-, 0.1-, and 0.01-volt null ranges, and 1 megohm (R7 thru R11) for the 0.001-volt null range, this is not the actual resistance of the 823A.

> NOTE: The input resistance of most differential voltmeters is determined by dividing the unknown terminal voltage by the current drawn from the unknown. The current drawn from the unknown is equal to the difference between the unknown terminal voltage and the internally known voltage divided by the resistance of the input attenuator. Since the reference voltage is equal to the unknown voltage and the source voltage at null, no current is drawn from the unknown; the input resistance is therefore infinite.

In the a-c vtvm mode, null switches S3C and S3D, and a-c/d-c switch sections S5H provide connection to only one position on the vtvm, regardless of where the range switch is set. This is because the output of the a-c to d-c converter is 5 volts d-c for full input on each range. In the a-c differential mode, the voltage difference (converter output voltage minus reference voltage) is reduced by the same positions on the vtvm attenuator as for d-c differential measurements.

Because of this and the fact that the converter puts out 5 volts d-c for full input on each range, the null range used must be multiplied by the a-c null multiplier indicated by the range switch to find the full-scale difference between the unknown voltage and the reference voltage.

The input impedance for the a-c vtvm and a-c differential mode depends upon the input impedance of the a-c to d-c converter and its attenuator. The input impedance is thus dependent on the setting of the range switch and is 1 megohm, 35 pf for the 500-volt a-c range; 1.1 megohm, 35 pf for the 50-volt a-c range; and 1 megohm, 50 pf for the 5- and 0.5-volt a-c ranges.

### 0- to 500-Volt Reference Circuit

When the 823A is used for differential voltage measurements, an internal voltage is nulled or matched against an unknown voltage. An extremely accurate reference voltage is therefore required. This is obtained from the 0- to 500-volt reference. (See Figs. 5-5 and 5-6.) The 0- to 500-volt reference is composed of a well-regulated 500-volt supply, a range divider, and a Kelvin-Varley five-decade attenuator. The output of the 500-volt power supply is applied directly to the Kelvin-Varley attenuator for the 500-volt d-c range. In the 50-, 5-, and 0.5-volt d-c ranges the range divider reduces the voltage to 50, 5, and 0.5 volts before it is applied to the Kelvin-Varley attenuator. For any a-c range, the range divider always reduces the voltage to 5 volts. The Kelvin-Varley attenuator divides its input voltage (500, 50, 5, or 0.5 volts) into 50,000 equal increments, any number of which may be selected by setting the five decades with the voltage readout dials. The output of the Kelvin-Varley attenuator, therefore, provides an extremely accurate reference voltage.

*500-Volt Power Supply*—The 500-volt power supply (Fig. 5-5) uses three diodes (CR101, CR102, and CR103) and a filter network (R109, R110, C101, and C102) to supply unregulated d-c voltage to V101. The voltage is regulated by comparing a sample of the output voltage tapped off a divider string (R123, R124, R121, and R12) with the voltage from reference tube V102 in differential amplifier V104. The control action of the differential amplifiers continuously adjusts the voltage drop across V101 in order to keep the sample voltage equal to the voltage of the reference tube. Any difference in voltage or error is amplified by V104 and V105, and thus changes the voltage drop across V101 to maintain the output at a constant 500 volts.

For proper operation, a highly stable and accurate balance of amplification must be maintained between the two halves of differential amplifier V104. The filament supply for V104 must be regulated to maintain this balance. Regulation is provided by a transistor regulator which supplies constant voltage to the filaments of V104. (The filaments of the a-c to d-c converter tubes V501 and V502 are also supplied by this regulator.) A filtered, full-wave rectifier (CR104, CR105, and C103), which is regulated by a three-transistor network (Q1, Q2, and Q3), supplies the regulated d-c filament voltage. One side of the filaments is connected to the reference-supply common (0 volts), while the other side is maintained at approximately 5.9 volts through the emitter-collector junction of Q3. The output of Q1 drives a Darlington connection of Q2 and Q3. Any variation in the unregulated supply causes a corresponding change in the voltage across the emitter-collector junction of 3 so that the filament voltage of V104, V501, and V502 remains stable.

*Range Divider*—In the 500-volt d-c range (Fig. 5-6), the output of the 500-volt supply is passed directly to the Kelvin-Varley attenuator by range switch section S2F. In the 50-, 5-, and 0.5-volt d-c positions, range resistors (R320 through R330) selected by section S2E, divide the reference voltage to 50, 5, and 0.5 volts before it is switched to the Kelvin-Varley attenuator by section S2F. With the a-c/d-c switch set to a-c, section S5G provides connection to the range resistors that divide the reference to 5 volts. This signal of 5 volts is then passed to the Kelvin-Varley attenuator by section S5F. The voltage applied to the Kelvin-Varley attenuator is always 5 volts for alternating current because the a-c to d-c converter always supplies up to a maximum of 5 volts to the vtvm attenuator.

*Kelvin-Varley Attenuator*—In Fig. 5-6 the five Kelvin-Varley decade resistor strings R401 through R455 and associated voltage dials A through E (S6 through S10) provide a means of making the four precision voltages (500, 50, 5, and 0.5) adjustable. Note that each string, with the exception of the first, parallels two resistors of the preceding string. Between the two wipers of S6 (voltage dial A), then, there is a total resistance of 40K (80K paralleled by 80K). With the range switch in the 500-volt position, a total voltage of 100 volts d-c will appear across these two wipers.

There will be 10, 1, and 0.1 volts d-c across the wipers of S7, S8, and S9, respectively. Voltage dial E (S10) picks increments of 0.01 volt d-c from the last decade. These voltages are reduced by a factor of 10 for each lower voltage range. The Kelvin-Varley resistors are

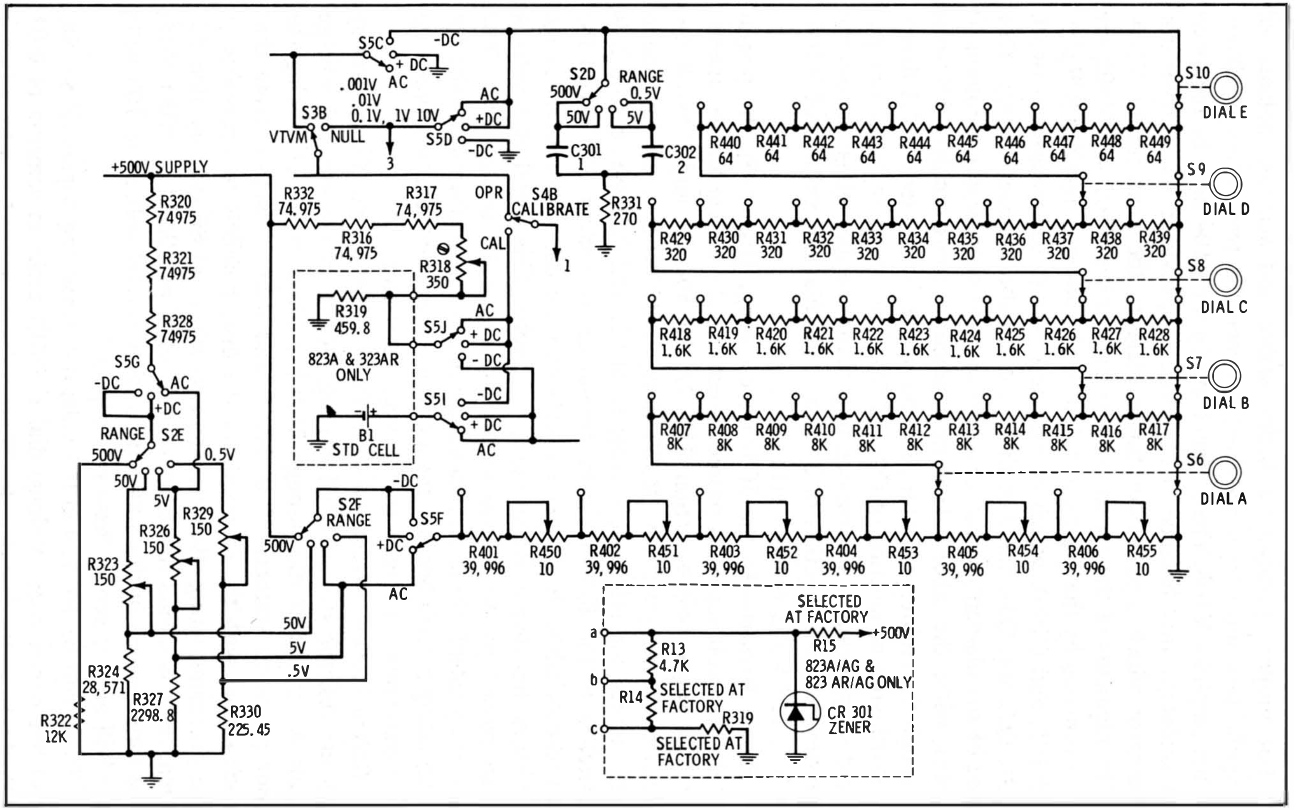



Fig. 5-6. Schematic of a range divider and Kelvin-Varley attenuator.

matched for both temperature coefficient and tolerance, thus providing an overall attenuator accuracy of 0.002 percent absolute. With the null switch in any null range the output of the Kelvin-Varley attenuator is connected in series with the vtvm attenuator, providing the accurate 0- to 500-volt reference required.

*Reference Supply Adjustments*—Variable resistor R121 (Fig. 5-5) is used during calibration to set the 500-volt supply to 500 volts with calibrate control R12 (Fig. 5-5) set at its center of rotation. This allows the reference supply to be adjusted to 500 volts by means of the calibrate control at any time deemed necessary. With the operate-calibrate switch held at calibrate, a fixed percentage of the 500-volt supply is compared with the precise potential of an internal standard cell or to a zener diode (Fig. 5-6) depending on the model. Any difference in potential is fed to the null detector to give an indication on the meter so that the 500-volt supply can be set with the calibrate control. The fixed percentage of the reference supply is set accurately during calibration by means of R318 (Fig. 5-6). The 50-, 5-, and 0.5-volt range resistor networks are set accurately during calibration by means of R323, R326, and R329 (Fig. 5-6).

### A-C to D-C Converter Circuit

The a-c to d-c converter is composed of an attenuator, an operational amplifier, and a rectifier-filter circuit. (See Fig. 5-7.) A diode in the rectifier filter circuit is used to convert the unknown alternating current into pulsating direct-current, which is then filtered to obtain a d-c voltage that is proportional to the average value of the a-c input voltage. The output, however, is calibrated to indicate the rms value of a pure sine wave.

An operational amplifier containing three resistance-capacitance coupled amplifier stages with high negative feedback is used to make the rectification characteristics of the diodes linear and stable. The amplifier achieves a midband loop gain of approximately 70 db with virtually a flat frequency response from 20 cps to 10 kc. The high negative feedback makes the amplifier practically noise-free and relatively insensitive to gain changes in individual tubes due to aging and tube replacement.

At the output to the amplifier, full-wave rectification is used to return negative feedback to the grid of the first amplifier tube. The attenuator is used to reduce the a-c input voltage by a factor of 10 to 100, as required to restrict the operational amplifier input to 5 volts maximum for full-scale inputs of 50 to 500 volts, respectively.

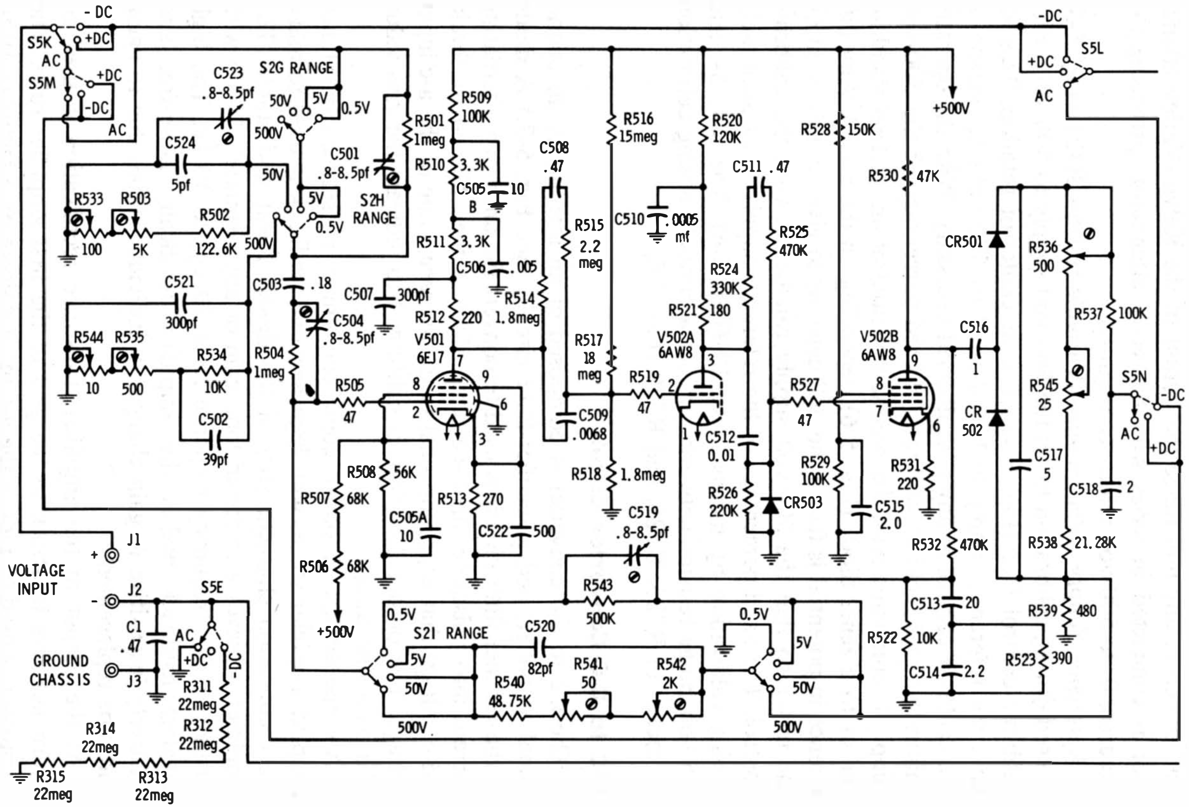



Fig. 5-7. Schematic of alternating-current to direct-current converter circuits.

*Circuit Operation*—All a-c measurements are made by first converting the a-c input voltage to a d-c voltage. The converter provides a d-c output of 5 volts when full-range voltage is applied in each a-c range. In the 5-volt a-c position, range switch sections S2G and S2H connect the input binding posts directly to the converter input. In this case the converter feedback is of such value that the d-c output voltage is equal to the rms value of the converter input a-c voltage. In the 500- and 50-volt a-c positions, an input attenuator reduces the unknown alternating current by a factor of 100 and 10, respectively. The operation of the converter is then the same as for the 5-volt position. In the 0.5-volt a-c position, range switch section S2I and section S2J provide connection to feedback resistors that allow the converter to produce an output that is equal to ten times the a-c input. Thus, an output of 5 volts d-c is provided for all full-scale inputs.

*Null Indications With A-C-Measurements*—When making a-c differential measurements, the null range used (times the applicable a-c null multiplier) must be used to represent full scale on the 832A meter. This is due to the way that the converter is designed. For the 500-volt a-c range the converter produces a d-c output voltage equal to 1/100 of the a-c input voltage. Thus, the a-c null multiplier for the 500-volt range position is X100. For example, when the range switch is set to 500 and the null switch is set to 0.01, full-scale meter deflection represents a one-volt ($100 \times 0.01$) difference between the unknown voltage and the amount set on the voltage-readout dials. By similar reasoning, the multipliers for the 50-, 5-, and 0.5-volt a-c ranges are X10, X1, and X0.1, respectively.

*Converter Circuit Adjustments*—For 0.5-volt converter gain, R536 and R545 at the output of the converter are adjusted. The gain of the amplifier for the 5-volt range is adjusted by means of R542 and R541 in the feedback network. The high-frequency response of the amplifier input is adjusted by means of C504, while C519 adjusts the high-frequency response of the 0.5-volt feedback circuit. The attenuation of the 500-volt attenuator is adjusted with R535 and R544, and the attenuator high-frequency response is adjusted with C501. The attenuation of the 50-volt attenuator is adjusted by R503 and R533; the attenuator high-frequency response is adjusted by C523.

### A-C/D-C Polarity Switch Circuit

The a-c/d-c polarity switch is provided for selecting either the a-c or d-c mode of operation. The connections provided by this switch can best be understood by reference to the diagram of Fig. 5-8.

When the a-c/d-c polarity switch is set to a-c, the a-c to d-c converter is switched into the circuit by sections S5K, S5L, S5M, and S5N. Also sections S5F and S5G are used to switch 5 volts d-c to the Kelvin-Varley divider. When in the a-c vtvm mode, S5H is used to provide proper attenuation of the d-c attenuator.

The a-c/d-c polarity switch may be set to either of two positions, positive or negative, for the d-c mode of operation. As shown in Fig.

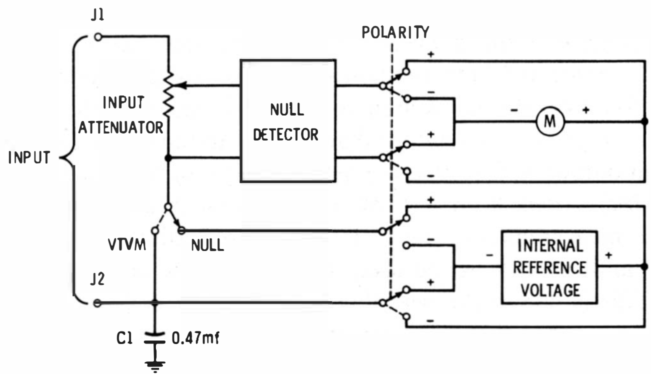



**Fig. 5-8. Function diagram of a polarity switch.**

5-8, the a-c/d-c polarity switch does not reverse the input posts when going from the positive to negative d-c position. Instead, it reverses the meter and the internal reference supply. If the instrument did not contain positive and negative d-c positions, the grounded side of any unknown voltage that is negative with respect to ground would have to be connected to the upper-input binding post. This would ground the upper post and effectively place C1 across the input terminals. Even if C1 were disconnected, there would still be considerable capacitance across the input due to transformer winding capacitance and stray wiring capacitance. With this capacitance across the circuit being measured, several problems would arise. The polarity switch provides equal convenience in measuring positive or negative voltages without the occurance of these problems.

