Weston

ENGINEERING NOTES

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A POWER FACTOR METER FOR UNBALANCED POLYPHASE SYSTEMS

CONVENTIONAL polyphase power factor meters measure the phase angle between one of the line currents and two of the line voltages and, therefore, only indicate power factor correctly when the voltages and currents are balanced and are sinusoidal. In 1914, W. N. Goodwin, Jr., and B. P. Romain developed, in the Weston laboratory, a polyphase power factor meter for unbalanced loads. It was based on balanced voltages and sinusoidal current and voltage, but was otherwise correct for any degree of current unbalance. This instrument was not made commercially

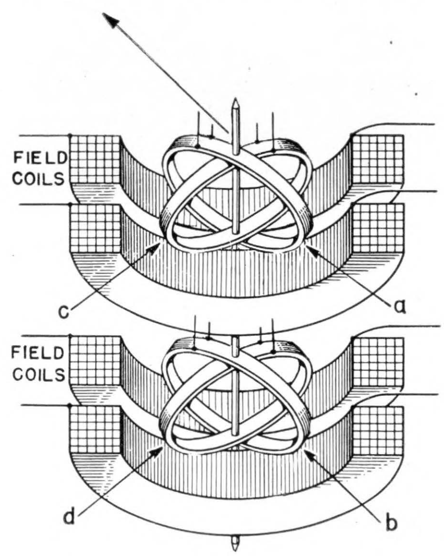

Figure 1

at that time since the only recognized definition of power factor was the ratio of power in watts to the voltamperes, and the instrument measured the ratio of vars to watts which corresponds to the tangent of the power factor angle. These two ratios are only the same for sinusoidal currents and voltages and, therefore, the instrument indicates true power factor only under sine wave conditions.

Since that time, the term vector power factor has become widely used and this is the power factor as derived from the ratio of vars to watts. The instrument is, therefore, a true indicator of vector power factor, and since it is correct for any degree of current unbalance, its indications have real meaning and value. This instrument is now being made as the Weston Model 928, type 2.

Figure 1 shows the basic instrument, which is similar to a two-element wattmeter, but has two moving coils in each element, spaced at an angle of 90 degrees to each other. The field coils and moving coils "a" and "b" are connected to a three-phase, three-wire circuit exactly the same as a two-element wattmeter. The torque of these moving coils is then proportional to the actual power in the circuit.

The other moving coils "c" and "d" are so connected to the circuit that their currents are in quadrature with the currents in the corresponding wattmeter coils "a" and "b." This is the same connection that is used on a two-element varmeter and, therefore, the torque of coils "c" and "d" is proportional to vars. The equilibrium position is where the two torques balance, at which position the ratio of vars to watts is indicated, corresponding to the tangent of the power factor angle.

It is well known that a two-element wattmeter will indicate three-phase, three-wire watts for un-

In This Issue

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balanced as well as balanced currents and voltages, and the conventional two-element varmeter using phase voltage and line current will likewise indicate three-phase, three-wire vars correctly if the current is unbalanced, provided the voltage is balanced. Since the two-element power factor meter is essentially a two-element wattmeter and varmeter, it will indicate correctly for any condition of current unbalance as long as the voltages remain balanced in a three-wire line.

This two-element power factor meter can also be considered as two single-phase power factor meters having a common moving element staff. The single-phase instrument has a resistor in series with one moving coil and a reactor in series with the other moving coil. This provides currents in each coil 90 degrees from each other, which is the same condition that exists in the two-element instrument. The single-phase instrument measures the ratio of vars to watts or vector power factor since the torque on one coil is proportional to watts and the torque on the other is proportional to vars. Since two single-phase wattmeters will indicate total watts when combined into a single two-element wattmeter, and two single-phase varmeters will indicate total vars when combined into a single two-element varmeter, two single-phase power factor meters will indicate total power factor (vector) when combined into a two-element power factor meter.

Goodwin has also made a mathematical analysis and proof of the theory of the two-element power factor meter, and it can be shown mathematically that the instrument will measure vector power factor for the conditions stated previously.

Figure 2 shows the actual internal and external connections of the instrument for three-phase, three-wire circuits. Moving coils "a" and "b" are provided with a center tap and coils "c" and "d" are each connected to the center tap. The other ends of the coils are connected through resistors to the lines and form two artificial Y networks.

Just as in a two-element wattmeter and varmeter, the field coils use line currents I_1 and I_3. In the bottom element, the effective current in the moving coil "a" is in phase with and proportional to line voltage E_{1-2}. This is the same voltage as used for this element of a wattmeter. The current in moving coil "c" is in phase with and proportional to phase voltage E_{0-3}, which is the voltage used for this element of a varmeter. In like manner, the currents in the moving coils of the top element are in phase with and proportional to the wattmeter voltage E_{3-2} and varmeter voltage E_{1-0}.

