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SOME ELECTRONIC ACCOUNTING DEVICES FOR DATA LOGGING APPLICATIONS

S. H. Jury

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CHEMICAL TECHNOLOGY DIVISION

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S. H. Jury

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S. H. Jury

SUMMARY

Several analog accounting circuits have been designed and investigated to determine their applicability to pulse counting in connection with metering of flows in the waste calcination process. A circuit involving an amplifier and comparator with appropriate passive elements has been found most suitable for the accounting application with pulse rates up to about 50 K pps.

Errors due to drift have been reduced to less than one count in 2,000 by appropriate choice of feedback capacitor and input resistor to reduce amplifier drift. Errors due to "learn" time have been investigated analytically and it has been shown that these errors can be controlled through appropriate choice of time constant in relation to the total count involved. These errors, in general, can easily be reduced to less than one count in 2,000.

A number of meter switches have been investigated to determine the effect of "contact bounce" in the production of serrations in a pulse. In general, it has been found that snap action micro-switches are most suitable for accounting purposes. Upon closing, they give rise to 4-8 distinguishable serrations lasting up to about 2-5 ms total time before a clean solid pulse sets in. It has been shown analytically that a serration filter can be used to avoid counting of serrations. It has also been shown experimentally that with high speed counting and in the absence of such a filter the serrations are counted to an extent determined by the time constant of the accounting operation. Errors due to serrations can be eliminated through calibration of the circuit if one wishes to omit the filter. In low speed counting (up to 50 pps), the effect of serrations is made negligible due to the longer interval time constant that may be employed.

1.0 INTRODUCTION

In the "off-on" type of control used with the waste calcination process and perhaps others, the variables being controlled change abruptly with time. Wattmeters, fluid flowrators, etc., register at full speed at one moment and are turned off or stepped down at another. Also the output of the meter is generally a mechanical displacement which must be translated into its voltage equivalent to be compatible with conventional digital data logging equipment. If this translation is made directly, any instantaneous values are considerably less meaningful than those integrated over the period of time elapsed between data loggings or any other period of time longer than the data logging interval.

In view of the foregoing requirements, the various meters involved in the calciner operation have been equipped with a cam operated switch. Each time the meter shaft to which the cam is attached makes one revolution the switch contacts make one closure and are opened again. If one connects this switch into an appropriate electrical circuit, an electrical pulse can be generated for each shaft revolution. By counting the pulses during an interval of time, one obtains an electrical indication of the total or integral flow during that time interval. Thus, the problem is reduced to one of simply counting pulses and generating an output emf proportional to their sum.

The so called "DCAT" has been developed at ORNL to count the pulses and translate the count into an output emf which is fed to the logging equipment. The DCAT involves a series of three stepping switches which are actuated by the pulses through an RCL network. In principle, a pulse causes the first stepping switch to step one step, a second pulse a second step, and so on to the end of the first stepping switch. At the end the first stepping switch resets and in so doing causes the second one to step one step. These cycles are repeated to the end of the second stepping switch at which time it resets and steps the third switch one step. At the end of the third switch, all three switches return to reset and the entire process is ready for repeating.

Resistors which are connected to each step of the switches are cut into a voltage divider network so that an output voltage is obtained which is proportional to the sum of the steps registered. This voltage is then fed to the logging equipment.

Although the stepping switch principle is a sound one for modest counting rates, the DCAT version has not been reliable in the calcination process.

The foregoing experience has suggested the need for improvement and perhaps an independent approach to the problem. It is the purpose of this report to summarize the results of the latter approach.

2.0 CHARACTER OF PULSES

The ideal pulses for accounting purposes are the uniform square ones. This means that the meter switch involved must make "clean" closures and breaks and for this purpose a snap action type without "contact bounce" is to be preferred. If a partial closure with a non-snap type occurs at a time when the meter stops, it is quite likely that intermittent contacting would produce spurious pulses which, if of long enough duration, would increase the registered count compared to the actual. It is quite unlikely, however, that it would reduce the count as has been observed unless the intermittent events were associated with some other less obvious one.