### Differential Voltmeter Applications

A differential voltmeter is well adapted for the measurement of resistance. There are two basic procedures, one of which is well suited for measurement of resistances above 1 megohm, while the other is better suited for precision measurement of lower resistance

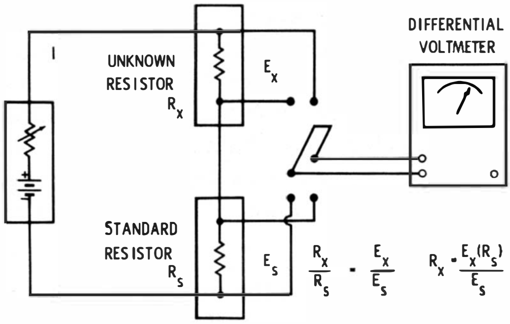



**Fig. 5-9. Method for measuring standard resistors.**

values. The following sections describe how a differential voltmeter can be used to make such resistance measurements.

### *Measurement of Standard Resistors*

It is possible to make accurate resistance measurements of lower value (1 megohm or less) resistors by comparing the voltage ratio between a standard resistor and the unknown resistor to be measured.

If a standard resistor $R_s$ is connected in series with the unknown resistor ($R_x$) as shown in Fig. 5-9, and the current through both resistors is the same, the ratio of the voltage drops, as determined by the differential voltmeter, is the same as the ratio of their resistances.

To obtain accurate and valid results it is essential that the current (I) remains constant throughout the measurement. This is easily verified by repeated measurements of the voltage drop with $R_s$ and $R_x$ alternately connected in the circuit.

The accuracy of resistance measurements using this method depends on the accuracy of the voltmeter and standard resistor.

The accuracy and resolution of the voltmeter used will be a prime factor in determining the measurement accuracy. If the ratio of the resistors is such that the voltage drop across $R_x$ and $R_s$ is measured on the same voltage range, then the basic error of the voltmeter will cancel out. If, for example, the voltmeter had a specified accuracy of better than ±0.1 percent, and the voltmeter was in error by +0.005 percent, then the value of the unknown resistor ($R_x$) would be

$$R_x = \frac{E_x\,(1 + 0.005\%)}{E_s\,(1 + 0.005\%)}\,(R_s) \quad \text{or} \quad R_x = \frac{E_x}{E_x}\,(R_s)$$

indicating that the error of the voltmeter cancels out. The error contributed by the voltmeter will consist only of the error caused by the linearity characteristics of the Kelvin-Varley divider. If the ratio of resistors $R_x$ to $R_s$ is large enough that the voltage measurement requires a voltmeter range change, then it is necessary to consider the basic error contributed by the voltmeter. This illustrates the need for knowing the capabilities and limitations of the voltmeter and compensating for these limitations by knowing the errors associated with your voltmeter. The resolution of the voltmeter must also be known.

NOTE: *Resolution* is defined as the smallest quantity discernable using a given measurement-device. The resolution of any measurement should always be greater than the accuracy expected.

Standard resistors can be obtained which typically provide calibration accuracies of ±0.005 percent, or better, and will remain within 50 parts per million of their certified value over a temperature range of 20° to 30° centigrade. The power coefficient of the resistors used is an important factor when attempting accurate measurements. Tables 5-1 and 5-2 indicate the current and power permissible for standard resistors as determined by the degree of accuracy required. It is suggested that the values of current and power indicated in Tables 5-1 and 5-2 be considered as guidelines in order to minimize the possibility of errors caused by self-heating in the resistors being measured.

The following example is intended to give an idea of the results obtainable by measuring resistors with a precision differential voltmeter.

*Measurement of a 10-ohm standard resistor*

$R_s$ = value of the standard resistor = 10.000 ohms.
$E_s$ = voltage drop across the standard = 1.0000 volt.
$E_x$ = voltage drop across the unknown = 1.0035 volt.

$$R_x = \frac{E_x}{E_s}(R_s) = \frac{1.0035}{1.0000}(10.000) = 10.0035 \text{ ohms.}$$

Accuracy of standard resistor = ±0.005%.
Accuracy of the voltmeter Kelvin-Varley divider = ±0.002%.
Total inaccuracy of the measurement ±0.007%.
The total error of the measurement is shown as the sum of the standard resistor and voltmeter errors.

One of the major advantages of this type of measurement is that the circuit under test need not be disturbed. Therefore, there is no

**Table 5-1. Maximum Current and Power for Certificate Limit of Error**

| Resistance (ohms) | Current (amperes) | Power (Watts) |
|---|---|---|
| 1 | 0.3 | 0.09 |
| 10 | 0.1 | 0.1 |
| 100 | 0.03 | 0.09 |
| 1000 | 0.01 | 0.1 |
| 10,000 | 0.003 | 0.09 |
| 100,000 | 0.001 | 0.1 |

loading effect, and the lead and contact resistances have no effect. This type of measurement is well suited for the measurement of four terminal resistors that are part of complicated networks. However, the measurement circuit should be grounded at only one point to prevent the possibility of ground loops.

**Table 5-2. Maximum Current and Power for ±0.02-Percent Limit of Error**

| Resistance (ohms) | Current (amperes) | Power (watts) |
|---|---|---|
| 1 | 1.0 | 1.0 |
| 10 | 0.3 | 0.9 |
| 100 | 0.1 | 1.0 |
| 1000 | 0.03 | 0.9 |
| 10,000 | 0.01 | 1.0 |
| 100,000 | 0.003 | 0.9 |

### Measurement of High Resistance

Differential voltmeters provide a feature that allows them to be used for the direct measurement of high resistances. The use of a differential voltmeter for this purpose provides a convenient and rapid method for measuring resistances from 1 megohm to 250,000 megohms. The applications for this type of measurement include the measurement of leakage resistance of capacitors, transformers, and insulators.

Fig. 5-10 shows the basic circuit configuration for accomplishing the measurement. This circuit consists of a known power supply voltage that can be selected by the voltage range switch and applied to a precision Kelvin-Varley divider. The setting of the Kelvin-Varley divider determines the exact voltage that is applied across the null detector connected in series with the unknown resistance, $R_x$. The input impedance of the null detector (which is usually on the order of 1 or 10 megohms, depending on the null range setting and the model of the instrument) is used as the standard resistor ($R_x$) in the circuit.

The measurement of resistance with a differential voltmeter is a direct application of Ohm's law: the measurement of the current through the unknown resistance and the voltage drop across it. With an unknown resistor connected to the input terminals of a differential voltmeter, the current through the unknown resistance is equal to the voltage indicated by the null detector divided by the null-detector

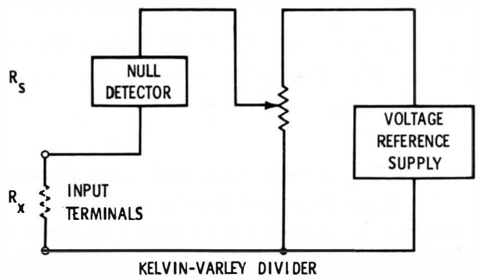



**Fig. 5-10. Method for measuring high resistances.**

input resistance. Likewise, the voltage drop across the unknown resistance is equal to the voltage measured on the Kelvin-Varley divider minus the voltage indicated by the null detector.

Since R=E/I, the voltage drop across the unknown resistance divided by the current through the unknown resistance is equal to the unknown resistance value.

For practical measurements, it is recommended that resistances up to 50,000 megohms be made with the null-detector meter needle at end (full) scale. With the most common differential voltmeter, the current will be known to within approximately 4 percent (3 percent from the null detector and 1 percent from the null-detector input resistance).

However, the accuracy of the measurement depends not only on the accuracy of the current through the circuit, but also on the percentage of the applied voltage that is dropped across the unknown resistance, compared with the voltage dropped across the null detector. For example, if the Kelvin-Varley dials indicated 1.1 volts, and the null detector indicates an end-scale reading of 1.0 volt (assuming a 10-megohm input resistance), then using the following equation we can determine the value of the unknown resistance, $R_x$:

$$R_x = \frac{R_n}{E_m}(E - E_m)$$

where,

$R_x$ = unknown resistance in megohms,
$E$ = voltage indicated on the Kelvin-Varley divider,
$E_m$ = voltage indicated on the null detector,
$R_n$ = resistance in megohms of the null detector.

If we compute the value of $R_x$, we find it to be 1 megohm:

$$R_x = \frac{10}{1}(1.1 - 1) = 10 \times 0.1 = 1 \text{ megohm}$$

But, if we recompute it, assuming that the null detector voltage reading is in error by +3 percent, we find that the value of $R_x$ to be low by 32 percent. Although the accuracy of the null detector is +3 percent, this 3 percent error in 1 volt represents 32 percent error in the 0.1 volt dropped across the unknown resistance. Conversely, we see that as the ratio of voltage drop across the unknown resistance becomes large with respect to the voltage drop across the null detector this error will be practically eliminated.

# 6

# Typical Integrating Digital-Meter Circuits

This chapter is devoted to typical integrating digital-meter circuits. While the basic operating principles of all integrating digital meters are the same, it was shown in Chapter 1 that the circuits used in the different meters vary considerably.

## MODEL 510

The instrument selected for discussion is manufactured by Vidar Corporation, Mountain View, California, and it is designated as their Model 510. It is representative of the basic integrating meter principle in that it combines two instruments in one (a voltage-to-frequency converter and a frequency counter) to provide a direct-reading digital display of an input voltage integrated over the gate (sampling) time selected. A step attenuator provides six calibrated full-scale sensitivity settings from 10 millivolts to 1000 volts. Fixed gate times of 10 msec, 100 msec, and 1 second are available, or any desired gate time may be used from an external source. Counter sensitivity of 0.1 to 100 volts rms up to 300 kc is available with both front and/or rear inputs.

An internal precision ±1-volt calibrating source is controlled from the front panel for both checking and calibrating the instrument.

The binary-coded decimal (bcd) output on the standard 510 is the 1-2-4-8 code (with "0" positive) and is available at a connector located on the rear panel. Bcd outputs are provided for five digits, decimal location in negative exponents of ten, and the measurement units of plus volts, minus volts, or frequency. This connector also supplies a readout command and accepts a readout hold signal. A photograph of the unit is shown in Fig. 6-1.


Courtesy Vidar Corporation

**Fig. 6-1. Vidar Corp. Model 510 integrating digital voltmeter.**

## Basic Theory of Operation

The overall theory of operation is best understood by referring to the block diagram of Fig. 6-2. Surrounding most of the instrument is an outer shield which is formed by the instrument sheet-metal exteriors (Fig. 6-3). Contained within the outer shield and enclosing the voltage-controlled oscillator (vco) of the Model 510 is a guard shield. In operation the guard shield is connected to source ground. The effect of the guard shield is to prevent the flow of common-mode currents in the input circuit. (Common-mode signals are discussed in later sections.) Contained within the guard shield is an inner shield. The inner shield is connected to the common rail of the vco power supply and forms an electrostatic shield for internally generated a-c voltages. Within the inner shield is the vco proper. Multiple-shielded transformers provide for the coupling of a-c power to the vco and coupling of the pulse train output to the counter section of the instrument, while maintaining the triple shielding.

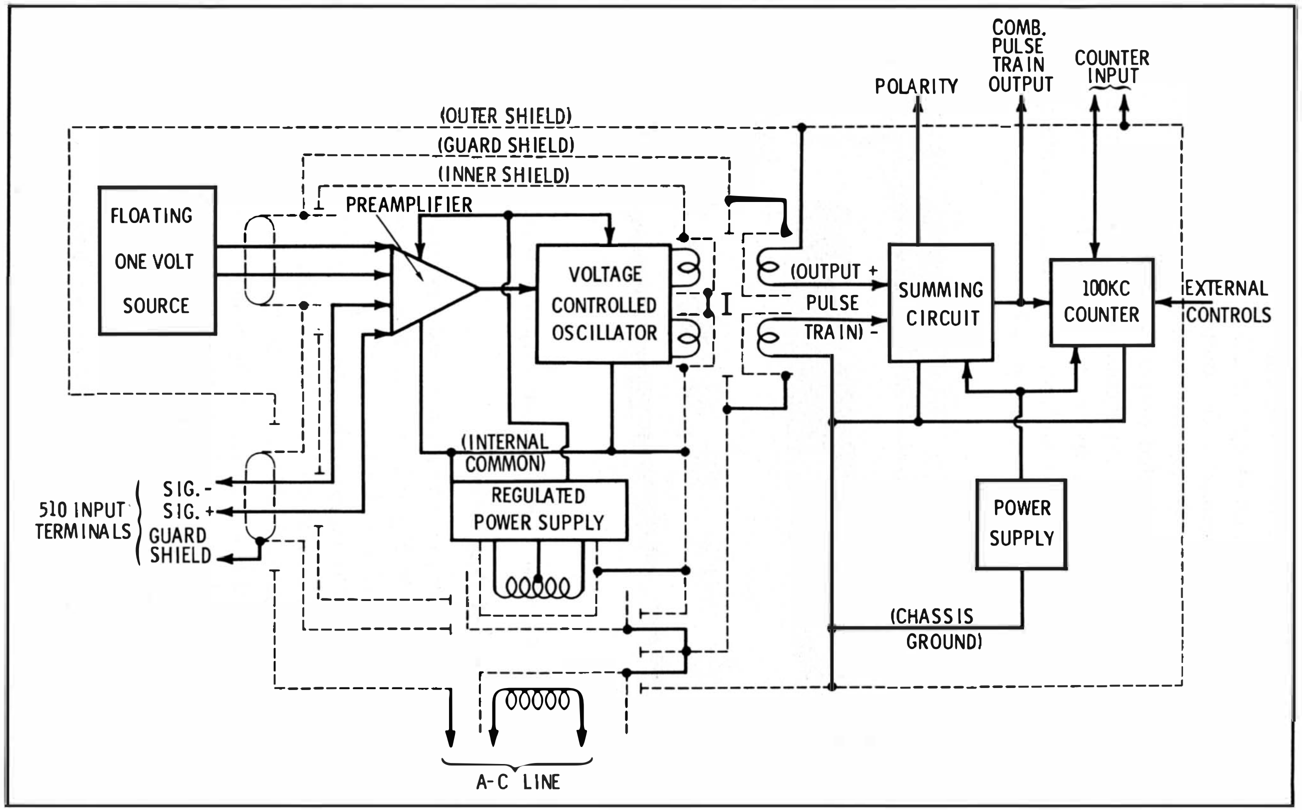



Fig. 6-2. Block diagram of Model 510.

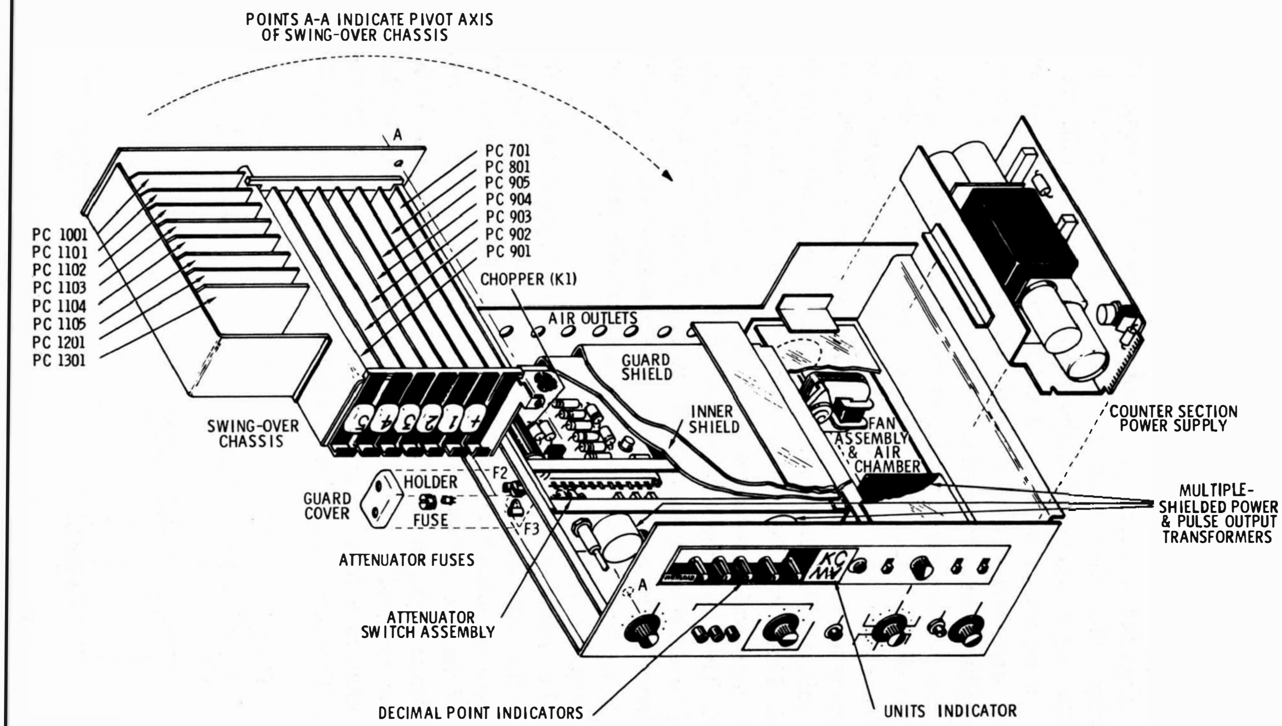


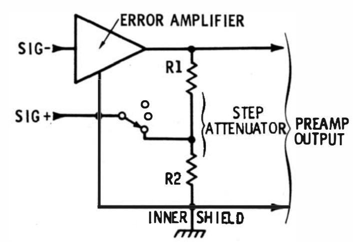

**Fig. 6-3. Exploded view of Model 510.**

*Basic VCO Circuit*—The vco consists of a preamplifier, a voltage-to-frequency converter, and a regulated power supply. The preamplifier is shown in block diagram form in Fig. 6-4. The preamplifier consists of an error amplifier and a precision divider (step attenuator) and operates in either of two modes, potentiometric or operational. The error amplifier is a high-gain, chopper-stabilized, low-pass, inverting amplifier. The input (source) signal is applied between the error amplifier input and an adjustable tap on the precision divider, for the millivolt ranges, or the inner shield for the volt ranges. An amplified output signal—out of phase with the input signal—appears at the preamplifier output terminals.

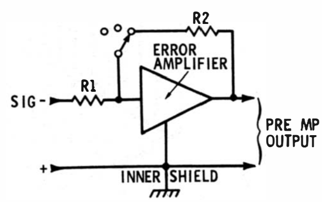
In the *potentiometric* mode, where the millivolt ranges are used, a fraction of the output signal appears at the junction of the precision divider resistors (R1 and R2). This divided output signal is fed back in opposition to the input signal. Since the error amplifier has very high gain, the divided output signal must very nearly equal the input signal. This is true because any difference in magnitude between the two signals will cause a corrective change in the preamplifier output signal. Therefore, the voltage gain of the preamplifier is very nearly equal to the division ratio of the voltage divider (R1-R2), and since the input signal is opposed by an equal-amplitude, opposite-phase signal, the input current required is essentially zero. Thus the input resistance is very high.