The resistors connected to "a" and "b" coils are all equal to each other. The resistors in series with "c" and "d" coils are equal to each other and the relationship of these resistors to the others determines the scale range of the instruments.

For three-phase, four-wire circuits the so-called two and one-half element instrument is used. This meter is connected the same as the two and one-half element wattmeter and varmeter and will indicate vector power factor for any condition of current unbalance provided the voltages are balanced. Its theory of operation is identical to the two-element power factor meter.

The conventional single-element power factor meter will even indicate incorrectly for resistive loads if they are unbalanced. For example, assume a delta load having impedances of

5+j0 across lines 1-2.
10+j0 across lines 2-3.
15+j0 across lines 3-1.

Using a conventional power factor meter with its single current coil in the various lines (potential circuits connected with same relationship to current coil in all cases), the following indications are obtained:

With current coil in line 1, indicated power factor is 0.961 lead.
With current coil in line 2, indicated power factor is 0.985 lag.
With current coil in line 3, indicated power factor is 0.993 lag.

The two-element power factor meter indicates unity power factor, which is the true vector power factor as computed.

Other unbalanced loads can cause much larger errors. The following impedances of a delta load have the resistance balanced with the inductance unbalanced in one phase.

5+j5 across lines 1-2.
5+j10 across lines 2-3.
5+j10 across lines 3-1.

For this load condition, the single-element power factor indicates as follows:

With current coil in line 1, indicated power factor is 0.72 lag.
With current coil in line 2, indicated power factor is 0.50 lag.
With current coil in line 3, indicated power factor is 0.45 lag.

The computed vector power factor is 0.57 lag, which is also the reading obtained from the two-element meter.

Changing the load across lines 3-1 in the last example from inductive (5+j10) to capacitive (5−j10) rules out the single-element meter since it will read lead, lag, or off scale, depending upon the location of the current coil. The actual indications are:

0.999 lead with current coil in line 1.
0.50 lag with current coil in line 2.
cosine 180° with current coil in line 3.

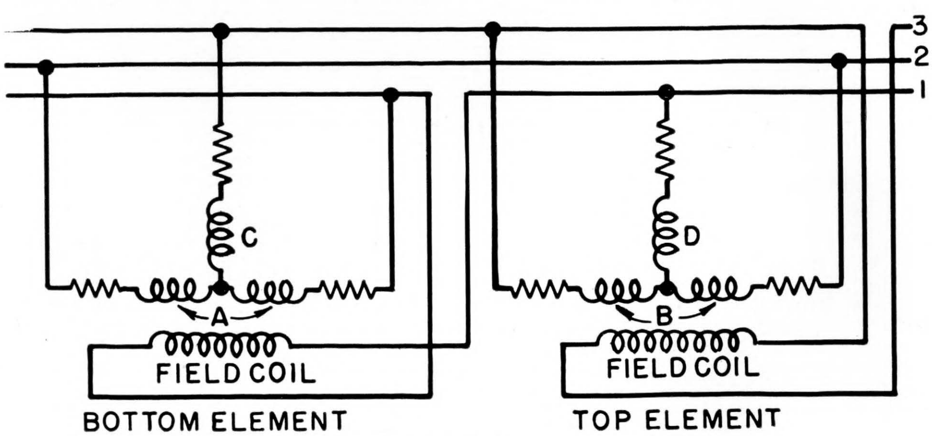


Figure 2

The computed vector power factor is 0.87 lag, which is the power factor indicated by the two-element meter.

This power factor meter is of special value where it is necessary to keep the over-all power factor above a certain value in a three-phase system having a number of single-phase loads connected to the various phases. Loads such as electric furnaces may cause the power factor to decrease below a desired minimum, and when the power factor can be read at a glance, the loads may be changed to raise the power factor. In a system of this type, the currents in each phase may differ widely from each other both in phase and magnitude, making it necessary to use a power factor meter giving a true indication for any load condition.

E. N.—No. 86. —*R. F. Estoppey.*

THERMAL PROBLEMS RELATING TO MEASURING AND CONTROL DEVICES—PART VIII—CONTINUED

Electric Heating of a Conductor, the Resistance of Which Varies With Temperature

INTRODUCTION

In the first installment of Part VIII which appeared in ENGINEERING NOTES, *Volume 6, Number 1, a study was made of the heating of conductors by current, the resistances of which vary with temperature, where the heat not absorbed by the conductors is dissipated to the surrounding medium. In the present installment, consideration will be given to similar conductors but from which the heat dissipated to the surrounding medium is practically negligible relative to the amount generated, as might occur during short circuits, or heavy overloads of relatively short duration.*

24. TO DETERMINE THE INCREASE IN TEMPERATURE OF A CONDUCTOR, THE RESISTANCE OF WHICH VARIES WITH TEMPERATURE, HEATED BY AN ELECTRIC CURRENT, AT ANY TIME AFTER THE INITIAL APPLICATION OF THE HEATING CURRENT, WHEN PRACTICALLY ALL OF THE HEAT GENERATED IS ABSORBED BY THE MATERIAL, AND A RELATIVELY NEGLIGIBLE AMOUNT IS DISSIPATED TO THE SURROUNDING MEDIUM.