The pulses from the meters are non-ideal in the sense that the pulse width decreases with increasing cam speed. In fact, the total switch closure time at one cam speed is the same as at any other cam speed. If, for example, the cam speed is doubled, the dwell time is halved.

3.0 ACCOUNTING CIRCUIT BASED ON ANALOG MEMORY

In high speed iterative analog computation, it is necessary to keep track of the number of the iterative step being computed. For this purpose the circuit of Fig. 1 is in common use. It is this circuit which is also applicable to the accounting problem at hand.

Basically, the circuit involves two integrators labeled R-M and M in Fig. 1. Each integrator involves a high gain chopper stabilized DC operational amplifier having an amplification factor, $-\mu$, of the order of -10^8. The C μf capacitors are the feedback capacitors. The fixed resistors attached directly to the output of the amplifiers along with the B comparator form the initial condition (I.C.) circuits. The buffer amplifier of gain +1 may for the moment be considered to be shunted because it is not pertinent to the immediate discussion of basic principles of operation.

It is important to note that no summing networks are shown at the inputs to the integrators in Fig. 1. Consequently, the integrators can only hold, i.e. remember, or track their I.C. input as determined by whether the terminals A-A and B-B are disconnected or connected. Thus, since the integrator circuits are not being used to integrate but to remember, we will refer to them as memories.

The difference in labeling of the memories is associated with the meter switch and B comparator hookup. When the meter switch is open only the y volts are applied to the B comparator coil. Consequently the B comparator blades are in the minus position and terminals A-A are connected while the B-B terminals are disconnected. In other words, the "reverse memory" abbreviated R-M is tracking its I.C. voltage, u + z volts, and the "forward memory" abbreviated M is holding, i.e. remembering. When the meter switch closes, the sum of +x -y volts are applied to the B comparator coil. Since $|x| > |y|$ the B comparator blades flop to the + position and now the R-M holds and the M tracks.

While the meter is running, the pulses of x volts repeat the foregoing events and if one plotted out the Z voltage he would observe that Z increases according to the step plot labeled M in Fig. 2. The absolute value of the R-M output voltage is also shown as the step plot labeled R-M in Fig. 2. The height of each voltage step is u volts.

If Z were allowed to increase indefinitely, the memories would of course overload were it not for the A comparator which resets the R-M to -w +u volts when Z reaches the voltage $|v|$ volts. This accounts for the abrupt drop (reset) in Z at the top of the steps shown in Fig. 2. After this reset operation, the accounting procedure starts over again.

The divided voltage z instead of Z is fed to the digital data logging equipment because z = $\pm$ 50 mv will cause it to operate full scale whereas operational amplifiers operate in the ranges Z = $\pm$ 10, $\pm$ 50, $\pm$ 100 volts depending on the brand of operational amplifiers used.