In the *operational* mode, where the volt ranges are used, the function is the same as just described except that the opposing signal is a current to balance the one provided by the input signal. The high gain of the error amplifier in this case affords essentially a zero input impedance at the mode. The preamplifier therefore has an input resistance essentially equal to R1 and a gain of the ratio of R2 to R1.

Contained within the error amplifier is a nonlinear feedback network. This internal feedback becomes effective only when the input




(A) Potentiometric.




(B) Operational.

**Fig. 6-4. Block diagram of Model 510.**

signal exceeds a predetermined limit. When this limit is exceeded, both the gain and input resistance of the error amplifier are greatly reduced, thus preventing further increases in either amplifier input or output voltages. The effect of the nonlinear feedback is to prevent overloading of the error amplifier by abnormal input signals. Within a few milliseconds following removal of the abnormal input the internal feedback drops out and normal preamplifier operation is restored.

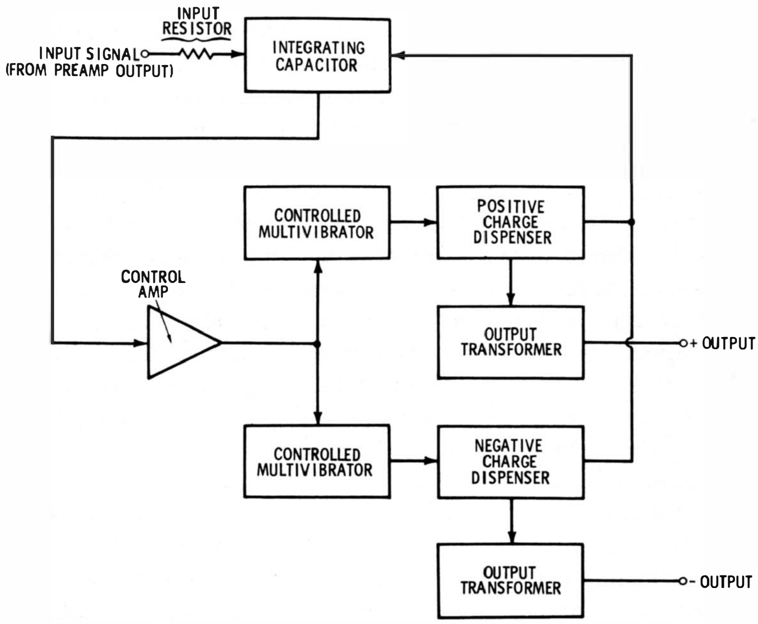



**Fig. 6-5. Block diagram of voltage-to-frequency converter.**

The output of the preamplifier is fed to the *voltage-to-frequency converter* (vfc) which is shown in block diagram form in Fig. 6-5. A current proportional to the preamplifier output voltage is applied to an integrating capacitor. The nearly constant current applied to the capacitor causes the voltage across it to increase linearly with time (assuming that there is a positive output from the preamplifier). The output of the control amplifier, therefore, falls linearly (the control amplifier inverts) until the output becomes sufficiently negative to allow the controlled multivibrator (cmv) to produce one pulse. This pulse, applied to the positive charge dispenser, causes it to send a charge into the integrating capacitor. This in turn causes the voltage across the capacitor to decrease. This action results in the waveform across the capacitor having a sawtooth shape. The resultant decrease in output voltage from the amplifier causes the cmv to stop oscillating until the amplifier output voltage again reaches the triggering level. The voltage appearing across the integrating capacitor is kept at a very low level because of the high gain of the control amplifier. To keep this voltage low the average current from the charge dispenser must be equal to $Q$ (the charge quantum per cycle) and time, $F$ (the number of charge quanta per second). Consequently, the frequency is precisely proportional to the input current. The accuracy of the system thus depends on maintaining the charge quantum precisely constant and keeping the voltage across the integrating capacitor very small, so that the input current is proportional to the input voltage.

If the input signal is of negative polarity, the integrating capacitor tends to charge in the negative direction. Consequently, the output of the control amplifier is positive, causing the negative rather than the positive charge dispenser to operate. The quantum charge from the negative charge dispenser is of opposite sign to that from the positive charge dispenser and therefore again tends to reduce the voltage appearing across the integrating capacitor. The pulse-train outputs are derived from the standard charge dispensers (scd) by means of transformers. Since only one scd operates at a time, there is only one pulse-train output at any given time.

The *counter section* is shown in block diagram form in Fig. 6-6. This circuit consists basically of a combining circuit and a conventional 300-kc counter.

The combining circuit adds the pulse trains from both the positive and negative outputs of the vco and provides a single pulse train equal to the total. The vco output, which at any instant is providing the pulses, is sensed and used as a polarity indication. The combined output is discriminated in the overload circuit, the output of which controls a relay. When an overload exists, the relay is de-energized and the numerical Nixie display is extinguished and the word Overload appears in the lower-left corner of the display window. The combined output is also delivered to the counter input through the Gate Time switch when in the voltmeter sector. In the frequency sector the counter is connected to the output of a 300-kc bandwidth preamplifier, whose input is connected through an attenuator to the counter input.

To describe the general operation of the counter, let us consider that the Reset one-shot has been triggered in some manner. The

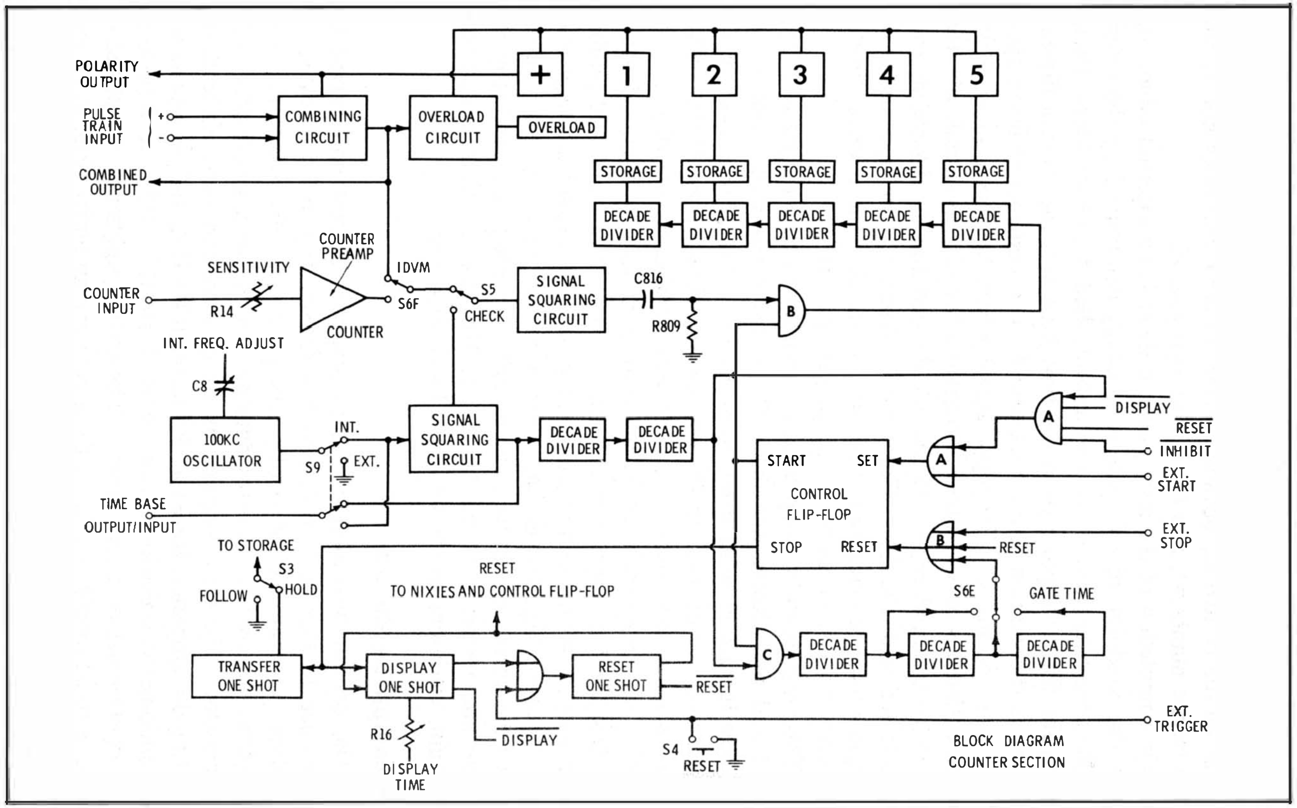



Fig. 6-6. Block diagram of counter section.

reset pulse acts to (1) reset the display one-shot to its relaxed state, (2) reset the control flip-flop to the stop state through OR gate B, (3) clear all the controlled decade dividers to zero (this does not include the stored numbers), and (4) inhibit AND gate A.

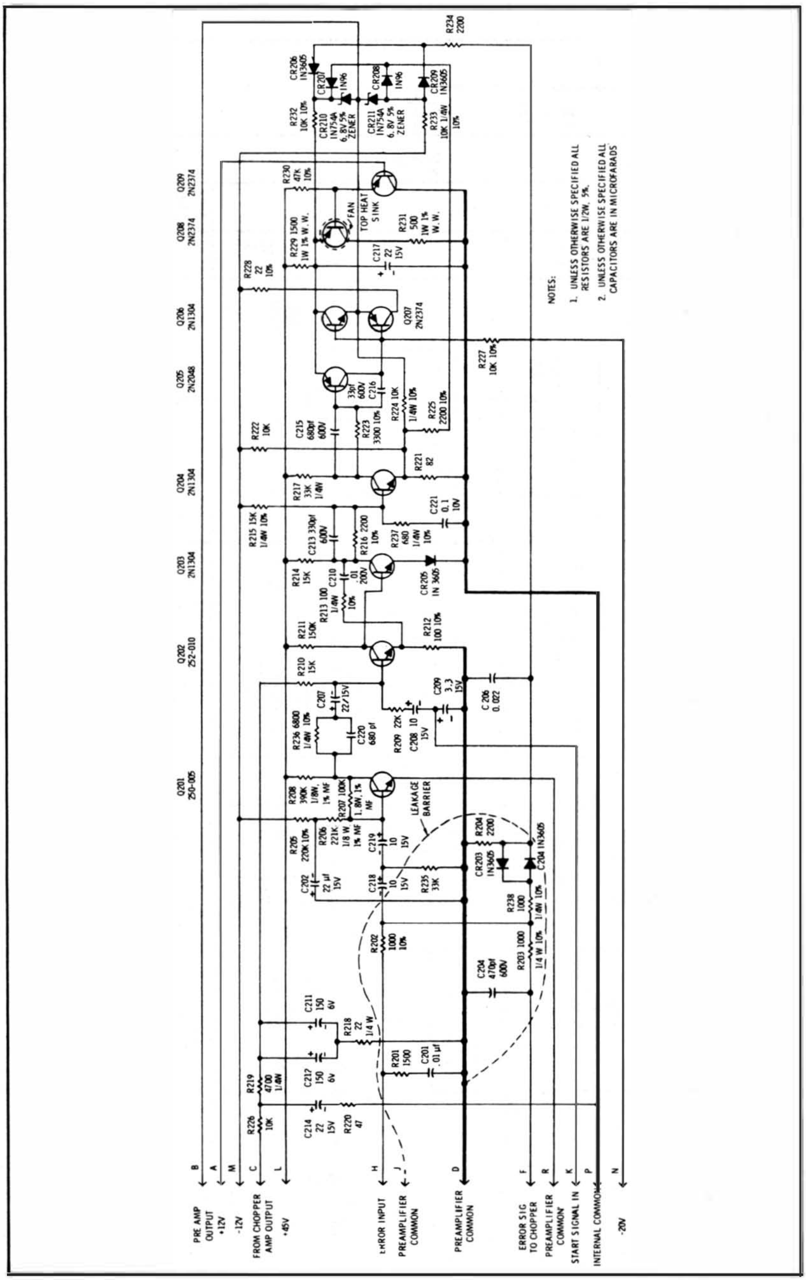
On completion of the reset pulse, if there is no external inhibit signal, the condition of AND gate A is totally determined by the output of the time base through the two uncontrolled decade dividers. Therefore, the first true pulse from the time base sets the control flip-flop to the start condition through OR gate A. This control condition opens AND gates B and C. The input signal which has been squared and differentiated is now counted through the five display decade dividers. The time base pulses are also decade divided and the desired division is selected with the Gate Time switch. When a pulse appears at the output of the selected time-base decade divider, the control flip-flop is reset to the stop condition through OR gate B.

This stop condition then initiates a trigger to both the transfer one-shot and the display one-shot. The transfer one-shot pulse, if the storage feature is being used, transfers the count accumulated in the display decade dividers to storage and the Nixie tubes. If the Follow state is used, the count is continuously carried on the display. The display one-shot, when set by the stop as previously discussed, starts the Display Time, which is controlled at the front panel and inhibits AND gate A. At the end of the display one-shot period the reset one-shot is again triggered to start a repetition of the cycle. Note that at any time during the entire cycle described, a reset pulse will stop the cycle and restart it at the beginning. External controls can initiate certain conditions as shown in the block diagram.

### Preamplifier Circuits

The overall operation of the preamplifier is best understood by reference to Figs. 6-7, 6-8, and 6-9 (Note: Fig. 6-7 is located on foldout page at end of book). As previously discussed, the preamplifier consists of an error amplifier and a precision divider (step attenuator). The error amplifier is a high-gain, chopper-stabilized, inverting d-c amplifier. Resistors R5 through R9 (Fig. 6-7) comprise the potentiometric precision divider, while R10, R11, R3, and R30 form the operational precision divider. The error amplifier is comprised of a wide-band amplifier (Fig. 6-8), a chopper amplifier (Fig. 6-9), and a chopper (Fig. 6-7).

The precision attenuator has six positions, three for the potentiometric mode and three for the operational mode. The desired pre-




Courtesy Vidar Corporation

**Fig. 6-8. Schematic of wide-band amplifier.**

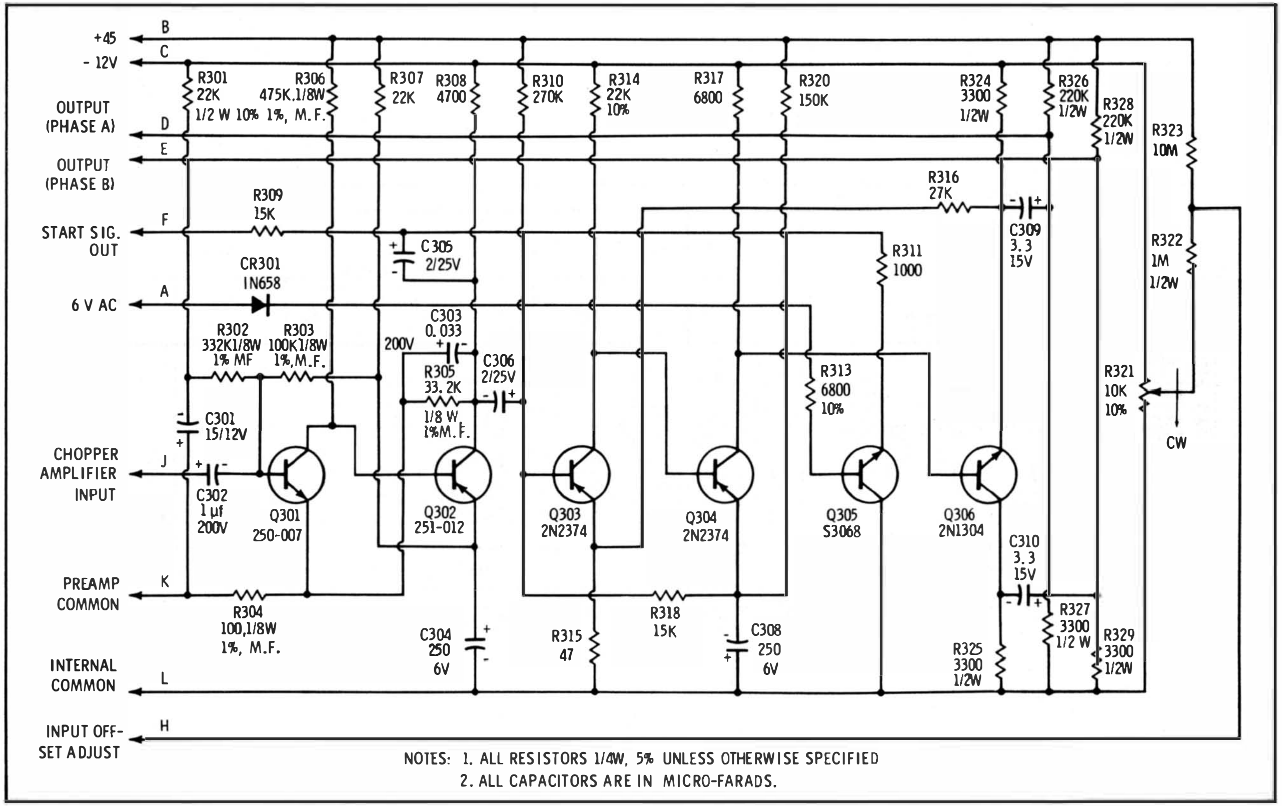



**Fig. 6-9. Schematic of chopper amplifier.**

Courtesy Vidar Corporation

amplifier gain is selected by the appropriate position of the three-pole switch (S10A-C), which is maximum in the 10-millivolts (potentiometric) position and minimum in the 1000-volts (operational) position. The input to the preamplifier is applied between the terminals marked Signal Input Lo and Hi. The terminal marked Hi is connected to a tap on the precision divider in the potentiometric mode and to internal common in the operational mode. The terminal marked Lo is connected to the error amplifier through a combination of series resistors (R1, R2, and R4) and fuses (F2 and F3) in the potentiometric mode and to the input resistor (R10) in the operational mode. Resistors R1, R2, R4, R12, and R31 and capacitors C5 and C10 along with RC network R201-C201 provide high-frequency stabilization for the error amplifier.

Back-to-back diodes CR1 and CR2 in conjunction with fuses F2 and F3 provide protection for the error amplifier in case of heavy overload (in normal operation the diodes do not conduct). R202 limits the current applied to the error amplifier in case of heavy overload. At the right end of R202, the error-amplifier input signal is split in terms of frequency.