It is frequently necessary or desirable to know the temperatures which may be reached in electrical circuits which are subject to the heavy currents produced by short or partial short circuits or definite overloads. Under such conditions, the heat is generated at a rate so large relative to the cooling rate that for practical purposes it may be considered that all of the heat is effective in raising temperature. This will be considered under two conditions: (a) for constant current, and (b) for constant voltage.

24(a). *When the Heating Current Through the Conductor Is Constant.* In the general Equation (161), if the rate of loss of heat, h, to the surrounding medium is made zero, we have,

$$\theta = \frac{1}{a}\left(\epsilon^{+\left(\frac{W_o a}{Ms}\right)t} - 1\right) \tag{171}$$

where $W_o = I^2R_o$, I is the constant current applied, R_o the initial resistance, and a the temperature coefficient of resistance in ohms per ohm per deg. cent.

This equation shows that under the conditions specified for a positive temperature coefficient, the temperature increases indefinitely with time. In practice, of course, the conditions will change with temperature such as for example, a change of state by melting or vaporizing.

Equation (171) can be modified to show the relation between the temperature increase which results when the resistance varies with temperature and that resulting if the resistance were constant, that is $a=0$.

If the resistance remained constant, and all of the heat added were absorbed by the material, then the increase in temperature, θ_i, in the time t would be

$$\theta_i = (W_o t)/(Ms) \tag{172}$$

since $W_o t$ is the total heat added in time t, and Ms is the amount of heat absorbed per degree.

If this relation is substituted in Equation (171), we have for the temperature at any time

$$\theta = \frac{1}{a}(\epsilon^{\theta_i a} - 1) \tag{173}$$

Example:

Assume a copper coil, from which the heat dissipated to the surrounding medium is relatively negligible, heated at such a rate by a constant current that its temperature increase in a given time would be $\theta_i = 100$

deg. cent. if its resistance had been constant. Determine the increase in temperature if the resistance of the coil had the temperature coefficient $a=+0.004$ at the initial temperature, with all other conditions unchanged.

Then from Equation (173), the increase in temperature during this given time would be

$$\theta=\frac{1}{0.004}\left[\epsilon^{+(100\times 0.004)}-1\right]=123 \text{ deg. cent.}$$

That is, under the assumed conditions of time and heating, the effect of the temperature coefficient is to increase this temperature by 23 per cent.

24(b). *When the Conductor Is Heated by Maintaining a Constant Difference of Potential Between Its Terminals.*

If V = difference of potential between terminals of the conductor, held constant.

R = resistance of the conductor at any time and temperature.

$R=R_o(1+a\theta)$.

R_o = initial resistance.

The amount of heat added in the time dt is $(V^2/R)dt=V^2dt/[R_o(1+a\theta)]$ joules.

The amount absorbed by the material in time dt during which the temperature increases $d\theta$ is $Msd\theta$.

Since all of the heat is assumed to be absorbed by the material, then

$$V^2dt/[R_o(1+a\theta)]=Msd\theta \tag{174}$$

Integrating this and remembering that when $t=0$, $\theta=0$, we have

$$\theta=\frac{1}{a}\left(\sqrt{\frac{2V^2at}{R_oMs}+1}-1\right) \tag{175}$$

This may be expressed in terms of the temperature which would result under the same conditions if the resistance remained constant, that is when $a=0$, namely, $V^2t/(R_oMs)$, which let us designate θ_r. Then Equation (175) becomes:

$$\theta=\frac{1}{a}\left(\sqrt{2\theta_ra+1}-1\right) \tag{176}$$

Example:

Assume a device having a copper coil connected directly to a source of excessive but constant voltage and that after some definite time the temperature would have increased 100 deg. cent. had the resistance been constant. Find the increase in temperature which actually occurs during the same time when the temperature coefficient of resistance is +0.004 ohm per ohm per deg. cent.

Then from Equation (176) we have

$$\theta=\frac{1}{0.004}\left[\sqrt{200\times 0.004+1}-1\right]=85.4 \text{ deg. cent.}$$

which shows that under the conditions assumed, the effect of the positive temperature coefficient is to reduce the increase in temperature about 15 per cent from that which it would have had, if the coefficient had been zero.

25. *Derivation of Equation (161) for Constant Current.* The heat generated in the conductor by the constant current I through the resistance at any time R in the time dt is

$$Wdt=I^2Rdt=I^2R_o(1+a\theta)dt \tag{177}$$

For brevity let the rate that heat is generated at the initial temperature, I^2R_o, be designated W_o. Then Equation (177) for the total heat generated becomes $Wdt=W_o(1+a\theta)dt$.