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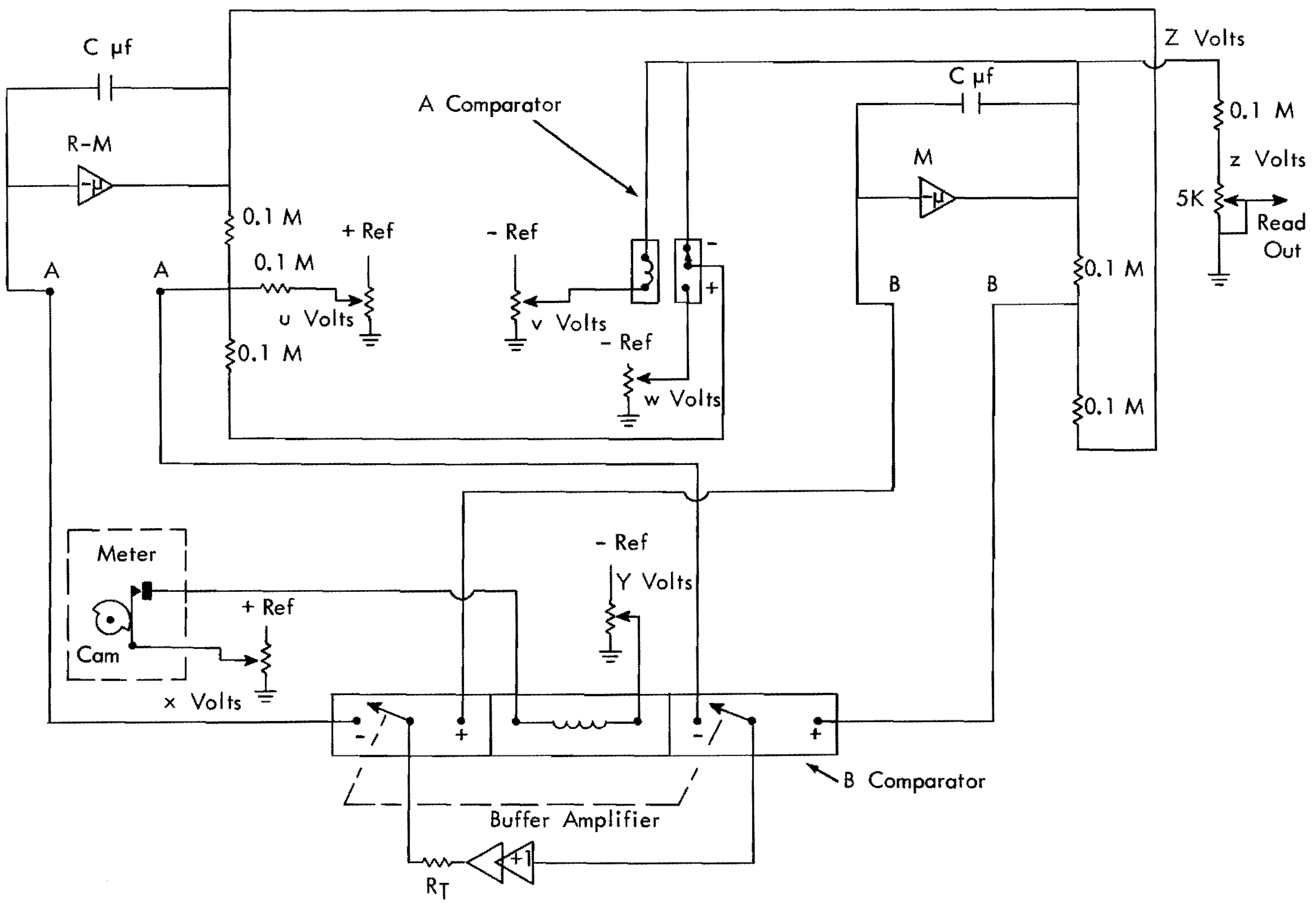

Fig. 1. An electronic counting device.