The signal components from direct-current to about 15 cps are applied to the chopper amplifier via R203 and the input section of chopper K1. Alternating-current signal components are capacitively coupled (C218-R235-C219) to the first section of the wide-band amplifier, which amplifies signals in the range from about 15 cps to several hundred kilocycles.

The filter network of R203 and C204 prevents spikes associated with the action of the chopper from reaching the wide-band amplifier. This amplifier is a cascade of three individual amplifier sections. The first amplifier section is a single-stage, inverting a-c amplifier. The first amplifier section, as do the second two sections, uses local negative feedback to ensure gain stability. The second amplifier section is a two-stage, noninverting d-c amplifier. The second amplifier section amplifies both the 15 cps to 100-kc output of the first amplifier section and the d-c to 15-cps output of the chopper amplifier. The third amplifier section has a relatively high output current capability. This capability is required to drive the parallel combination of the precision divider section, the vco section, and the overload circuit.

The chopper amplifier operates in conjunction with chopper K1 and consists of two individual amplifier sections. The first amplifier section is a two-stage, noninverting a-c amplifier. The first amplifier section, as does the second, uses local negative feedback to ensure gain stability. The second amplifier section is a three-stage amplifier and phase inverter. The operation of the chopper-amplifier combination is best understood by imagining a small positive d-c voltage present at the input of the error amplifier (the small positive voltage might be due to a difference between the tap on the precision divider and the input signal voltage). The chopper is driven at line frequency (50 to 60 cps), and the chopper arms are energized in phase (as drawn, both arms are either up or down). The arm of the chopper input section (the input of the chopper amplifier) is therefore connected alternately between the small positive d-c voltage and a point which is near zero potential.

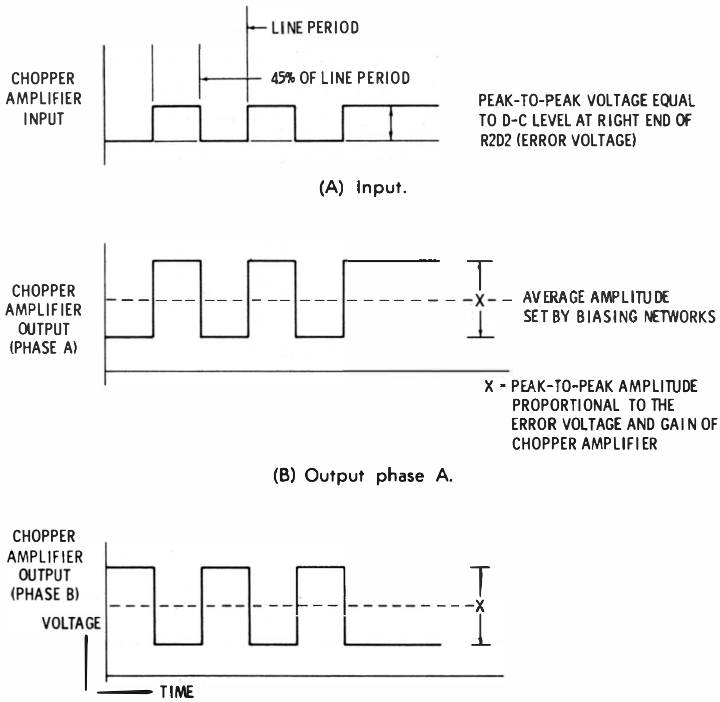
It should be noted that the upper contact of the chopper input section is not at zero potential, due to the d-c drop across R22. The voltage drop across R22 is adjusted by means of potentiometer R321 such that the potential between the chopper input-section contacts is zero with no input to the preamplifier.

The input to the chopper amplifier is therefore a series of line-rate pulses with peak-to-peak amplitude equal to the small positive d-c voltage (error voltage). These pulses are then amplified and phase-inverted by the chopper amplifier. The chopper-amplifier output signals are designated as phase A and phase B. Phase A is in phase with the chopper-amplifier input, while phase B is 180° from it. Divider networks R326-R327 and R328-R329 bias the two output phases to an average positive level. The chopper-amplifier input and output waveforms are shown in Fig. 6-10.

Note that the waveforms are not symmetrical square waves. The chopper is constructed so that the contacts in the input section are closed for 45 percent of the line period, while the contacts on the output section are closed for 35 percent of the line period. Connection to the output section of the chopper is made in such a way that the chopped arm alternately samples 35 percent of line-period portions of the phase A and phase B outputs. The waveforms at the output section of the chopper are shown in Fig. 6-11.

Note that the chopper output, relative to the d-c bias level, is an average negative d-c voltage (the output of the chopper is inverted relative to the input).

The chopper-amplifier output is applied to a low-pass filter (resistors R226, R220, R219, R218, and capacitors C214, C211, C212). The smoothing action of the low-pass filter removes the a-c component of the chopper-amplifier output (the a-c component is due to noise in the chopper amplifier). The output of the low-pass filter is




(A) Input.

(B) Output phase A.

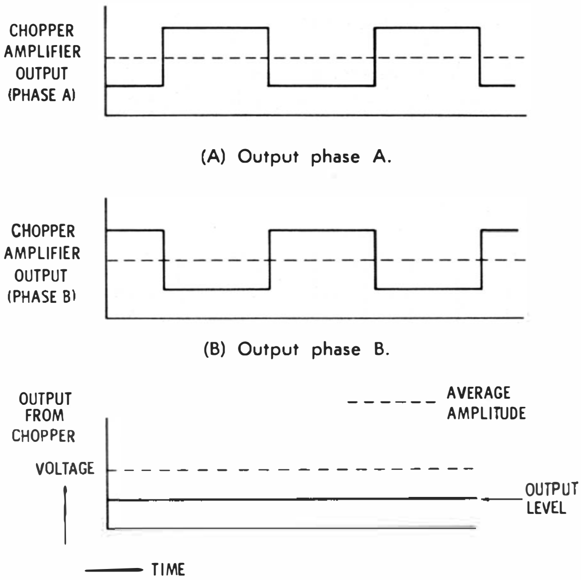
(C) Output phase B.

**Fig. 6-10. Diagram of chopper waveforms.**

combined (via resistor R210) with the output of the first amplifier section of the wide-band amplifier. Also combined at this point is the output of the start circuit (Q305).

The operation of the start circuit is as follows. Q305 functions as a synchronous rectifier by clamping to zero (common) potential alternate half-cycles of the pulse waveform which appear at the output of the first amplifier section of the chopper amplifier. The base of Q305 is driven (via CR301 and R313) by a 6-volt a-c line rate signal. This same signal energizes chopper K1. Q305 conducts during alternate half-cycles of the line-rate drive signal. The waveform at the junction of R309 and R311 is shown in Fig. 6-12.

This waveform is applied to low-pass filter (R309-C209). The d-c voltage across C209 is proportional to the chopper amplifier input voltage, while the a-c voltage across C209 is proportional to the change of chopper-amplifier input voltage. The start circuit makes use of the a-c signal across C209. This signal is summed (via C208-R209)




(A) Output phase A.



(B) Output phase B.



(C) Output from chopper.

**Fig. 6-11. Diagram of chopper output-section waveforms.**

with the filtered chopper-amplifier output and the output of the first amplifier section of the wide-band amplifier. The a-c signal across C209 anticipates a change of voltage at the filtered output of the chopper amplifier. A greater time is required to change the voltage at the output of the chopper amplifier. This is due to the relatively long time constant of the chopper-amplifier output filter, it is compounded by the saturation of the second amplifier section of the chopper amplifier. This occurs when the instrument is first turned on. The anticipating signal derived from the start circuit enables the error amplifier to reach a stable operating condition following the

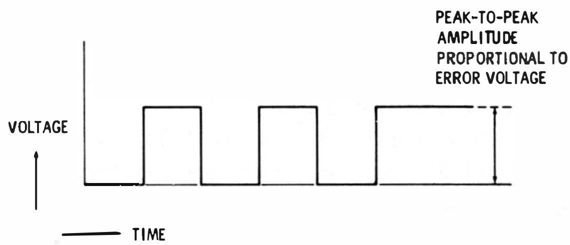



**Fig. 6-12. Waveform at junction of R309 and R311.**

initial condition which exists when power is first applied to the instrument.

The combined outputs of the start circuit, the chopper-amplifier–chopper combination and the first amplifier section of the wide-band amplifier, are amplified by the second and third (output) sections of the wide-band amplifier. The output stage of the wide-band amplifier consists of a complementary symmetrical emitter-follower connected between +12 volts and −12 volts. It is desirable to restrict the preamplifier output range to a value slightly in excess of the normal operating range (+6.25 volts to −6.25 volts). Restriction of the preamplifier output range is accomplished by the overload circuit. The effect of the restriction is to decrease the susceptibility of the error amplifier to overload from abnormal input signals.

Operation of the overload circuit is as follows. Current flows through zener diodes CR210 and CR211 via resistors R232 and R233, respectively (note that R232 and R233 are returned to +12 volts and −12 volts, respectively, and that the junction of CR210 and CR211 is returned to the preamplifier output). The voltage at the junctions of R233-CR211 and R232-CR210, therefore, differs from the preamplifier output voltage by the zener voltage (6.8 volts ±5 percent). The voltage at the junction of R233-CR211 is 6.8 volts negative with respect to the preamplifier output, and the voltage at the junction of R232-CR210 is 6.8 volts positive with respect to the preamplifier output. Therefore, the voltage at the junction of R233-CR211 will be zero when the preamplifier output voltage is 6.8 volts, and the voltage at the junction of R232-CR210 will be zero when the preamplifier output voltage is −6.8 volts.

Diodes CR206-CR207 (connected to the junction of R232-CR210) and CR202-CR209 (connected to the junction of R232-CR211) are connected to near zero (internal-common) potential points. Diodes CR208 and CR209, therefore, conduct when the potential at the junction of R233-CR211 is about zero. Similarly, CR206 and CR207 conduct when the potential at the junction of R232-CR210 is about zero.

Note that since CR207 and CR208 are germanium diodes and CR206 and CR209 are silicon diodes, CR207 will conduct prior to CR206 and that CR208 will conduct prior to CR209. Therefore, R207 or CR208 will go into conduction when the preamplifier output reaches about plus or minus 6.8 volts. When CR207 or CR208 go into conduction, R224 the wide-band amplifier output-stage feedback resistor) is shunted by R225, thereby reducing the wide-band amplifier gain, hence reducing the error-amplifier gain. A further increase in preamplifier output causes either CR206 or CR209 to conduct. When CR206 or CR209 conduct, overall feedback is applied around the error amplifier via divider network R234, R204, C206), back-to-back diodes CR203 and CR204, and resistor R238.

The effect of the overall feedback is to greatly reduce the input impedance of the error amplifier (the current fed back around the error amplifier opposes the input current). The greatly reduced input impedance tends to prevent a further rise in input voltage with consequent saturation of the error amplifier and subsequent prolonged recovery time.

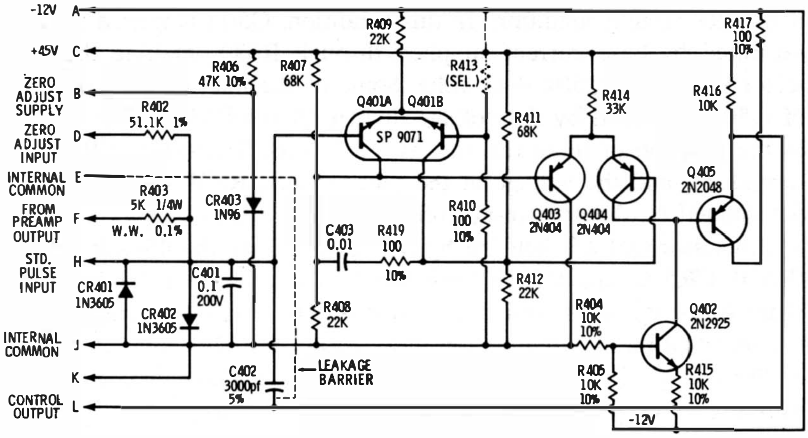
The +12 volts necessary for operation of the wide-band amplifier output stage and the overload circuit is derived from the +45-volt supply by means of a shunt regulator (Q208-Q209). The +12-volt output of the vco power supply provides the reference voltage for the +12-volt supply.

Several precautions are taken to enhance the performance of the preamplifier in terms of stability and noise. These include the use of high-accuracy resistors in the precision divider, the use of noise-selected transistors in low-level stages, the use of shielded wiring in low-level stages, and a carefully worked-out wiring arrangement.

### Voltage-to-Frequency Converter Circuits

The *control amplifier* (Fig. 6-13) amplifies the sawtooth waveform appearing across the integrating capacitor along with any d-c level appearing at this point. The output of the control amplifier drives the standard charge dispensers. The amplifier must have wide bandwidth in order to amplify the sawtooth waveform with good fidelity. The amplifier must also have high gain and low input offset voltage in order to maintain the d-c voltage appearing across the integrating capacitor at nearly zero, thus making the input current accurately proportional to the input signal voltage.

The control amplifier consists of two cascaded differential amplifiers (Q401A-Q401B and Q403-Q404) followed by an emitter follower (Q405). Q402 acts as a high-impedance load for Q404. The control-amplifier input signal (the preamplifier output) is fed to the base of Q401 via input resistor R403. Back-to-back diodes (CR401 and CR402) protect the control amplifier from excessive input voltages. Q401 base-to-ground voltage is set at zero by adjustment of the first differential-amplifier operating point. This adjustment is provided by voltage divider R413-R410 in the base circuit of Q402 (note that




NOTES: 1. UNLESS OTHERWISE SPECIFIED
A. ALL RESISTORS 1/2W, 5%
B. ALL CAPACITORS IN MICROFARAD

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**Fig. 6-13. Schematic of control amplifier.**

selected resistor R413 may be connected to either +45 volts or −12 volts.) A variable offset current derived from the +45-volt supply via potentiometer R23 (the zero calibrate control) and resistors R402 and R406. The offset current may be adjusted to give zero-frequency output from zero input to the meter. The control-amplifier output is taken from the emitter of Q405. The capacitor (C402) provides the amplifier with overall integrator-type feedback which aids in forming the sawtooth triggering waveform.

Several precautions have been taken to reduce d-c drift in the amplifier. The differential amplifiers both make use of low leakage silicon transistors. In addition, the input pair (Q401A-Q401B) shares a common header and uses transistors which have been matched for emitter-base voltage tracking over the operating temperature range.

*Positive Standard Charge Dispenser*—The positive standard charge dispenser (scd) is driven by the output of the control amplifier and supplies the precision quanta of charge to the integrating capacitor. The charge dispenser is made up of a controlled multivibrator (cmv), an emitter-follower amplifier, a switch transistor which drives the precision capacitor, switch diodes, and a constant-current source (Fig. 6-14).

Transistors Q501 and Q502 comprise the cmv. The output of the cmv is amplified by emitter follower Q503 and is used to drive the switch transistor (Q504). Consider the operation of the circuit when it is in the resting condition. In this condition, Q504 is turned on and saturated by base current supplied through R511, making the collector voltage of Q504 +45 volts. Some of the load on the collector of Q504 is caused by the voltage divider (R516-R517). The voltage at the junction of these resistors is +6.3 volts. Therefore, CR505 is turned off, and the voltage at the junction of CR501 and CR502 is determined by the control-amplifier output voltage.

Transistor Q502 is held off by the current from the 45-volt supply (R510, CR502, and R505). Since one half of the cmv is held off, the other transistor is on, but not saturated. The current through Q501 is controlled by maintaining the base voltage at −2.9 volts by means of the voltage divider (R501-R502). Current in Q501 is determined by the emitter voltage and R502. The cmv is thus held in a stable state for all voltages from the control amplifier which are more positive than the −2.7 volts on the emitter of Q502. In the operating condition, however, the output of the control amplifier becomes negative and will eventually become more negative than the voltage at the base of Q502. When this occurs, Q502 turns on, and the cmv is triggered.

The normal operation of the cmv then turns off Q501 because of the rise in the collector voltage on Q502 which is transmitted through C502. The collector voltage on Q501 falls and causes Q502 to stay on, regardless of the voltage at the output of the amplifier. The rise in voltage at the collector of Q502 is also coupled to the emitter follower and then to the base of the switch transistor, Q504. The switch transistor is thus turned off, and its collector voltage falls toward ground level. The fall in collector voltage makes the junction of R516 and R517 become negative, thereby turning on CR505 and turning off both CR501 and CR502. The influence of the control amplifier is removed from the cmv until the switch transistor is again turned on. The switch transistor (Q504) is turned on at the end of the pulse produced by the cmv.

In normal operation the output of the control amplifier will have decreased below the triggering level in this length of time due to the standard charge dispenser will have sent a standard unit charge into the summing point. If this is true, the circuit will remain in the resting state until the sawtooth waveform at the output of the control amplifier again becomes sufficiently negative to trigger Q502. Consequently, in this mode of operation the cmv operates as a controlled multivibrator producing one pulse each time the output of the control amplifier becomes more negative than −2.7 volts. If the unit is over-

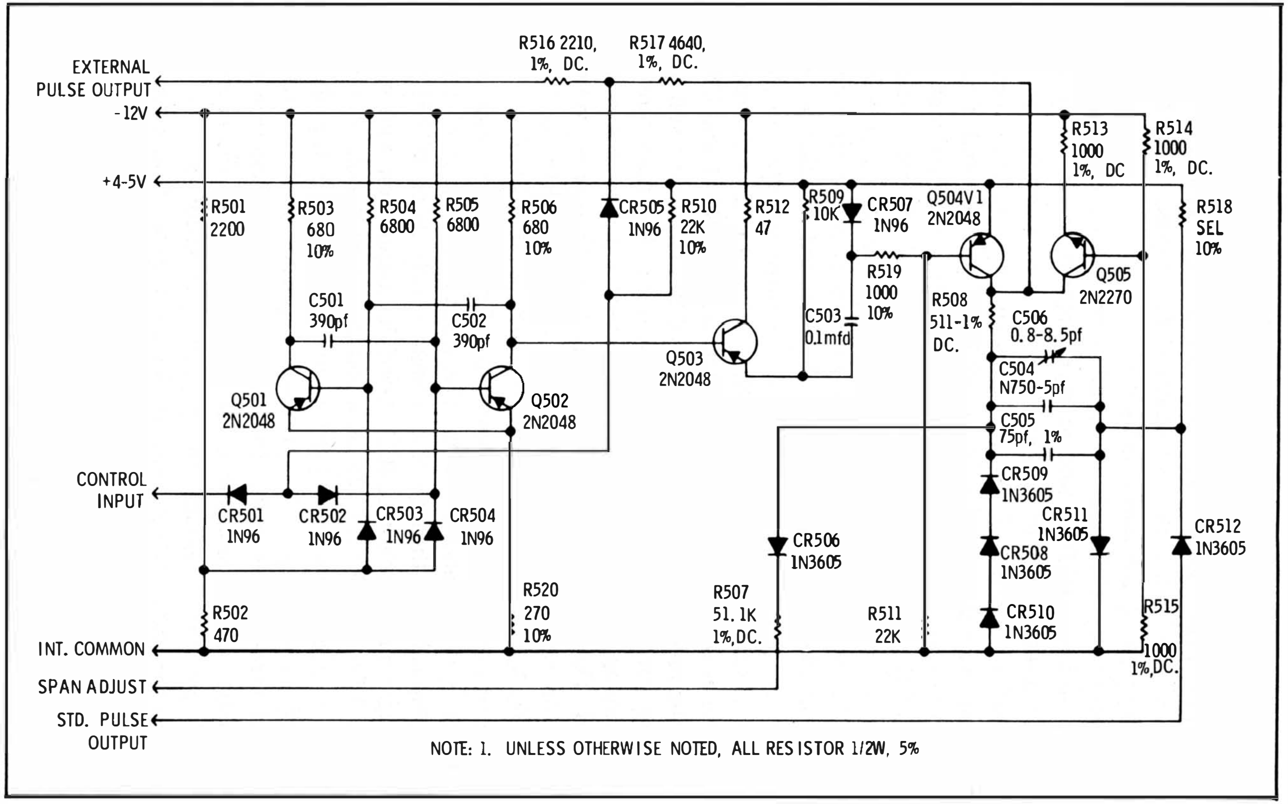


**Fig. 6-14. Schematic of positive standard charge dispenser.**

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loaded by the application of a large positive signal, the output of the control amplifier will become more negative than $-2.7$ volts, and the multivibrator will operate as a free-running multivibrator. The cmv also "free-runs" when the instrument is first turned on and has not established its equilibrium condition.