The heat dissipated to the surrounding medium in the time dt is $h\theta dt$, and the heat absorbed by the material during the time dt during which the temperature increases $d\theta$ is $Msd\theta$.

Then, since the heat absorbed by the material plus the heat dissipated to the medium in dt, must be equal to the total heat generated in that time, we have

$$W_o(1+a\theta)dt=h\theta dt+Msd\theta \tag{178}$$

from which we obtain

$$dt=\left(\frac{Ms}{h-W_oa}\right)\frac{d\theta}{\dfrac{W_o}{h-W_oa}-\theta} \tag{179}$$

Integrating Equation (179) under the condition that when $t=0$, $\theta=0$, we have

$$t=\frac{Ms}{h-W_oa}\log\left(\frac{W_o/(h-W_oa)}{W_o/(h-W_oa)-\theta}\right) \tag{180}$$

Putting this into the exponential form it becomes

$$\theta=\frac{W_o}{h-W_oa}\left(1-\epsilon^{-\left(\frac{h-W_oa}{Ms}\right)t}\right) \tag{181}$$

In this Equation, $W_o/(h-W_oa)$ is the final temperature reached, provided h is greater than W_oa, which may be designated θ_r; and $Ms/(h-W_oa)$ has the dimensions of time, and therefore it represents the time constant of the conductor which may be designated by t_r. Then Equation (181) may be written

$$\theta=\theta_r\left(1-\epsilon^{-\frac{t}{t_r}}\right)$$

which is Equation (161).

Derivation of Equation (168) for Constant Voltage:

The heat generated in the conductor having a resistance $R=R_o\ (1+a\theta)$ at any time, during a time dt, when a constant voltage V is applied to its terminals, is

$$\frac{V^2}{R}dt=\frac{V^2dt}{R_o(1+a\theta)}$$

The heat dissipated to the surrounding medium in dt is $h\theta dt$, and the heat absorbed by the material in the time dt, during which the temperature changes $d\theta$, is $Msd\theta$. Then the total heat added in dt is equal to the heat dissipated plus the heat absorbed in that time, that is,

$$W_o\left(\frac{dt}{1+a\theta}\right)=Msd\theta+h\theta dt \qquad (182)$$

where for brevity $W_o=V^2/R_o$, which is the initial rate at which heat is generated.

Separating the variables and putting it into a form for integration, Equation (182) becomes

$$dt=Ms\left[\frac{d\theta}{W_o-h\theta-ha\theta^2}+\frac{a\theta d\theta}{W_o-h\theta-ha\theta^2}\right] \qquad (183)$$

Integrating this equation, and applying the condition that when $t=0$, $\theta=0$, we obtain after simplification,

$$t=\frac{Ms}{2}\left[\frac{1}{h}\log\frac{W_o}{W_o-h\theta-ha\theta^2}+\frac{1}{\sqrt{4W_oha+h^2}}\log\left(\frac{2W_o-h\theta+\theta\sqrt{4W_oha+h^2}}{2W_o-h\theta-\theta\sqrt{4W_oha+h^2}}\right)\right] \qquad (184)$$

This equation can be changed to a simpler and dimensionless form by dividing through by h, and designating W_o/h as θ_m, and Ms/h as t_o, which are, respectively, the final temperature and time constant which would result if the resistance did not vary with changes in temperature, that is if $a=0$. Making these substitutions in Equation (184) we obtain

$$\frac{t}{t_o}=\frac{1}{2}\left[\log\frac{1}{1-\frac{\theta}{\theta_m}-a\theta_m\left(\frac{\theta}{\theta_m}\right)^2}+\frac{1}{\sqrt{1+4a\theta_m}}\log\left\{\frac{\left[1-\frac{1}{2}\left(\frac{\theta}{\theta_m}\right)\right]+\frac{1}{2}\left(\frac{\theta}{\theta_m}\right)\sqrt{1+4a\theta_m}}{\left[1-\frac{1}{2}\left(\frac{\theta}{\theta_m}\right)\right]-\frac{1}{2}\left(\frac{\theta}{\theta_m}\right)\sqrt{1+4a\theta_m}}\right\}\right]$$

which is Equation (168).

Reference:

WESTON ENGINEERING NOTES, Vol. 2, Page 3, 1947.

E. N.—No. 85 Cont. —*W. N. Goodwin, Jr.*

REACTIVE DAMPING OF THE WESTON MODEL 1411 INDUCTRONIC® D-C AMPLIFIER

This is one of a series of applicational articles. The original description of the Model 1411 Inductronic D-C Amplifier appeared in the April 1951 (Vol. 6, No. 1) issue of WESTON ENGINEERING NOTES.