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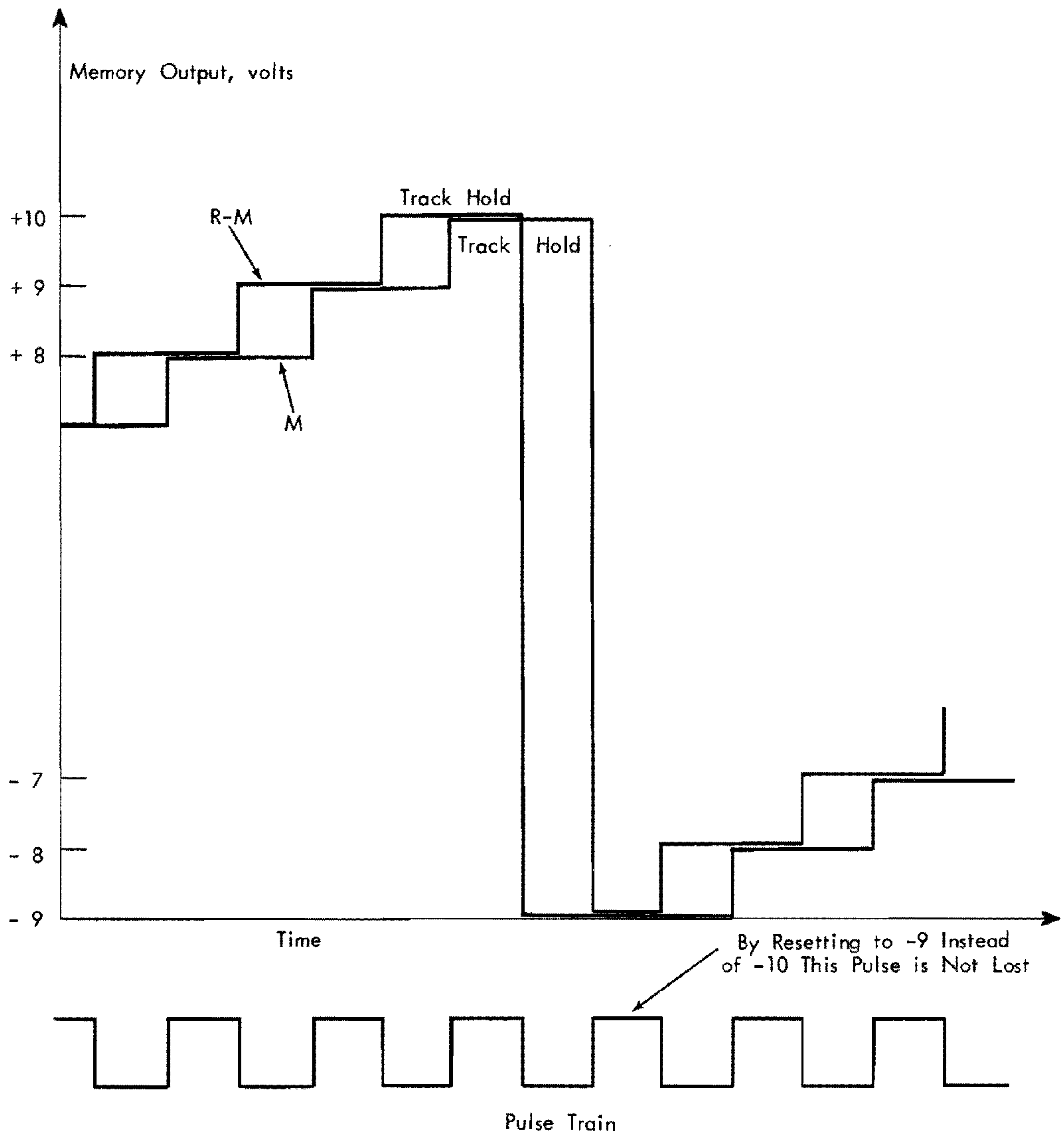

Fig. 2. The memory output voltage histories.

With $\pm$ 100 volt amplifiers and u = 0.1 volt, for example, it is possible to count 2,000 steps before resetting the memories.

4.0 ANALYSIS OF THE UNASSISTED TRACKING OPERATION

For simplicity sake, the risers of the steps in Fig. 2 have been shown as vertical. This in fact is not actually the case. When either terminals A-A or B-B are connected, it takes time to charge the capacitors to their new voltages because infinite currents are impossible. In fact, it is this very consideration which contributes to the determination of the minimum pulse width and pulse spacing that can be accommodated without loss of information. Thus, it takes time for a memory to be "taught," i.e. "learn," a new input voltage value.