As previously discussed the output of the cmv is amplified by the emitter follower (Q503) and is then applied to the base of the switch transistor (Q504). This transistor is normally on when the standard charge dispenser is inoperative. In this case the precision capacitor (C504 in parallel with C505 and C506) is charged to some positive value. The exact voltage is determined by the setting of R24 (the plus calibrate control). R24 sets the voltage applied to the catching diode (CR506). The effect of changing R24 is to adjust the positive voltage to which the precision capacitor charges.

The standard quantum of charge is generated by alternately connecting the precision capacitor to the positive voltage and then to ground as Q504 is turned on and off. The right end of the precision capacitor is connected to the two switch diodes (CR511 and CR512). When the precision capacitor is connected to the positive voltage, the charging current is through CR511 to ground. The time constant of the charging circuit is such that the capacitor becomes nearly fully charged. Thus a charge ($Q = CE$, where E is the positive voltage) is placed on the capacitor. Turning the switch transistor (Q504) off causes the left end of the precision capacitor to fall towards $-12$ volts, but the left end is caught at about ground potential by clamp diodes (CR508, CR509, and CR510). The discharge current of the precision capacitor is through the second switch diode (CR512) into the integrating capacitor (C401). Q505 in conjunction with resistors (R513, R514 and R515) acts as a constant-current sink, feeding the charge to the precision capacitor and thereby reducing the time required to discharge. The precision capacitor is completely discharged through this path. Therefore, the charge delivered to the integrating network is again equal to CE. To make the charge delivered precisely constant the precision capacitor must alternate between the same voltage limits during each cycle of operation. This is accomplished by making the upper limit of the voltage excursion the preadjusted positive voltage and lower limit the diode clamp voltage. The precision capacitor is chosen to have a very small temperature coefficient. The residual coefficient is compensated by the negative temperature-coefficient capacitor (C504). The exact amount of charge delivered per cycle can be adjusted by the trimming resistor

(R24). Adjustment to compensate for component tolerances in the charge dispenser is accomplished by the trimmer capacitor (C506). R24 allows adjustment over a range of about ±1 percent.

The pulse-train output of the unit is taken from the collector of Q504 via the output transformer (T2). The transformer is loaded by R26.

*Negative Standard Charge Dispenser*—The negative and positive standard charge dispensers are similar in all respects except for input signal polarity and output charge polarity. In the negative scd (Fig. 6-15), transistors Q601 and Q602 comprise the controlled multivibrator (cmv), Q603 is the emitter-follower amplifier, Q604 is the switch transistor, and Q605 is the curent source.

In the negative scd, the switch diodes are reversed and the cmv transistors are npn types. The cmv is triggered when the input from the control amplifier becomes more positive than the voltage across R603. The cmv provides the signal input to Q603 and the npn emitter follower which drives switch transistor Q604. Q604 is normally in the off condition. Therefore, in the resting condition the precision capacitor (C606 in parallel with C607 and C608) is discharged. Note that in the negative scd the polarities of the two switch diodes (CR611 and CR612) are reversed from those in the positive scd. Therefore, the sign of the charge delivered to the integrating capacitor is also reversed.

The pulse-train output is taken from the collector of Q604 via the output transformer (T3). The transformer is loaded by R27.

*VCO Power Supply*—The voltage-controlled oscillator (vco) power supply provides two regulated outputs, −12 and +45 volts, and one unregulated output of −22 volts. Operation of these circuits is typical for power supplies of this type. These circuits will not be discussed since they are not unique to an integrating digital voltmeter.

*Pulse-Combining and Counter Circuits*—A simplified composite schematic of the counter section is shown in Fig. 6-16 (Note: Fig. 6-16 is located on foldout page at end of book). This entire section is contained on six basic types of printed-circuit boards. They are the input processing circuits (PC801), the counter control circuits (PC701), the time-base source (PC1001), the decade dividers (PC1101-PC1105), the decade displays (PC901-PC905), and logic conversion circuits (PC1201-PC1301). The manually switched logic and circuit interconnections are also shown in schematic form on this drawing. The decimal-point location logic is done with switches S8A, S8B, S8C, and S6B. The unit's logic is performed by switches S8D

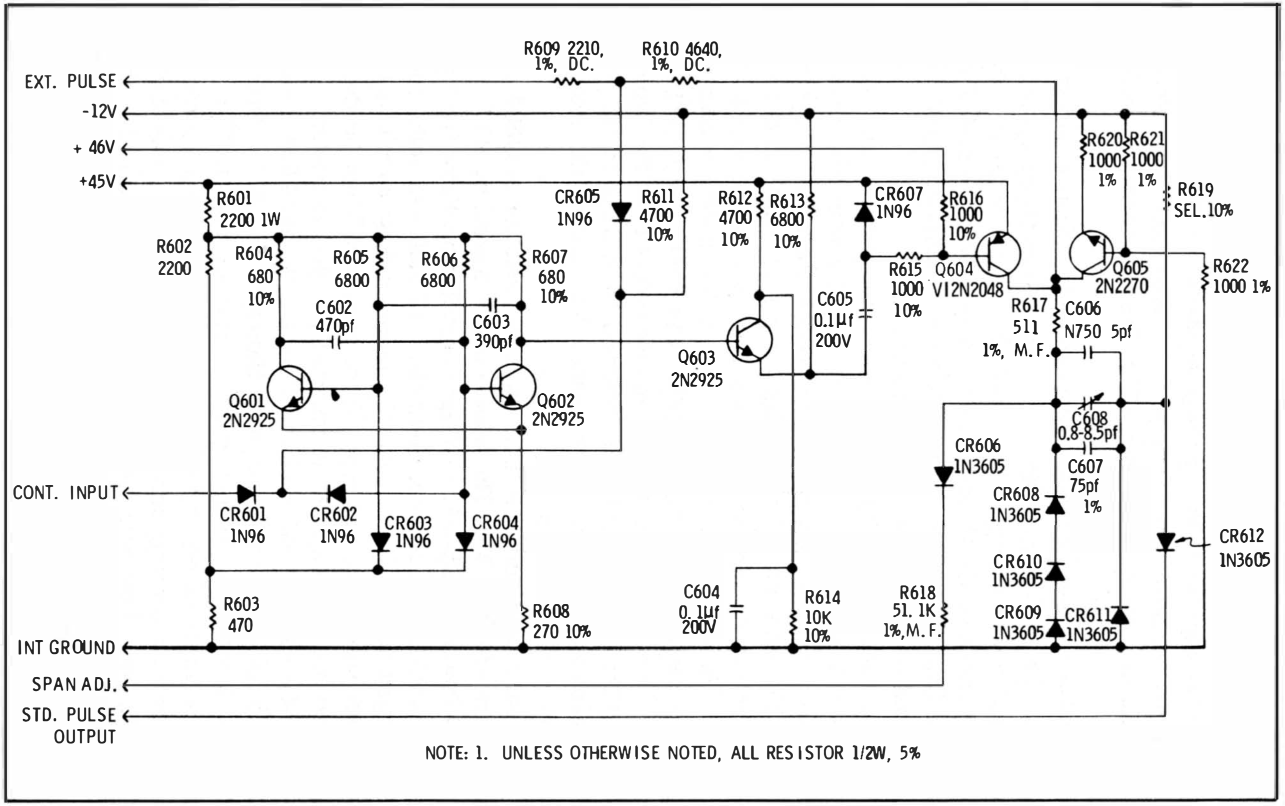


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**Fig. 6-15. Schematic of negative standard charge dispenser.**

and S6D. Negative-true logic is employed throughout the counter section with "zero" equal to or greater than $-1.2$ volts and "one" equal to or less than $-9.5$ volts.

*Pulse-Combining and Polarity Circuits*—These circuits are contained on the input board (PC801 in Fig. 6-16). The positive pulse-train output from the voltage-to-frequency converter is amplified and inverted by Q801, then differentiated by C804 and R814, while the negative pulse-train output is directly differentiated by C802 and R813. The positive swing of either pulse train will therefore trigger the one-shot comprised of Q806 and Q807. The secondaries of the transformer load (T801) of Q807 are cascaded and applied to the counter input. The center winding of T801 is loaded by R20 and available as the combined output pulse train. The negative swing of the positive and negative differentiated pulse trains is applied to the bistable multivibrator composed of Q804 and Q805 in such a manner as to set it to one state or the other, depending on which output of the voltage-to-frequency converter is producing pulses.

The high voltage is applied to the plus or minus Nixie tube through the Gate Time switch (S6C) in the voltmeter sector only and through the current-limiting resistor (R812) to provide a continuous minus sign display. When a pulse appears at the positive output of the converter, Q804 is turned on and the plus sign lights through the driver transistor (Q809). A logic level is also provided from this flip-flop for polarity indication, a "one" signifying positive voltage input.

*Overload Circuit*—The output pulses of the combining circuit are constant in duration and therefore may be used in a standard charge dispenser much the same as previously described. In this case the average current and supplied curent is integrated by a capacitor and the differential current is sensed. When this differential current which is proportional to the pulse repetition rate, exceeds a predetermined value, a relay is deactivated which illuminates the overload indicator, extinguishes the display, and provides a contact break from ground on pin 45 of J10.

The overload detection circuit is contained on the control board (PC701) and is shown at the lower left of Fig. 6-16. The relay and indicator are mounted on the bottom of the decimal point assembly.

The combined output at J7 is applied to the base of Q713 which is normally biased off by the low source impedance of T801 and the difference between the base-emitter drop of the silicon transistor (Q713) and the forward drop of the germanium diode (CR710). When a positive pulse occurs both Q713 and Q714 are turned on for the duration of the pulse, taking the collector of Q714 to the +15-volt level and charging capacitor C701 through R719 and CR720. At the end of the input pulse, Q713 and Q714 turn off again, and C701 is discharged through CR721, R719, and R712. The left end of capacitor C701 is caught at ground potential by diode CR719. The original charge on the capacitor C701 was $C \times 15$ volts, since the drop in the switch transistor is negligible. The charge delivered to the integrating capacitor (C722) is thus the initial charge minus the final charge which is very nearly zero because the choice of the pulse length is sufficient to allow complete discharge of C701.

Each pulse transfers the same quanta of charge. Therefore, the average current out of the integrating capacitor is proportional to the pulse repetition rate. Resistors R720 and R722 provide current to the circuit, and Q715 is turned on as long as insufficient curent is being removed from the circuit by the charge dispenser (as a result of the pulse rate being too low). As the pulse rate increases, a point (adjustable by R720) is reached at which insufficient current is available to maintain conduction of Q715, and it turns off. Q716 provides temperature compensation and allows Q715 to be a nonsaturating switch. When Q715 turns off, Q717 also is turned off, and the relay (K2) is thereby de-energized. One contact of K2 is available as a break in continuity on the Bcd Output connector, J10. The other opens the short circuit across the overload indicator neon lamps and disconnects the high voltage from the Nixie numerical display. The full high voltage applied across the series overload neons causes their breakdown, and the sustained breakdown voltage of both neons in series is large enough to hold the Nixie high-voltage line below their breakdown voltage level. Resistor R40 limits the current through the overload neons.

*Counter Preamplifier*—The counter input is applied through C6 and R13 to the arm of the potentiometer (R14). C7 produces high-frequency boost to offset cable losses due to the high impedance of the attenuator. Transistors Q810, Q811, and Q814 located on the input board (PC801) comprise a 300-kc bandwidth, high-gain amplifier with voltage feedback across R830. The high-frequency roll-off is accomplished by the capacitor (C814) and the low-frequency cut-off by C809. Emitter follower Q814 reduces loading of Q811 to maintain high open-loop gain and also provides low output impedance to reduce loading effects.

*Time-Base Oscillator*—The time oscillator and Schmitt trigger are located on the source board (PC1001). The oscillator is crystal controlled and is a switching-type network consisting of transistors (Q1004 and Q1005) and the crystal (X1001). The crystal can be considered as a series-tuned circuit, driven from a current source and resonant at 100 kc. Due to the low input impedance of Q1004 nearly all the current through the crystal must flow into the base of Q1004. The regenerative action of the emitter coupling requires the collector of Q1005 to be in phase with the base of Q1004. Therefore, the series resonant circuit allows sustained oscillation only at the resonant frequency when zero phase shift exists. The trimmer capacitor (C8) allows a small amount of phase shift to be introduced at the base of Q1004 and thereby "pull" at the frequency of oscillation slightly (about ±3 cps).

The output of the oscillator is capacitively coupled through C1005 and R29 to switch S9 where it is either grounded when an external time base is used or applied to the base of the amplifier transistor (Q1001). The collector of Q1001 is then coupled to a conventional Schmitt trigger circuit composed of Q1002 and Q1003. The collector of Q1002 provides a signal at time base used in the counter Check mode. The output of the squaring circuit goes through R1021 and C9 to S9 where it is connected through R21 to the output connector J9 when the internal time base is used. When S9 is in the external position, J9 becomes an input which is applied through R21 and C1001 to the amplifier Q1001 in the same manner as the oscillator output previously described. The Schmitt trigger output also goes to the first decinary.

*Decinary*—Five identical decade dividers are cascaded from the time base to provide output frequencies of 10 kc, 1 kc, 100 cps, 10 cps, and 1 cps. The lower three frequencies are used to establish the three fixed gate times. Gating of the time base is done at 1 kc. The three gate-time decinaries are reset each count to ensure accurate timing.

Each decinary consists of four binaries in series. A typical decinary schematic is shown in Fig. 6-17. Feedback is so arranged as to produce a bcd weighting of 1-2-4-8 for each successive binary. The carry line thereby produces one pulse for each ten applied to the input.

*Display Decades*—The five display decades each perform three basic functions: they divide the counted pulses by ten and carry every tenth pulse to the next decade in the same fashion as the decinaries; they store the accumulated count when a transfer command is received; and they convert the bcd output to ten-line code and drive the Nixie tube with the resultant information. Each of the display boards

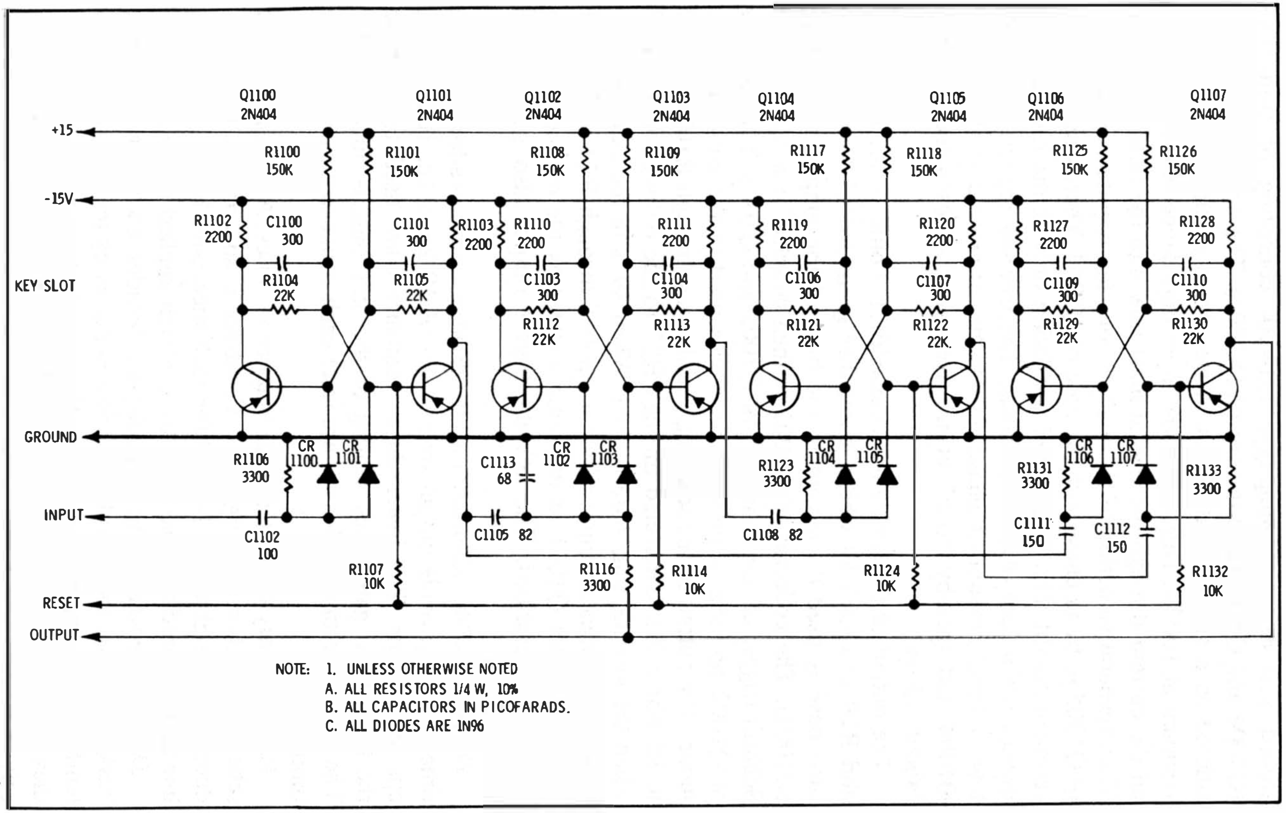


**Fig. 6-17. Schematic of decinary PC1100.**

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are identical, except for the first binary of the unit's decade which is of higher speed to ensure accurate counting through 300-kc input pulses. A typical display-decade circuit is shown in Fig. 6-18.