IN SOME applications of the Model 1411 Amplifier, it may be desirable to damp the indication, for example, to average a fluctuating input. Some damping is intrinsic in the indicating instrument, but damping by this means is generally limited to a response time constant of seconds at best, and then at some sacrifice of accuracy in the instrument.

In such cases, the feedback system itself may be damped by including reactance in the feedback network, with no effect upon steady-state accuracy. The system then responds exponentially with a time constant that is a function of the transient response of the complex network. In practice, time constants much larger than obtainable by damping the instrument proper are possible; up to several minutes in some cases.

Method

Theoretically either inductive or capacitive feedback may be used, but for practical purposes a two-winding, iron-cored mutual inductance is best. It is applied by including a primary winding in the output current circuit and a secondary winding in series with the induction converter in the amplifier, in degenerative relation. The resistance of the primary winding should not be larger than 1,000 ohms, and the secondary winding not larger than 50 ohms, to avoid appreciable sensitivity loss. An R-C circuit is included across the primary to by-pass the primary self-inductance which would cause a feedback phase delay and possibly oscillation. This is a precaution rather than a major function.

The feedback system will then respond exponentially, with the indication at time (t) relative to the final indication expressible as:

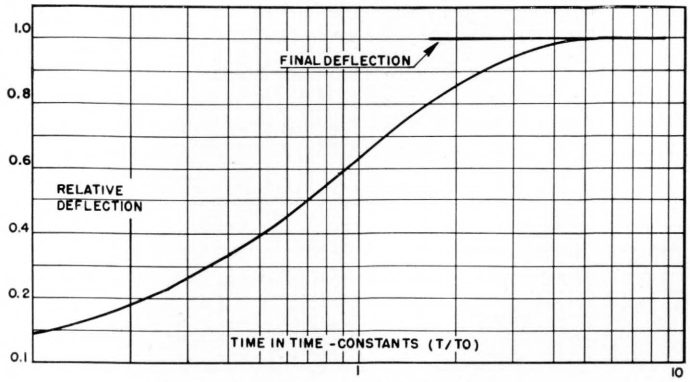


Figure 1—Exponential Response Curve.

$$\frac{\text{indication at time } (t)}{\text{final indication}} = 1 - e^{-t/t_o} \qquad (1)$$

wherein e is the natural log base (2.7---) and t_o is the time constant of the feedback network. The percentage of final indication with time in terms of time constants (t/t_o) is shown in the curve of Figure 1, which is a plot of expression (1).

Potential Circuit

The damping reactor is applied to potential-input circuits as shown in Figure 2. The range-determining resistor is R_1, which may be a mutual pi network in some cases. The steady state calibration relationship for this circuit is:

$$\text{(input) } e = IR_1 \qquad (2)$$

and with the standard output current range of 1 milliampere, the input range in millivolts will equal the value of R_1 in ohms.

The exponential time constant is the ratio of the feedback inductance to the effective feedback resistance, which is:

$$t_o = Lm \left[\frac{R_1 + R_g}{R_1 R_g}\right]\left[\frac{R_s + R_g}{R_g}\right] \text{ (seconds)} \qquad (3)$$

wherein Lm is the mutual coupling inductance in henrys, R_g is the converter shunting resistance, usually 200 ohms, and R_s is the input source circuit resistance. When the source resistance R_s is small with respect to the converter shunt (R_g), the last term may be neglected. And when, as is usually the case, R_g is large with respect to R_1, the first term reduces to $1/R_1$. So for practical purposes:

$$t_o = Lm/R_1 \text{ (seconds)} \qquad (4)$$

Current Circuit

A reactor-damped current-input circuit is shown in Figure 3, which has two range resistors, R_1 and R_2, forming a current ratio network. The steady-state calibration relationship for this circuit is

$$\text{(input) } i = I \frac{R_1}{R_1 + R_2} \qquad (5)$$

The exponential time constant is now:

$$t_o = Lm \left[\frac{R_1 + R_2 + R_g}{R_1 R_g}\right] \times$$

$$\left[\frac{R_g(R_1 + R_2) + R_s(R_1 + R_2 + R_g)}{R_s(R_1 + R_2 + R_g)}\right] \text{ (seconds)} \qquad (6)$$

wherein R_g and R_s are the converter shunt and the source resistance as in the potential case. When the source resistance is large with respect to the other network resistances, the last term may be neglected. And when, as is usually the case, R_1 and R_g are small with respect to R_2, the first term may be reduced. So for practical purposes:

$$t_o = Lm \frac{R_2}{R_1 R_g} \qquad (7)$$

By-Pass Circuit

The reactor primary R-C by-pass circuit is not at all critical, and in many cases the losses in the reactor are