The transient analysis of the unassisted (no buffer amplifier) learning operation can best be carried out with reference to the simplified diagram of a single memory shown in Fig. 3. The Kirchoff nodal equation for current through the summing junction with switch S closed is

$$\frac{e_i}{R} + \frac{e_o}{R} + C\,\frac{de_o}{dt} = 0 \tag{1}$$

the Laplace transform of which is

$$\bar{e}_i + \bar{e}_o + RCS\left[\bar{e}_o - e_o(0)\right] = 0 \tag{2}$$

Thus the transfer function of the tracking operation is

$$-\frac{\bar{e}_o}{\bar{e}_i} = \frac{1}{RCS + 1} \tag{3}$$

where RC is the time constant of the learning operation. The inverse transform of eq. 3 for constant e_i is

$$-e_o = e_i\left(1 - e^{-t/RC}\right) \tag{4}$$

Only when $t = \infty$ does

$$-e_o \equiv e_i$$

For practical purposes, however, one takes

$$t = mRC \tag{5}$$

where m is a number chosen such that learning is adequate for satisfying the learning specifications at hand. For example, if m = 10 then

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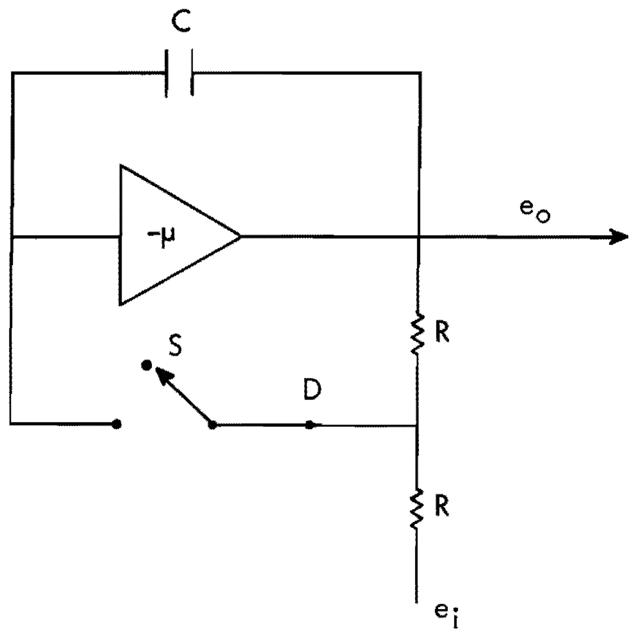

Fig. 3. Simplified diagram of a single memory.

$$1 - e^{-10} = 0.99995 \tag{6}$$

which is the reason why 10 time constants are frequently used for reset and compute intervals in high speed repetitive analog computation.

5.0 ANALYSIS OF THE CRITERION, m

Based on the previous section, it is obvious that one ought to have available some sound mathematical basis for determining m, the minimum number of time constants that must be provided for in a given design. For this purpose, we will use the black box diagram shown in Fig. 4 where we define

$$f = 1 - e^{-m} \tag{7}$$

= fraction learned of the voltage being taught during the tracking time $t = mRC$

and it is assumed that in Fig. 1 the cam closes the meter switch for one half revolution and opens it for the other half. In other words, while the meter runs at a given RPM the pulses are spaced in time equal to their width. From Fig. 2, it is obvious that during 1 RPM there occurs one cycle consisting of a pulse and pulse space during which each memory has experienced one learn period. Also during this cycle which we will call the n^{th} one, the voltage x_n which appeared at point A was added to the constant voltage u and $f(u + x_n)$ volts were learned at the output of the R-M before the R-M switched from track to hold and the M started tracking. When M stopped tracking it learned $f^2(u + x_n)$ volts of the original $u + x_n$. It is the $f^2(u + x_n)$ voltage which now appears at point A preparatory for the n + 1 cycle. Thus,

$$x_{n+1} = f^2(x_n + u) \tag{8}$$

When this finite difference equation is solved subject to the condition

$$x_{n=0} = -w \text{ volts} \tag{9}$$

we find that

$$x_n = - \left[|w| + \frac{f^2 u}{1-f^2} \right] f^2 n + \frac{f^2 u}{1-f^2} \tag{10}$$