The decade-divider portion of the circuit functions identically to the decinary previously discussed (the carry line delivers every tenth pulse to the next display-decade input). Each binary is connected to the reset line, and each binary output is coupled through gating networks (R998, R996, R997, and R999) and is a typical storage binary. The output of each storage binary is connected to the bcd ten-line decoding matrix (R934 through R963) and the bcd output through resistors R964-R967 to connector J10. When the transfer line is grounded the levels of the decade divider binaries are transferred continuously to the storage binaries. When the transfer line is held positive the resistor ratios prohibit the information transfer from decade divider to the storage binaries. The Nixie drivers Q908 through Q917 are directly coupled to the resistor logic matrix and the collector of each to the appropriate numerical indication in the Nixie tube.

*Counter-Control and Gating Circuitry*—The output of either the combining circuit or the counter preamplifier is selected by the Gate Time switch (S6F) and then either that signal or the 100-kc Check signal is selected by S5 and applied to the input Schmitt trigger composed of Q812 and Q813 (Fig. 6-16). Capacitor C816 and resistor R809 differentiate the squarred waveform at the collector of Q813 when the base of Q803 is held negative by the start signal from the control flip-flop. The negative-going pulses are then amplified by saturating amplifier Q802 and delivered to the input of the unit's decade display divider.

The reset one-shot may be triggered from three sources: (1) closure of Reset switch S4, (2) a positive-level shift from an external source through Ext Trig (J11), or (3) by termination of the display time (transistor Q706 turning on). The negative reset pulse, of about 180-microsecond duration, is current amplified through transistor Q709 and delivered to the reset line. Decinaries PC1103 through PC1105 and all decade dividers in the display decades are set to the zero state by the reset line. The reset pulse is applied to the control flip-flop to reset it in the stop mode (Q701 on and Q702 off). Note that only if Q702 is on at the time the reset pulse occurs will a transfer pulse be generated by action of the reset one-shot. The positive reset pulse is used to terminate the display and as one of the control AND gate signals.

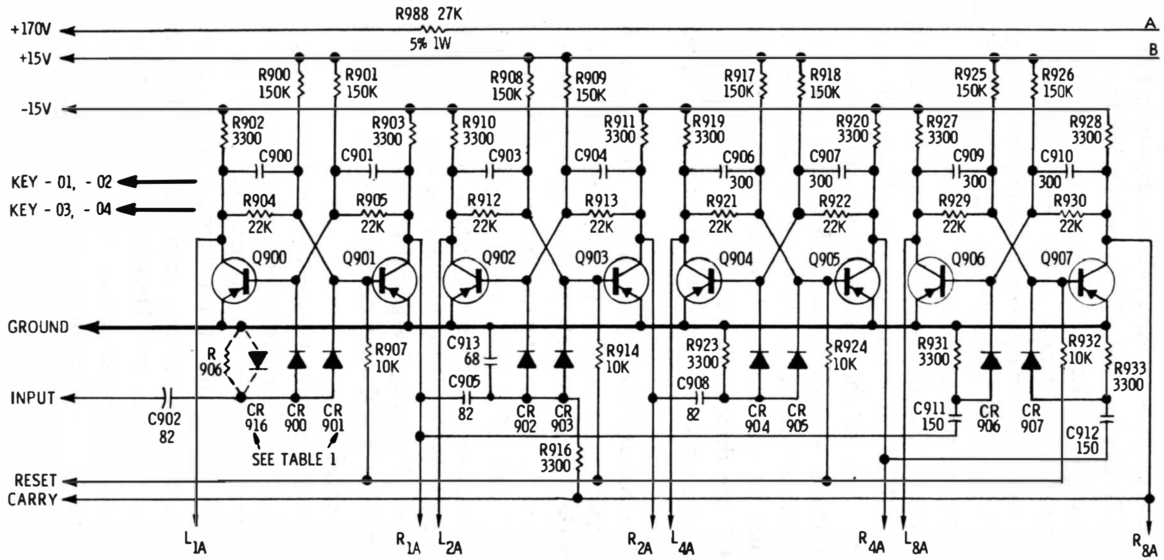
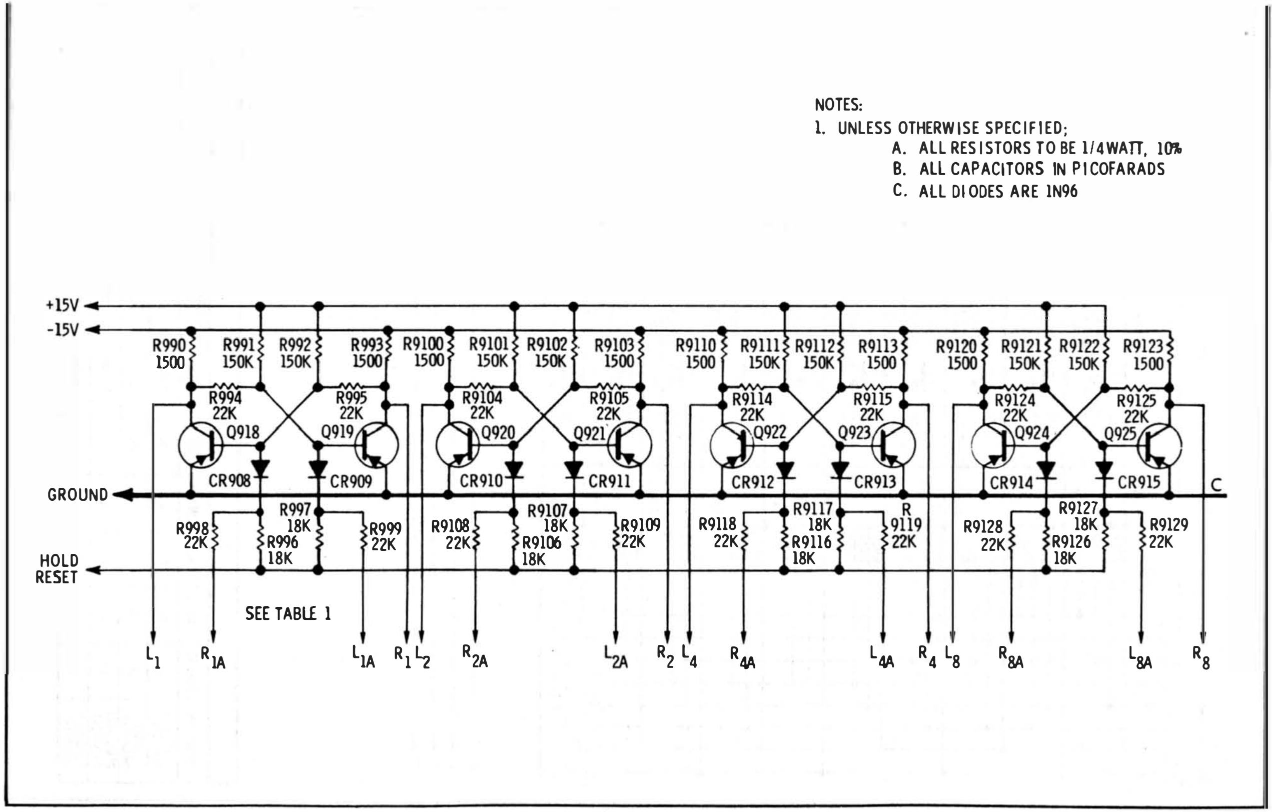



Fig. 6-18. Schematic




of display circuit (cont'd on page 118).

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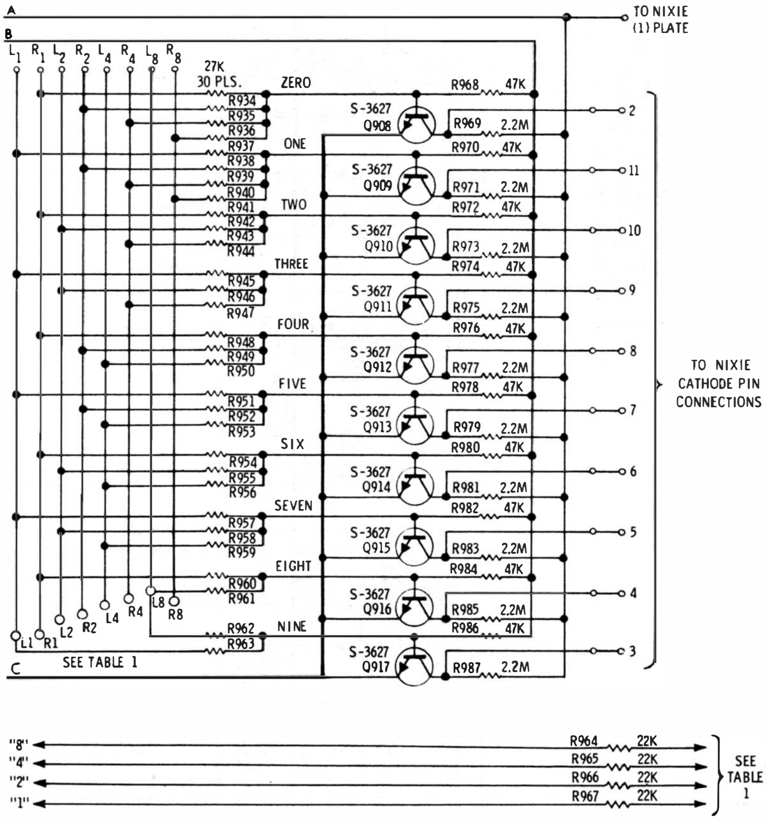


NOTES:

1. UNLESS OTHERWISE SPECIFIED;
   A. ALL RESISTORS TO BE 1/4 WATT, 10%
   B. ALL CAPACITORS IN PICOFARADS

TABLE 1

| FUNCTION | R964 TO | R965 TO | R966 TO | R967 TO | C800 C901 | Q900 Q901 | R906 OR CR916 | ASSEMBLY NO. | PC NO. | C903 C904 |
|---|---|---|---|---|---|---|---|---|---|---|
| ZERO POSITIVE | R8 | R4 | R2 | R1 | 300 | 2N404 | 3300 | 0770-01 | PC900 | 300pf |
| ZERO NEGATIVE | L8 | L4 | L2 | L1 | 300 | 2N404 | 3300 | -02 | PC900 | 300pf |
| ZERO UNIT POSITIVE | R8 | R4 | R2 | R1 | 100 | 2N2048 | 1N96 | -03 | PC901 | 120pf |
| ZERO UNIT NEGATIVE | L8 | L4 | L2 | L1 | 100 | 2N2048 | 1N96 | 0770-04 | PC901 | 120pf |

Courtesy Vidar Corporation

**Fig. 6-18 (Cont'd). Schematic of display circuit.**

Consider now the condition of the control AND gate at the base of the emitter follower (Q703). The beginning of the reset pulse has turned off Q704 in the display one-shot (assume that there is no external inhibit signal); at the completion of the reset pulse the base of Q703 will have the output of the 1-kc decinary applied to it (CR714, CR706, CR705, and CR708 all are back-biased). The first positive swing of Q703 base will set the control flip-flop to the start condition (Q701 off). The start signal applies the necessary negative voltage to the base of Q803 to "open" the input gate and also allows the time-base gate to follow the output of the 1-kc decinary. Therefore, the pulse which started the control flip-flop is also the first pulse counted in the 100-cps time-base decade dividers. Depending on the Gate Time setting, either 10, 100, or 1000 milliseconds after the start, a positive stop pulse is applied to the reset side of the control flip-flop. The stop condition of the control flip-flop (Q702 off and Q701 on) terminates the count by causing both the input and the time-base gates to "close," as well as triggering the transfer and display one-shots. The transfer one-shot negative pulse is alternating-current coupled through Q712 to effect a level shift and produce about a 33-microsecond negative pulse referenced to the positive supply on the transfer line.

There are actually three modes of operation of the display one-shot: first as a variable-duration pulse generator; second, as a bistable flip-flop; and third, as a fixed bias to one state.

In the pulse-generated mode, the Infinite switch (S2) and Gate Time switch (S6E) connect the negative supply to the junction of R18 and R17. The display one-shot duration is then controlled by the RC time constant of R16 plus R18 and C710. Emitter follower Q705 reduces effect of collector load on the time constant and provides a fast recharge time constant nearly independent of the discharge time setting of R16. Diode CR3 is back-biased and made of silicon, thereby offering very little shunt resistance to the timing resistance of the timing resistors. This period is variable from about 60 milliseconds to about eight seconds. When the Infinite switch (S2) is pulled the junction of R17 and R18 is connected to the anode of CR175 while CR3 is forward-biased. Therefore, R725 is placed in parallel with C710, and the flip-flop becomes bistable. When any of the external or manual functions are in use, S6E opens and CR3 becomes forward-biased, clamping the base of Q706 to the positive supply through R17. Under this condition Q706 is kept from conducting, and the control AND gate is clamped to chassis ground through CR706, Q704, and CR716.

In fixed gate time operation the negative trigger pulse from the "stop" signal of the control flip-flop is applied through CR717 to the base of Q704, turning it on and thereby starting the display one-shot pulse period. With Q704 on the control AND gate is biased "closed" until the termination of the display time.

The reset one-shot is triggered by this termination from the collector of Q706, and again the control AND gate is biased "closed" through CR708 and CR705, which completes the entire control cycle. The two series diodes are used to match the various output levels at the gate and thereby remove the possibility of false triggering.

### External and Manual Modes

In the External and Manual modes the display one-shot is biased to a state which will hold the control AND gate "closed." The control flip-flop then can only be triggered by signals through CR703 to start the count, and R704 to stop the count. For external control External Start (J12) is used to start, and External Stop (J13) is used to stop the control flip-flop. The external signal may be either a closure to ground or a three-volt positive step. The manual start signal is taken from the Gate Time switch (S6B) and the stop signal from S6A.

## USING AN INTEGRATED DIGITAL VOLTMETER

The integrated digital voltmeter has certain advantages over the nonintegrating type. However, the principal advantage is its ability to obtain the integral of a voltage over a given time interval. For example, assume that it is desired to obtain the integral of the nonrepetitive waveform shown in Fig. 6-19.

The voltage-time integral is equal to the area enclosed by the waveform. The area may be obtained by considering the waveform as being composed of a series of 0.2-second bars of increasing height

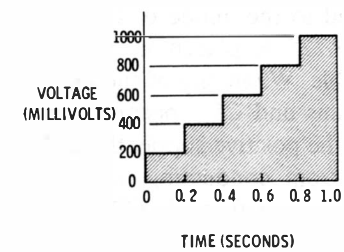



**Fig. 6-19. Graph of the integral of a nonrepetitive waveform.**

(voltage); the total area consists of the sum of the areas of the individual bars. The area (A) is then

$$A = (200 \text{ mv} \times 0.2 \text{ sec}) + (400 \text{ mv} \times 0.2 \text{ sec}) + (600 \text{ mv} \times 0.2 \text{ sec}) + (800 \text{ mv} \times 0.2 \text{ sec}) + (1000 \text{ mv} \times 0.2 \text{ sec}) = 600 \text{ mv-second}$$

If the waveform of the example were applied to an integrated digital voltmeter such as the Model 510 (set to the 1000-mv range) the voltage-to-frequency converter output would be as shown in Fig. 6-20.

When the frequency output is applied to the counter portion and if it is gated "on" at the beginning of the waveform and "off" at the end of the waveform, the reading would indicate the total number of

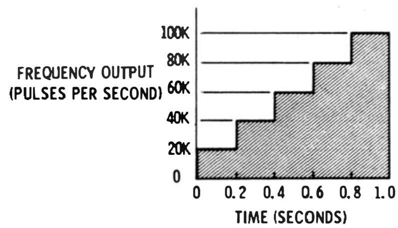



**Fig. 6-20. Graph showing output as integral in voltage-to-frequency conversion.**

pulses emitted by the converter during the one-second gating period. This number can be derived by considering the frequency output of the converter as consisting of a series of 0.2-second bars of increasing height (frequency), the total number of pulses consisting of the sum of the number of pulses emitted during each 0.2-second period. The total number of pulses (N) is

$$N = (20{,}000 \text{ pps} \times 0.2 \text{ sec}) + (40{,}000 \text{ pps} \times 0.2 \text{ sec}) + (60{,}000 \text{ pps} \times 0.2 \text{ sec}) + (80{,}000 \times 0.2 \text{ sec}) + (100{,}000 \times 0.2 \text{ sec}) = 4000 + 8000 + 12{,}000 + 16{,}000 + 20{,}000 = 60{,}000$$

Observe that not only is N (the reading) an accurate indication of the area under the waveform (the integral), but that the means by which N is derived is exactly analogous to the means by which the integral was originally derived (by adding a number of small areas).

To obtain the units of millivolt-seconds from the reading a factor must be applied to N as follows:

$$\frac{\text{full-scale millivolts (range)}}{10^5 \text{ (full-scale pulses per second)}}$$

In the example given, we have

$$\begin{aligned} N \times \frac{\text{range}}{10^5} &= 60{,}000\,\frac{1000}{10^5} \\ &= 600 \text{ millivolt-seconds} \end{aligned}$$

# 7

# Special-Purpose Analog Meters

This chapter is devoted to special-purpose analog meters that are typical of the laboratory instruments now in use.

## MODEL 202B D-C MICROVOLTMETER

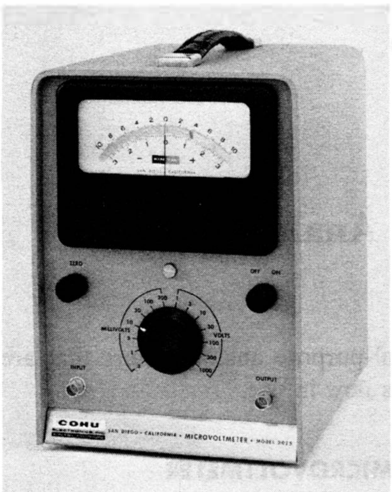
The first instrument selected for discussion is manufactured by Cohu Electronics, Inc., San Diego, California, and is designated as their Model 202B. The instrument is a combination d-c microvoltmeter and amplifier, providing a wide range of measurement and amplification. Fourteen voltage-ranges cover from 300 microvolts full scale to 1000 volts full scale. The zero-center meter face indicates polarity on a mirrored scale which covers all ranges. The amplifier has a maximum gain of 70 db to an output connector on the front panel, providing for its use as a low-drift d-c amplifier with high gain and very high input impedance. A photograph of the unit is shown in Fig. 7-1.