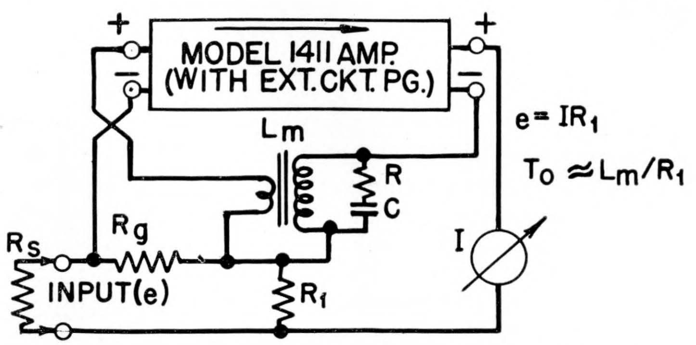


Figure 2—Damping Reactor applied to Potential Input Circuit.

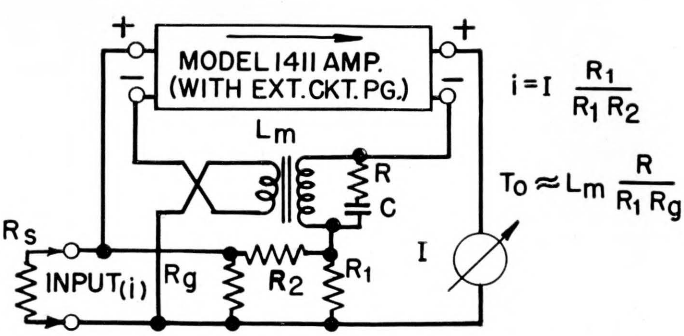


Figure 3—Damping Reactor applied to Current-Input Circuit.

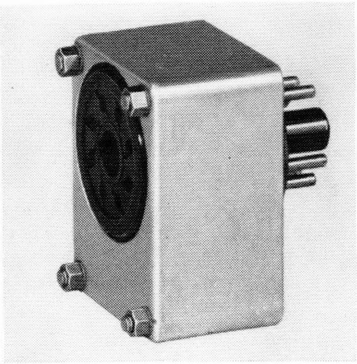


Figure 4—Model 9914—Damping Adaptor.

sufficient to avoid oscillation. But as a precaution, it is recommended.

Ideally, the *R-C* shunt should form a critically damped circuit against the primary self-inductance (Lp) of the reactor, at a resonant frequency (w_o) somewhere between 10 and 100 radians/sec. So:

$$C = \frac{1}{w_o^2 Lp} \tag{8}$$

and

$$R = 2\, w_o Lp \tag{9}$$

It is suggested that an w_o of 50 be selected. C may be a non-polarized electrolytic capacitor, and normal orders of leakage will do no harm. With large reactors where the winding resistance is low with respect to R, the capacitor may be omitted.

Reactors

The only limit to the reduction of response speed obtainable is the physical size of the reactor with regard to tolerable winding resistances. Figure 4 shows a small reactor available as a stock accessory and termed Model 9914 Damping Adapter. It is designed to mount within the Model 1411 Amplifier between the Range Standard and the Range Standard socket. It is sufficient for damping low ranges to a time constant of several seconds.

For greater damping, larger externally connected reactors are required. A typical reactor of standard Weston construction can be obtained for Model 1411 damping to Lm values of 10 henrys maximum.

E. N.—No. 87. —*R. W. Gilbert.*

PHOTOCELL RANGE NETWORK

For Use With the Weston Inductronic® D-C Amplifier

THE Model 1411 Inductronic® D-C Amplifier is ideal for use with Weston dry disc photoelectric cells because it operates the cell under essentially a condition of zero load resistance. Data given previously shows the lower the instrument resistance associated with barrier layer cells, the more linear the response. The photocell terminal potential is automatically balanced to zero, and better linearity and stability are realized than possible with a direct-connected indicating instrument. Also, operation on several ranges without serious mistracking is possible, and the following general design of circuit is suggested.

When a given photocell is used with a Model 1411 Amplifier on a single range, it is usual to match the range standard to the sensitivity of the particular cell being used, and the cell and range standard are paired by number or other means. However, for multirange work, the number of range standards for a given cell becomes burdensome, and damage to the cell would require replacement of all of its associated range standards.