or

$$x_n = - |w| f^{2n} + f^2 u \frac{1-f^{2n}}{1-f^2} \tag{11}$$

If f = 1, then

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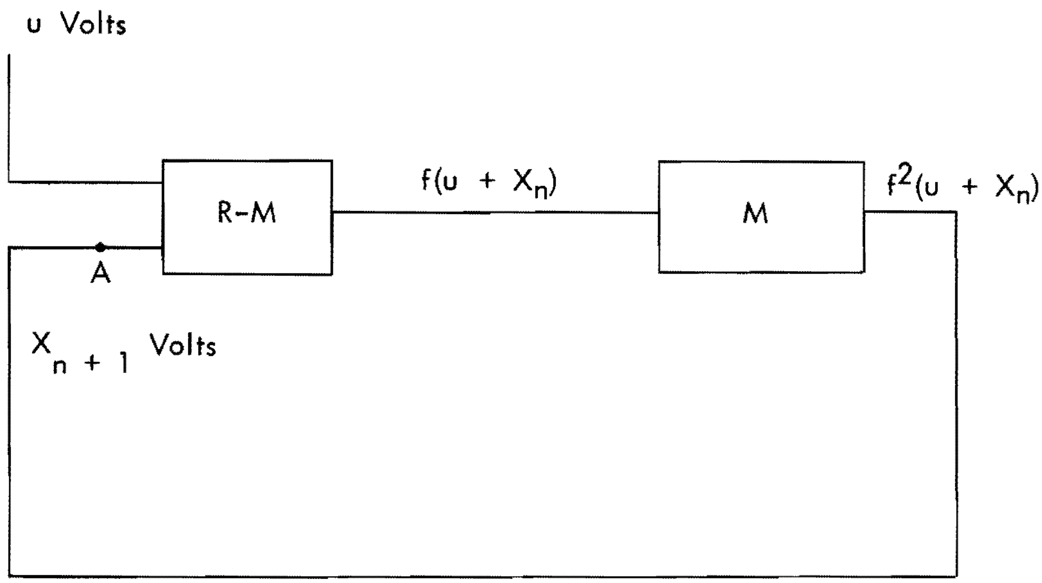

Fig. 4. Black box diagram of the accounting operations.

$$(x_n)_{f=1} = - |w| + nu \tag{12}$$

At the end of the n + 1 cycle we have

$$F_{n+1} = \frac{x_{n+1} + |w|}{(x_{n+1})_{f=1} + |w|} = \frac{\left[\frac{|w|}{u} + \frac{f^2}{1-f^2}\right]\left[1 - f^{2(n+1)}\right]}{n + 1} \tag{13}$$

where

F_{n+1} = ratio of the apparent number of cycles (or pulses) based on output meter reading to the actual number of cycles (or pulses) that transpired.

For a large number of pulses as we are concerned with here, F_{n+1} is not as important as the actual number of counts lost from start to reset. If, for example,

N = total number of counts from start to reset

$$= 2 \frac{|w|}{u}$$

then eq. 13 may be written

$$F_{N+1} = \frac{\left[\frac{N}{2} + \frac{f^2}{1-f^2}\right]\left[1 - f^{2(N+1)}\right]}{N + 1} \tag{14}$$

Now if we specify that we do not want to lose more than one count out of N, then based on eq. 14 we have

$$N = \left[\frac{N}{2} + \frac{f^2}{1-f^2}\right]\left[1 - f^{2(N+1)}\right] \tag{15}$$

This equation is cumbersome to evaluate as it stands due to the large N and relatively large m involved. For this reason we use eq. 7 to approximate as follows:

$$\frac{f^2}{1-f^2} = \frac{(1 - e^{-m})^2}{(2 - e^{-m})e^{-m}} \simeq \frac{e^m}{2} \tag{16}$$

Thus eq. 15 may be written

$$1 - 2\frac{N}{N + e^m} = \frac{1 - \frac{N}{e^m}}{1 + \frac{N}{e^m}} = (1 - e^{-m})^{2(N+1)} \tag{17}$$

so that

$$\ln \frac{1 - \frac{N}{e^m}}{1 + \frac{N}{e^m}} = 2(N + 1) \ln (1 - e^{-m}) \qquad (18)$$