### Theory of Operation

A schematic (as well as an overall block diagram) of the Model 202B is shown in Fig. 7-2.

*Power Supply Section*—The power supply section consists of a transformer with four secondaries and a fused primary isolated from the a-c line by an r-f choke and bypass capacitors. The third wire on the a-c line connects the chassis common ground to earth ground.

The top secondary winding of the transformer provides for properly phased alternating current to drive diodes CR1 and CR2 at the output of the chopper amplifier. The next secondary winding supplies a high-voltage, full-wave rectifier the output of which is connected across two series-connected voltage regulator tubes. The point to which the two tubes are connected is grounded. This divides the




Courtesy Cohu Electronics, Inc.

**Fig. 7-1. Cohu Electronics Model 202B microvoltmeter.**

voltage to provide −105 volts and +150 volts. The third winding supplies a-c heater voltage to all tubes (except V1), pilot-light voltage, and properly phased energizing voltage to the chopper. The bottom secondary winding supplies heater voltage which is rectified by two 1N91 diodes, supplying direct current to the heater of the first a-c amplifier tube (V1). This minimizes hum pickup in this low-level amplifier.

*The Amplifier System*—The amplifier system actually consists of two amplifiers: a chopper amplifier and a d-c amplifier. The chopper amplifier is on the input side and feeds the d-c amplifier on the output side. The chopper amplifier is more stable for amplifying low-level direct current than a straight d-c amplifier.

To prevent oscillation due to feedback loops created by minute voltage differentials between various chassis areas, all grounds in the amplifier are connected to a circuit common ground at the input; they are also brought out to an external earth ground through a third lug on the a-c power plug.

Since the microvoltmeter must indicate the magnitude of extremely small d-c signals, great amplification is necessary before the signal can be used to actuate a conventional d-c microammeter movement. Gain can be rigidly fixed by the use of a large value of feedback, thus adapting an electronic amplifier for measurement purposes. A conventional direct-coupled d-c amplifier cannot be used, however, because of the trouble encountered with drift and grid currents. If grid current is allowed into the input system, an apparent input signal is developed which results in an erroneous output. In the conventional d-c amplifier, drift may also be caused by changes in the direct current through the tubes. Such changes can be caused by small variations in heater voltage or by normal tube aging, resulting in a deviation in the emission characteristics of the tubes.

These difficulties may be overcome by the chopper amplifier, since it reduces the effect of grid current in the input circuit and plate-current drift to a negligible amount by using coupling capacitors.

Fig. 7-3 shows how the effect of grid currents and drifting plate current would give an erroneous output in a conventional d-c amplifier with very small d-c input and how these effects are overcome with the chopper amplifier. Part A shows greatly magnified plate-voltage drift in a d-c amplifier with the grid shorted to cathode. This plate-voltage drift is due to changes in plate current. Part B shows the same voltage drift, but at a different level, due to current through the grid resistance, thus changing the bias.

Part C shows plate voltages with a small positive d-c voltage applied at the grid. Plate voltage drifts as before, but again at a different level. Note that the plate-voltage change caused by the small input signal is on the same order of magnitude as the changes caused by grid current and plate-current drift. Unless a much larger input signal is used, d-c plate voltage change will not depend on signal-voltage change, and the amplifier will give erroneous readings.

Part D shows plate voltage with small d-c voltage being switched on and off at the grid. Note that while drift is still apparent, the rate of change of the square-wave component is constant.

Part E shows the voltage of part D as it would appear at the grid of the following stage in a capacitive-coupled amplifier. Since the coupling capacitors have blocked the drifting d-c component, all that remains is for the amplified input square wave to be proportional to the small d-c input. Long-term drift in the system due to thermal ef-

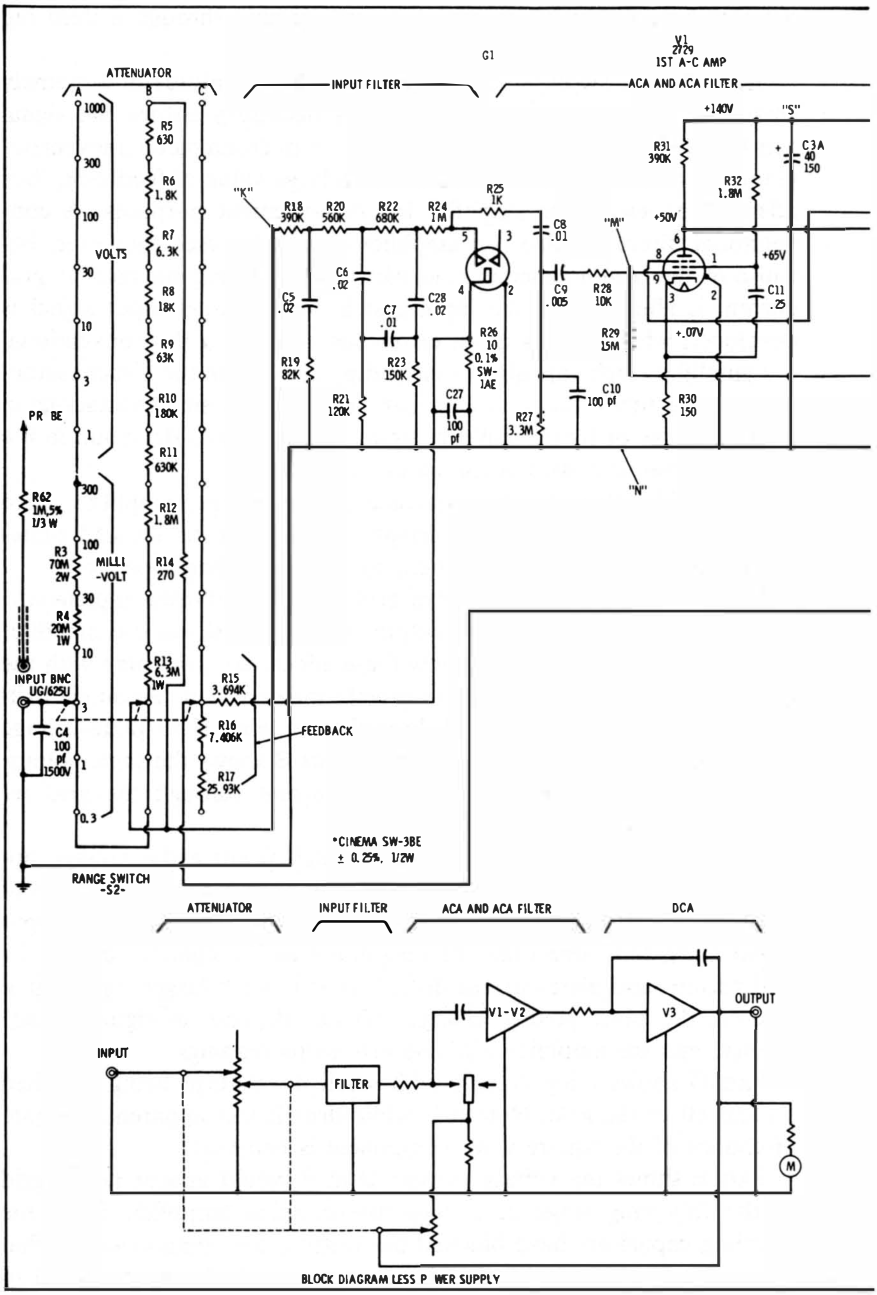
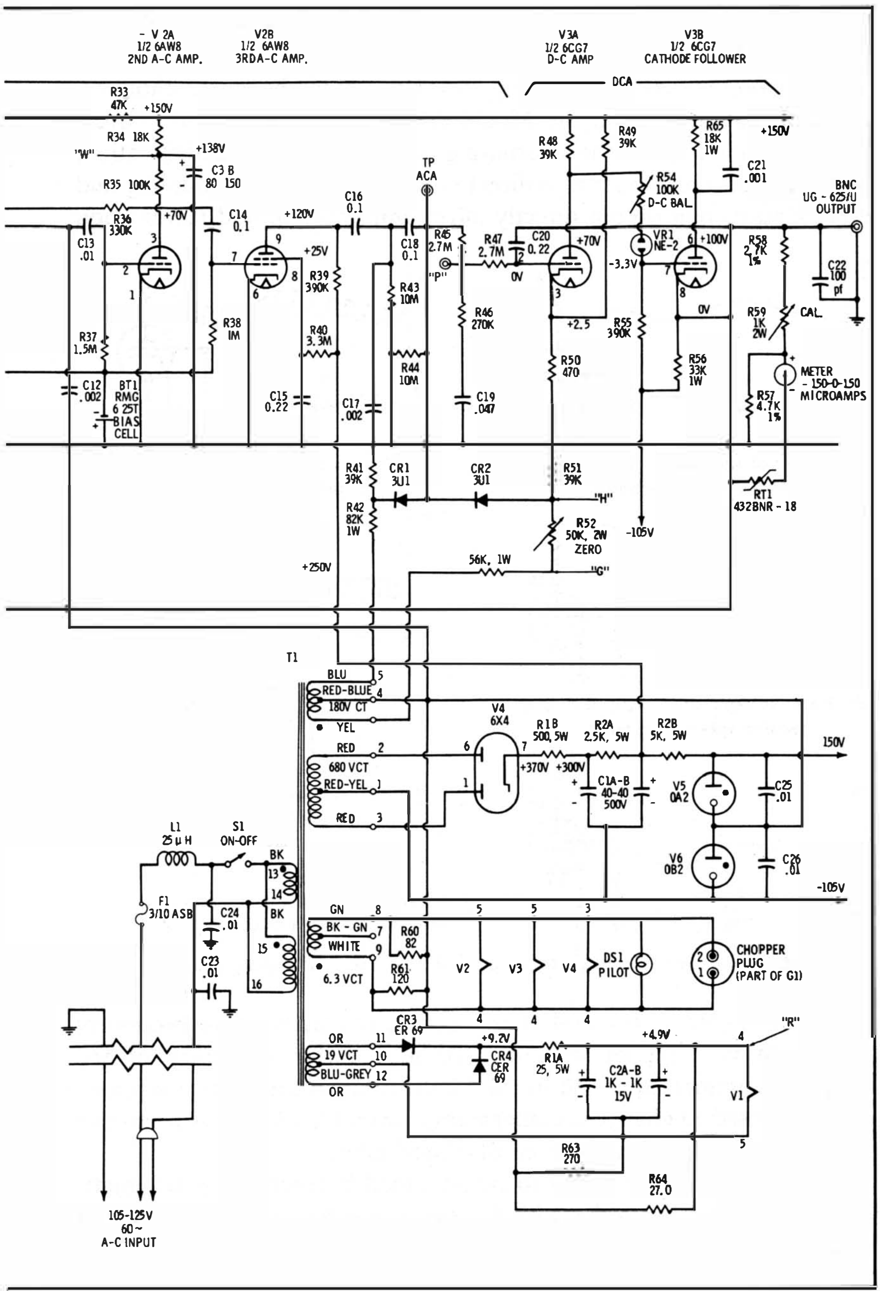


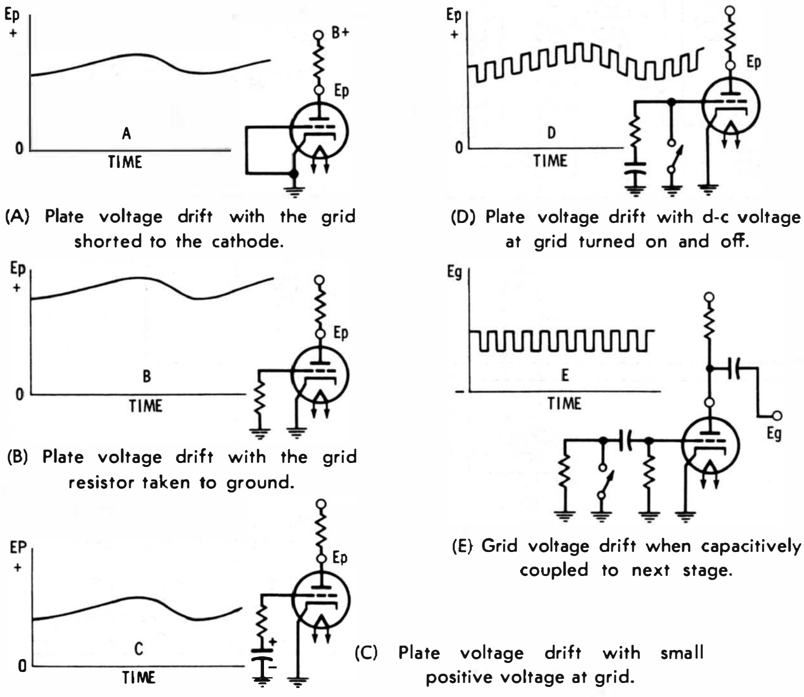

**Fig. 7-2. Schematic**



of Model 202B.

fects and circuit aging can be compensated when necessary by the use of a zero adjustment.

*The Chopper Amplifier*—Fig. 7-4 represents the basic chopper-amplifier circuit used in the Model 202B. Its purpose is to convert low-level direct current to alternating current for stable amplification, amplify it, and reconvert it to direct current. The plate decoupling and other circuits that do not directly affect amplification and rectification




(A) Plate voltage drift with the grid shorted to the cathode.

(D) Plate voltage drift with d-c voltage at grid turned on and off.

(B) Plate voltage drift with the grid resistor taken to ground.

(E) Grid voltage drift when capacitively coupled to next stage.

(C) Plate voltage drift with small positive voltage at grid.

**Fig. 7-3. Waveforms showing effects of grid currents and drifting plate current.**

are not included in Fig. 7-4. For the sake of clarity a synchronously driven output chopper will be used to describe operation of the chopper amplifier, although in the Model 202B a driven crystal-diode system is used to change the alternating current back to direct current after amplification. This will be discussed later.

In Fig. 7-4 the d-c signal to be amplified is represented schematically by a battery in series with a resistor, connected so as to provide a positive input signal. There are two possible positions for the chopper poles at the input and output of the amplifier: poles left and poles right.

With the poles left the input voltage is grounded by the input chopper; the input to the first stage is zero. With no input voltage and assuming a quiet-state plate voltage of 100 volts for the last stage, there is zero output with the output chopper-pole left because the junction of C2 and R6 is grounded and C2 is charged through R5 to the plate voltage ($E_p$) of 100 volts.

When the input chopper pole goes right, an input of the polarity shown in Fig. 7-4 is applied through C1 to the grid of the first stage.

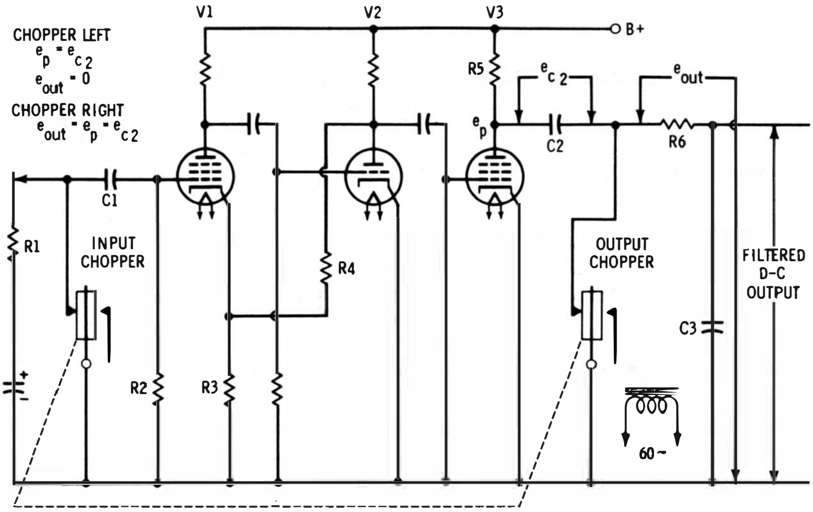



**Fig. 7-4. Schematic of basic chopper-amplifier circuit.**

This results in a decrease in the plate voltage of the first stage, an increase in the plate voltage in the second stage, and a decrease in the plate voltage of the third stage.

Assume that the plate voltage of the last stage drops to 90 volts. When the output chopper pole is thrown to the right C2 is no longer connected to ground. An output voltage is produced which is equal to the voltage that is now on the plate, minus the voltage level to which C2 was charged just prior to the chopper pole being thrown right. If C2 is charged to a nominal 100 volts and the plate voltage drops to 90 volts because of the positive input signal, the output is then ideally a square wave equal to $-10$ volts.

Although the input signal is positive with respect to ground, this d-c level is lost through capacitive coupling and except for an initial transient condition the plate voltage of the last stage will vary both above and below the quiet state d-c value. Fig. 7-5 shows input and plate waveforms as they would appear initially and as they would appear after several seconds.

The voltage variation appearing at the plate of V3 would, after pasing through C2, vary equally above and below ground potential, giving an average d-c output of zero without further treatment. To obtain a significant d-c output at the junction of C2 and R6, an output chopper must be used. Fig. 7-6 shows the effect of the output chopper on the output waveform at the junction of C2 and R6.

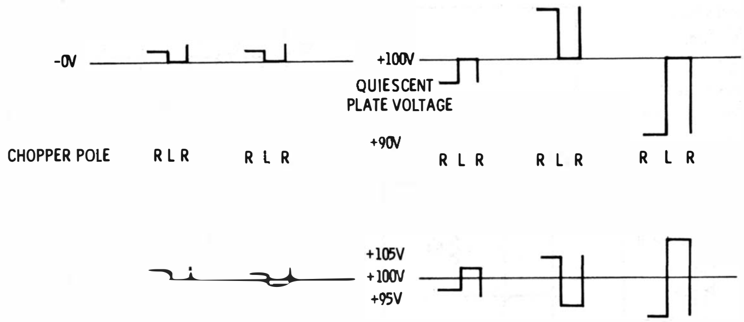



**Fig. 7-5. Sketches of input and plate waveforms.**

Fig. 7-6 shows that the output chopper serves to clamp the top of the output square wave with reference to ground. This is accomplished as follows:

When the chopper pole goes left the plate voltage of the last stage will rise to +105 volts and C2 will charge to +105 volts, being connected between plate and ground. The output voltage will be zero because the junction of C2 and R6 is grounded. When the chopper poles go right again, the plate voltage of V3 will decrease from +105 volts to +95 volts. Since C2 has no quick discharge path, the output voltage which was zero will drop to −10 volts. Thus the output voltage is again equal to the plate voltage present when the output chopper pole broke away from the left contact.