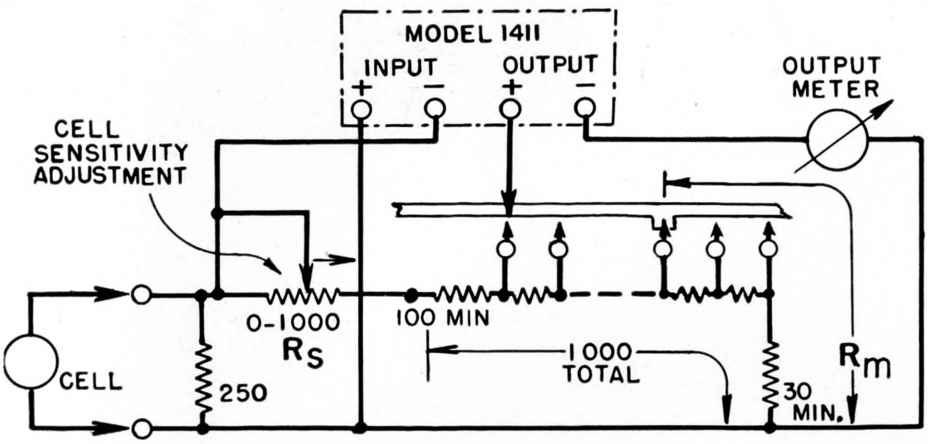
In such cases, it is apparent that switch selection of ranges together with a single-element means for correcting the entire network to specific cell sensitivities is preferred. The element may be either a plug-in resistor supplied with each cell, or preferably a calibrated adjustable element which can be set to the sensitivity of a specific cell. To this end, the following design arrangement is presented.

The Figure illustrates a linear range attenuator for use as the feedback element of the Model 1411 in place of the range standard. The range is a function of the included mutual portion (R_m) of the series of resistors, and the sensitivity adjustment is included as part of the total resistance. The fixed portion of the total resistance, including the mutual resistance, is 1,000 ohms, and the adjustable portion for sensitivity setting (R_s) is 150 to 1,000 ohms. The fixed 1,000-ohm portion is tapped to the range with minimum-range and maximum-range values of mutual resistance of 30 minimum and 900 maximum ohms, respectively, corresponding to 5 and 150 foot-candles, respectively. The relationship between the mutual resistance of each range tap and the range in ft-candles is:

$$R_m = \text{range in ft-candles} \times 6 \text{ (ohms)} \tag{1}$$

from which the series of resistors can be calculated for the required range complement between the limits of $R_m = 30$ ohms (5 ft-candles) and $R_m = 900$ ohms (150 ft-candles). The total of 1,000 ohms is made up by the sum of R_m and the fixed end-portion resistor, which has a minimum value of 100 ohms.

A 250-ohm shunt is included across the input terminals for proper galvanometer shunting resistance. The cell sensitivity adjustment (R_s) can accommodate cell sensitivities from 6 to 3 microamperes/foot-candles corresponding to resistance values of 0 and 1,000



Model 1411—Photocell Network.

R_m to each switch point = range in ft-cd × 6 (ohms)

$$R_s \text{ setting} = \frac{6{,}000}{\text{cell sensitivity } ua/fc} - 1{,}000 \text{ (ohms)}$$

ohms, respectively. The relationship is as follows:

$$R_s = \frac{6{,}000}{\mu a/fc} - 1000 \text{ (ohms)} \qquad (2)$$

or the sensitivity of the cell in $\mu a/fc$ relates to resistances as:

$$\mu a/fc = \frac{6{,}000}{R_s + 1{,}000} \qquad (3)$$

Thus the setting of the cell sensitivity adjustment may be specified from the output of the cell in microamperes/foot-candles (2), or vice versa (3).

In cases where multiple-cell targets or filtered cells falling outside of the 3-6 $\mu a/fc$ level are used, a range multiplying figure may be included and the sensitivity figure restated within the range of 3-6, with the range multiplier expressed as a denominator.

Example: A filtered cell has an output of 0.8 $\mu a/fc$ so that a range multiplier of 5 would be appropriate. The cell output may be stated as 4/5 $\mu a/fc$. The cell sensitivity adjustment from (2) then should be:

$$R_s = \frac{6{,}000}{4} - 1{,}000 = 500 \text{ ohms}$$

and the scale readings multiplied by 5.

A ten-turn 1,000-ohm Micropot or Helipot is ideal for sensitivity setting. This type of resistor can be equipped with a geared dual 10/1 dial figured to 1,000, and with a 1,000-ohm element reads directly in ohms. The suggested system has the advantage of being able to accept and standardize cells of any output sensitivity. It also is continuously adjustable, and can be set by test against a standardizing light source.