Now

$$\ln \frac{1 + x}{1 - x} = 2 \left[x + \frac{x^3}{3} + \frac{x^5}{5} + \ldots \right] \; ; \; x^2 < 1 \qquad (19)$$

$$\ln (1 - x) = - \left[x + \frac{x^2}{2} + \frac{x^3}{3} + \ldots \right] \; ; \; -1 < x < 1 \qquad (20)$$

so that

$$\left[\frac{N}{e^m} + \frac{\left(\frac{N}{e^m}\right)^3}{3} + \frac{\left(\frac{N}{e^m}\right)^5}{5} + \ldots \right] = (N + 1) \left[e^{-m} + \frac{e^{-2m}}{2} + \frac{e^{-3m}}{3} + \ldots \right] \qquad (21)$$

Even for the ridiculously low value

$$m = 2$$

we have

$$e^{-m} = 0.135$$

and the first term of the right hand series approximates the series within about 5 per cent. Thus eq. 21 reduces to

$$\frac{\left(\frac{N}{e^m}\right)^2}{3} \left[1 + \frac{3\left(\frac{N}{e^m}\right)^2}{5} + \ldots \right] = \frac{1}{N} \qquad (22)$$

Now if for the moment we assume the answer namely that N/e^m is small so that we need use only the first term of the left hand side, then

$$m = \frac{2.303}{2} \left[3 \log N - \log 3 \right] \qquad (23)$$

which is plotted in Fig. 5. Now from the plot and a table of exponentials, we can determine, for example, that

$$\frac{N}{e^m} = \frac{40}{148}; \quad m = 5$$

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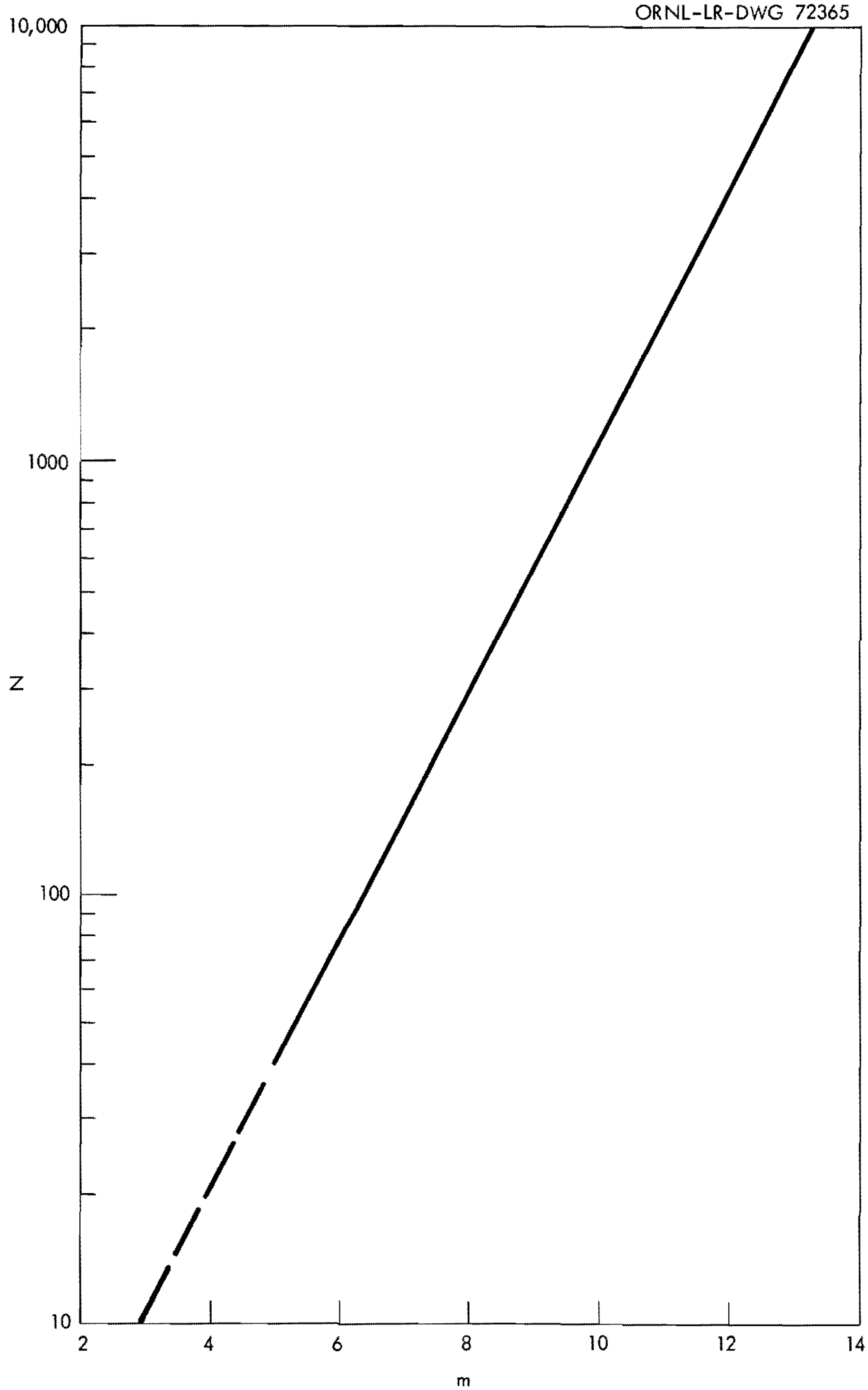