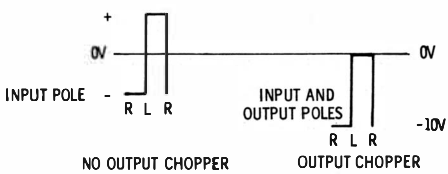



**Fig. 7-6. Sketch of output chopper effect on output waveforms.**

The voltage across C2 closely follows the plate voltage of V3. This is true since C2 has a relatively fast charge path through R5 and the chopper and a relatively slow discharge path through R6, compared with the half-period of the chopper frequency. Because of this, C2 is unable to discharge to the point that the gain would be affected. It thus becomes apparent that by means of the chopper amplifier a pulsating d-c output is obtained which is proportional to, but larger than, the d-c input.

Note that the d-c voltage gain of the chopper amplifier is less than the a-c voltage gain. If the negative pulsating d-c voltage appearing at the junction of R6 and C2 varies from zero to $-10$ volts, then its average d-c value is $-5$ volts. Thus, in this case the d-c voltage gain is half the a-c voltage gain. This is the minimum loss possible. In a practical three-stage chopper amplifier a d-c input to d-c output gain of 10 to 12 db below the a-c gain is to be expected.

The output of the chopper amplifier is filtered by a network designated as R6 and C3 in Fig. 7-4. The time constant of the filter is long compared with a half-period of chopper frequency. The filter actually used in the Model 202B is more complex. In the Model 202B with no signal applied and with the a-c amplifier balanced, there is no output at the filter. With maximum plus d-c signal applied for any range there is a negative d-c output of between 0.1 and 0.25 volt, depending on the amplifier load. The a-c sawtooth ripple is approximately 0.7 volt when measured with the usual averaging-type voltmeter.

Note that in the three-stage, a-c chopper amplifier just described, a phase reversal has taken place in such a way that a positive input voltage produces a negative output voltage. However, the d-c amplifier which follows produces an additional phase reversal for an overall positive gain with the polarities described. Had the input to the chopper amplifier been negative instead of positive, the direction of all waveforms would have been reversed. The plate voltage on the output stage would have been higher than the voltage of C2 during the period when the output chopper was to the right.

Therefore, the chopper output voltage, which is equal to the plate voltage of V3 minus the voltage on C2 at the time the output chopper switches from left to right, would have been positive. Again, the d-c amplifier would give an additional phase reversal such that a negative input voltage would give a negative indication on the meter. Thus it is seen that with a d-c voltage applied to the chopper amplifier described, a larger inversely proportional voltage appears at the chopper output. This, when further amplified by the following d-c amplifier, gives a d-c output directly proportional to the d-c input.

*Driven Diode Chopper*—Because of certain design considerations a driven diode system containing 3U1 diodes is used in the Model 202B, instead of the synchronous output chopper, to change a-c voltage back to d-c voltage after amplification. The method of utilizing diodes to replace the output chopper is shown in Fig. 7-7. The function of the diodes is similar to that of the output chopper.

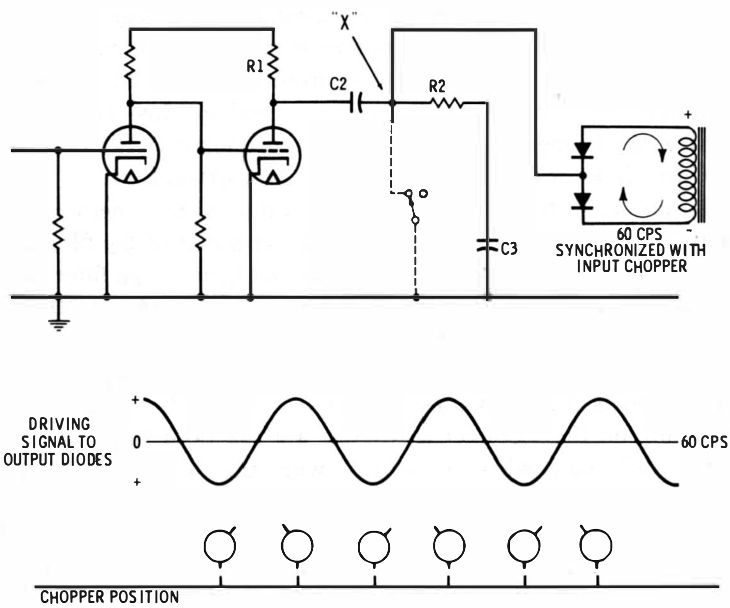



**Fig. 7-7. Schematic of driven-diode chopper circuit.**

When the polarity of the driving voltage is as indicated in Fig. 7-7, maximum conduction takes place through the diodes and point X is essentially at ground. When the polarity is the reverse of that indicated, point X is isolated from ground.

By proper phasing of the diode a-c supply voltage and the a-c voltage which drives the input chopper, the diodes are in effect synchronized with the input chopper. By using sufficient driving voltage across the diodes, plate resistance is made much lower than would be the case with more conventional rectifiers, and a linear relationship between applied a-c signals and output d-c pulses is obtained throughout the range of the applied voltage. Also, inversely proportional d-c output is obtained, regardless of the polarity of the amplifier d-c input voltage.

*The D-C Amplifier*—As has been discussed, the purpose of the chopper amplifier is to prevent drift and other undesired effects due to grid current, component faults, voltage changes, etc. However, after a number of chopper stages have brought the signal up to a point where the signal to noise ratio is more favorable, circuit economy therefore demands that a direct-coupled d-c amplifier should be employed.

The two-stage d-c amplifier in the Model 202B consists of a triode amplifier and a cathode follower (see Fig. 7-2). A neon bulb is used in the interstage divider to minimize gain loss, since it has low dynamic impedance with a relatively large d-c voltage drop. R54, the d-c balance control, sets the direct-current operating conditions in the amplifier, and the chopper amplifier is not required to furnish a steady bias voltage. The Thyrite resistor across the output and a plate resistor on the cathode follower limit the output voltage and current.

Capacitive feedback is connected between the input and output of the d-c amplifier. The purpose of this feedback is to decrease the necessary size of the filter capacitor, to make system response-time independent of d-c amplifier gain variations, and to minimize the effects of internal disturbances in the d-c amplifier.

### Meter Circuit

The meter is connected across the amplifier output in series with the proper multiplier and shunt resistance (see Fig. 7-2). R59, a screwdriver adjustment, is provided for initial factory calibration. The output connector is for an external indicating device capable of handling $\pm 1$-volt maximum output signals and presenting not less than a 1000-ohm load to the amplifier.

## MODEL 204A ELECTRONIC GALVANOMETER

The next instrument selected for discussion is manufactured by Cohu Electronics, Inc., San Diego, California, and is designated as their Model 204A. The instrument is a combination d-c null detector, microvolt ammeter, and an inverting d-c amplifier. It is functionally equivalent to suspension galvanometers, but the chopper-stabilized, all-transistor circuit provides many added features and numerous advantages over conventional moving-coil and electronic galvanometers. The instrument is insensitive to vibration, shock, micro-

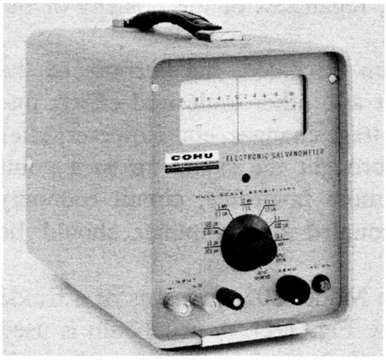



Fig. 7-8. Cohu Electronics Model 204AR electronic galvanometer.

Courtesy Cohu Electronics, Inc.

phonics, earth's magnetic field, and stray pickup. A photograph of the unit is shown in Fig. 7-8.

### Theory of Operation

A block diagram of the Model 204A is shown in Fig. 7-9. The Model 204A consists of an input divider followed by a four-stage a-c amplifier incorporating an input chopper, a synchronous rectifier, and a three-stage d-c output amplifier.

The chopper converts the d-c input into a proportional a-c carrier signal which is amplified by the a-c amplifier. The output of the a-c amplifier is synchronously rectified, filtered, and applied to the d-c output amplifier.

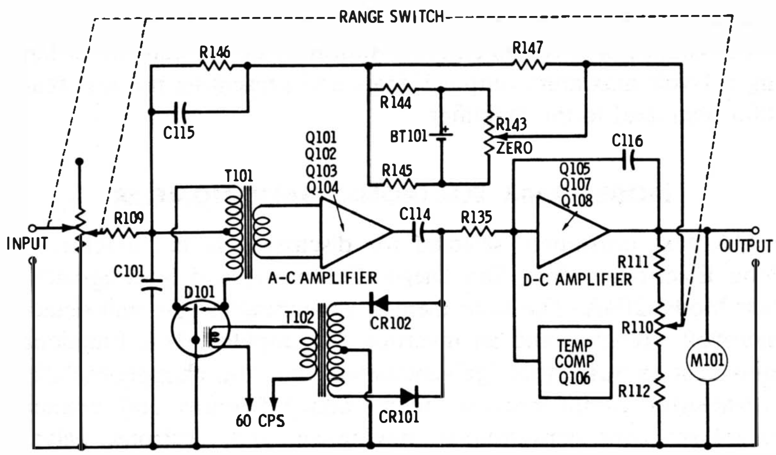



Fig. 7-9. Block diagram of Model 204AR.

Stability and constant input-impedance are maintained by using a large value (approximately 50 db) of inverse feedback. With this value of feedback, overall gain accuracy is determined, essentially by the accuracy of the feedback and input-divider networks, which consist of stable precision resistors. The range switch sets both these networks for the appropriate range. This control functions as an Ayrton shunt, which is usually an accessory for conventional galvanometers.

*Input and Chopper Circuit*—The input signal is fed through R109 to the center tap of the input transformer. R109 sets the d-c gain, either direct or attenuated (depending on the position of the range switch), and forms (in conjunction with C101) a filter. This filter attenuates the 60-cps component that may be present on the signal (due either to leakage or stray fields).

*A-C Amplifier Circuit*—The secondary of the input or chopper transformer feeds the first stage of a four-stage, transistorized, a-c amplifier (Q101 through Q104). The output of this amplifier is coupled to the d-c amplifier.

*Rectifier Circuit*—Diodes CR101 and CR102 are connected to form a half-wave bridge rectifier circuit and are supplied with a 60-cps signal phased to the chopper drive. These diodes rectify the output of the a-c amplifier, and the resultant d-c signals are applied to the input of the d-c amplifier.

*D-C Amplifier Circuit*—The d-c amplifier consists of three transistorized amplification stages (Q105, Q107, and Q108). Temperature compensation is provided by a fourth transistor (Q106), which is connected as a diode. The output of the d-c amplifier is applied to the meter and the output terminals.

*Battery Feedback*—Battery BT101, its associated resistors, and a potentiometer form a bridge which can insert a positive, negative, or zero voltage in series with the feedback resistor. This is effectively equivalent to the insertion of an input zero-adjusting voltage.

# Glossary of Terms

**Accuracy of an indicated or recorded value (percent of reading).** This is expressed by the ratio of the error of the indicated value to the true value. It is usually expressed in percent.

**Accuracy rating of an instrument.** The limits which errors will not exceed when the instrument is used under prescribed conditions.

**Balance.** See *Null balance.*

**Balance detector.** In digital instruments, an electronic amplifier which compares input and feedback voltages and issues commands to the logic circuits to change the feedback voltage as needed to attain a balance.

**Balancing time.** In a d-c dvm ratiometer or ohmmeter, the elapsed time from when the instrument is commanded to make a measurement to when the computation has been completed and is ready for visual readout or automatic recording. In an a-c/d-c converter it is the time elapsed between application of a step input voltage of a stated amplitude and the time when the a-c/d-c converter d-c output has stabilized to within some specified amount of its final value. Balancing time may be expressed as a maximum value or average value for a single range and single polarity or for all range and polarity conditions.

**Conversion, analog servo.** The operating network in which the feedback-voltage generator consists of a motor-driven variable resistor, servo-controlled by an error detector.

**Conversion, prescribed sequence.** (Successive approximation, whole number, or discrete conversion). The operating principle in which the feedback-voltage generator provides a signal of regulated voltage corresponding in value to the input voltage. The voltage corresponding to the most significant digit is first compared with the scaled input voltage. If smaller, it is left in the circuit. The comparison continues with the voltage sources left in the circuit at the end of the process representing the input voltage.

**Conversion, ramp.** (Voltage-to-time conversion). The operating principle in which the feedback-voltage generator periodically produces a voltage which changes linearly with time (ramp voltage) and is compared with the scaled input voltage by an error detector.

**Conversion, voltage-to-frequency.** The operating principle in which a converter generates pulses of frequency directly proportional to the input voltage.

**Current-summing instrument.** One in which the sum of the input current and the feedback current approaches zero at null balance.

**Dead band.** The total range through which the input voltage can be varied without initiating a change in reading. It is usually expressed in digits.

**Decade.** See *Digit position designation.*

**Digit.** One of the symbols of the decimal number system. The term is also applied to the position of symbols in the decimal system. For example, in the number 7851 the term "second digit" identifies the position occupied by the 8.

**Digit position designation.** (Display position). The designation of positions of digits in the display or record is generally in accordance with the following table:

| Number of Digits | Designation (starting at left) |
|---|---|
| 3 | First, second, last |
| 4 | First, second, third, last |
| 5 | First, second, third, fourth, last |

**Digital instrument.** One in which the indicated or recorded value is expressed directly in terms of the decimal number system.

**Dynamic response.** Behavior of the instrument output as a function of the measured quantity when both output and the measured quantity are varying with time.

**Error detector.** See *Balance detector.*

**Extreme operating conditions.** The limits of specified variables or conditions within which the instrument may be operated. Under these conditions, performance ratings do not necessarily apply.

**Feedback voltage or current.** An internally-generated voltage or current used for comparison with the scaled input voltage or current by means of an error detector.

**Input voltage.** (Measured quantity). The voltage to be measured. (See also *Scaled input voltage.*)

**Interference.** Any physical phenomenon tending to adversely affect operation of the instrument.

**Least increment.** The value corresponding to the least digit displayed. It is expressed in volts in a voltmeter, in "per unit" in a ratiometer.

**Logic.** The sequence in which feedback-voltage range-divider operations and polarity-switching operations occur in attaining a balance.

**Logic, double scan** (or "double-duty no-needless-nines" or "tracking scan"). A later improvement of no-needless-nines logic in which the logic does not have to check polarity switch, range switch, and some or all decade switches before changing position of a given decade switch which requires change.

**Logic, no-needless-nines.** A form of scan logic which eliminates needless cycling to 9's and 0's position in each decade of stepping-switch type digital voltmeters. Logic sequence is: set polarity switch, set range switch, set decade switches starting with leftmost and proceeding to the right, changing only those switches whose numerical positions (values) differ from the final numerical reading.

**Logic, scan.** Logic in which each decade of the feedback voltage divider is changed in value only one per measurement and in which these changes progress from any decade (often, but not always, from the leftmost decade) toward the right. See also *Conversion, prescribed sequence.*

**Logic, successive trials.** See *Conversion, prescribed sequence.*

**Logic, tracking.** Logic in which an increase or decrease in any decade requires that all decades to the right of it be switched to 9 or 0, respectively, and in which decade changes start with some decade and progress leftward (one measurement often requires several cycles of skipping again to the rightmost decade and progressing leftward). In this logic, an input-voltage change immediately results in a change in value of the decade which was last changed (in some cases the change will occur in the decade to the right or left of the decade last changed if the decade last changed is at a 9 or 0).

**Longitudinal interference.** Interference which appears between measuring circuit terminals and ground.

**Monotonicity.** The characteristic of a digital instrument such that the readings of the total count appear successively, in proper order, when the input voltage is continuously varied throughout a range.

**Null balance.** The condition which exists in the circuits of an instrument when the absolute value of the error voltage or current is less than that required to initiate changes in the instrument indications.

**Potentiometric instrument.** One in which a feedback voltage is generated by means of a voltage divider.

**Range.** A continuous band of voltage values within which the instrument is capable of making readings to stated accuracy. Digital voltmeters normally provide more than one range by shifting the decimal point.

**Range, basic.** The range in which the input voltage is compared directly with the feedback voltage.

**Range divider.** The circuit element that converts the input voltage to the scaled input voltage.

**Reference voltage, external.** The external voltage source with which the input voltage is compared in digital ratiometers.

**Reference voltage, internal.** The voltage source for the feedback generator in digital voltmeters.

**Resolution.** See *Dead band.*

**Scaled input voltage.** A voltage proportional to the input voltage for comparison with the feedback voltage. The output of the range divider.

**Sensitivity.** See *Dead band.*

**Total count.** The total number of least increments available in a range. For example, in the case of a four-digit instrument, the total count is normally 10,000.

**Transverse interference.** Interference which appears as a voltage between measuring circuit terminals.

# Index

## D



## E



## F



## G



## H



## I



## K



## L

## R



## S



## T



## U



## V



## Z

# DIRECT READOUT METERS

**by John D. Lenk**

The need for trained industrial technicians is expanding rapidly. Increased knowledge about modern circuitry is necessary to keep abreast with industrial advances. This book fills that need by providing information and illustrations to enhance the technician's understanding of the subject of digital, differential, and analog meters.

Written to serve a threefold purpose, the text can be used as a textbook for student technicians, a training aid for the experienced technician desiring to enter the industrial field, and a handy reference for those who are already active in the industry. To fullfil these purposes, a wide variety of laboratory meters is explained in basic terms. Full technician-level circuit descriptions are also provided for the most widely used types of meters. Waveforms at various points throughout the circuits are also discussed.

Laboratory meters are covered in logical progression following signal sequence wherever possible from pickup devices, through the processing circuits, to the display functions. Preceding each group of circuits is a discussion of the basic considerations with typical operating limits outlined for each meter. Numerous illustrations are used throughout the text to enhance your understanding of the subject.

Whether the problem is one of testing, servicing, or simply understanding direct readout meters, this book provides the answers. The necessary information for technicians to qualify as laboratory meter specialists is thoroughly presented in easy-to-understand terms.

## ABOUT THE AUTHOR

John D. Lenk has been a full-time writer since 1949. His numerous books and technical articles have found wide acceptance. In addition, he is a technical writer for industry. His background includes electronics training in the U.S. Navy during World War II. He has held a First-Class Radiotelephone License and an Amateur Radio License since 1939. Other popular Sams books by Mr. Lenk include *Servicing With Dip Meters*, *Servicing UHF TV*, *Eliminating Engine Interference*, and *Electronic Corrosion Control for Boats*.

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