E. N.—No. 88. —*R. W. Gilbert.*

WESTON REPRESENTATIVES

ALBANY 7, N. Y.—Carse Electric Corporation.....100 State St.—Albany 3-3628
ATLANTA 3, GA.—E. A. Thornwell, Inc...217 Whitehall St., S.W.—Walnut 3548
BALTIMORE, MD.—Thomas H. Closs, 17 W. Pennsylvania Ave., Towson 4, Md.—TOwson 6335
BOSTON 16, MASS.—Cowperthwait and Brodhead, 131 Clarendon St.—COmmonwealth 6-1825
BUFFALO, N. Y.—Cooper-Morgan, Inc.....................Highland Ave. at Lake St., P. O. Box 152, Hamburg, N. Y.—EMerson 3405-6
CHARLOTTE 2, N. C.—Ranson, Wallace & Company, 116½ E. 4th St.—CHarlotte 4-4244
CHICAGO 6, ILL.—Weston Electrical Instrument Corp. Room 1222, 205 W. Wacker Dr.—FRanklin 2-4656
CINCINNATI 37, OHIO—The Beedle Equipment Co., Roselawn Center Building, Reading and Section Rds.—Jefferson 2640
CLEVELAND 14, OHIO—The Ambos-Jones Co., 1085 The Arcade—CHerry 1-7470
DALLAS 5, TEXAS—Butler and Land.........5538 Dyer St.—EMerson 6-6631
DENVER 16, COLO.—Peterson Company....4949 Colorado Blvd.—DExter 5434
DETROIT 21, MICH.—T. S. Cawthorne Company, 16607 James Couzens Highway—DIamond 1-2121
HOUSTON 2, TEXAS—Lynn Elliott Company...371 M & M Bldg.—CHarter 4715
JACKSONVILLE 4, FLA.—Ward Engineering Co., Inc., 903 West Adams Street—Jacksonville 4-6739
KNOXVILLE 2, TENN.—A. R. Hough, 402 Church St., P. O. Box 1452—KNoxville 2-1383
LITTLE ROCK, ARKANSAS—Curtis H. Stout, 400 Shall St., P. O. Box 107—Little Rock 4-8835
LOS ANGELES 27, CALIF.—Edward S. Sievers, 5171 Hollywood Blvd.—NOrmandy 2-1105
MEMPHIS, TENN.—Curtis H. Stout......730 M & M Bldg.—Memphis 8-7601
MIDDLETOWN, CONN.—Dittman & Greer, Inc., 33 Pleasant St., Middletown 6-9606
MINNEAPOLIS 3, MINN.—Pinkney & Hine, 552-3 Plymouth Bldg.—LIncoln 0523-4
NEWARK 5, N. J.—J. R. Hemion.....614 Frelinghuysen Ave.—BIgelow 3-4700
NEW ORLEANS 12, LA.—W. J. Keller......415 Natchez St.—MAgnolia 3603-4
NEW YORK 7, N. Y.—Weston Electrical Instrument Corp., Room 1775, 50 Church St.—WOrth 4-3510
ORLANDO, FLA.—Ward Engineering Co., Inc., 1217 West Central Ave.—ORlando 2-4295
PHILADELPHIA 2, PA.—Joralemon, Craig & Co., 112 So. 16th St.—RIttenhouse 6-2291-2
PHOENIX, ARIZ.—J. E. Redmond Supply Co., 625 W. Madison St., P. O. Drawer 869—Phoenix 4-8471
PITTSBURGH 22, PA.—Russell F. Clark Co., 1404 Clark Bldg.—ATlantic 1-8089, 1-8090
ROCHESTER 4, N. Y.—Rittenhouse & McGuire Co., 311 Alexander St.—BAker 2340
SAN FRANCISCO 5, CALIF.—Herman E. Held, 120 Main St.—GArfield 1-6130-1, SUtter 1-2652
SEATTLE 4, WASH.—Eicher & Co.......263 Colman Bldg.—MUtual 2411, 2412
ST. LOUIS 1, MO.—C. B. Fall Co...Suite 1003, 317 N. 11th St.—CHestnut 2433-4
SYRACUSE 2, N. Y.—Naylor Electric Company, 317 State Tower Bldg.—SYracuse 2-3894
TULSA 1, OKLA.—Riddle and Hubbell Company, 302 S. Cheyenne Ave., P. O. Box 727—TUlsa 3-4697
WASHINGTON 11, D. C.—Weston Electrical Instrument Corp., 6230 Third St., N. W.—TUckerman 2414
CALGARY, ALTA, CAN.—Northern Electric Co., Ltd...........709 8th Ave., W.
HALIFAX, N. S., CAN.—Northern Electric Co., Ltd.................86 Hollis St.
MONCTON, N. B., CAN.—Northern Electric Co., Ltd..............599 Main St.
MONTREAL, QUE., CAN.—Northern Electric Co., Ltd...1620 Notre Dame St., W.
MONTREAL 25, QUE., CAN.—Powerlite Devices, Ltd...1440 St. Catherine St., W.
OTTAWA, ONT., CAN.—Northern Electric Co., Ltd...........141 Catherine St.
REGINA, SASK., CAN.—Northern Electric Co., Ltd........2300 Dewdney Ave.
TORONTO 9, ONT., CAN.—Powerlite Devices, Ltd.....1870 Davenport Road
TORONTO, ONT., CAN.—Northern Electric Co., Ltd..............143 Fleet St.
VANCOUVER, B. C., CAN.—Northern Electric Co., Ltd...........150 Robson St.
WINNIPEG, MAN., CAN.—Northern Electric Co., Ltd.............65 Rorie St.