Fig. 5. A plot of Equation 23.

so that the error involved in using the first term of the series in eq. 22 is reduced to about 5% and less at and above m = 5. Thus, the dashed portion of the plot in Fig. 5 probably should not be used although there is a certain compensation of errors due to our having approximated series on both sides of eq. 21.

The important conclusion to be drawn from this work is that if we used $\pm$ 100 volt amplifiers and wanted to count 1,000 pulses to reset (u = 0.2 volts) as is the case at hand, then

$$m \geq 9.8$$

if we limit pulse losses to one or less in the 1,000.

6.0 DRIFT AND CAPACITOR LEAKAGE EFFECTS IN ANALOG MEMORY

There are two other important sources of error in analog memory operations. They are errors which are due to drift and capacitor leakage.

The new deposited film Mylar capacitors commonly used in analog computation have for all practical purposes essentially eliminated the capacitor leakage problem particularly when one considers that in the case at hand one to two hours is being considered between accounting circuit start and reset.

The drift problem is a bit more subtle. When the input voltage of a DC amplifier is kept constant, the output voltage changes with time from the original value to some other value. This phenomenon, called drift, is caused mainly by changes in plate and heater voltages, variations in resistor values, tube aging and grid current changes. Computing amplifiers are normally stabilized against drift by incorporating into the DC amplifier a chopper operated AC amplifier. The latter measures the drift as referred to the summing junction of the DC amplifier and introduces a correcting or stabilizing voltage into the DC amplifier.

Even the best chopper stabilizers do not completely eliminate drift which in the case of a memory (integrator) may be thought of in terms of the equation

$$-\frac{de_o}{dt} = \frac{e_D}{R_a C} \tag{24}$$

where

e_o = amplifier output voltage

e_D = amplifier drift voltage referred to amplifier summing junction

R_a = a resistance characteristic of the amplifier in the hold mode

C = the amplifier feedback capacitor

Equation 24 implies that drift in a memory can be more serious because e_o varies as the time integral of e_D. Of course, one might find some comfort in the fact that e_D may be positive or negative and that as a result there may be some self compensation. In actual experience such comfort is short lived because e_D is unpredictable and serious errors can be developed over the one to two hour period.

In a sense, both R_a and e_D are designed into an operational amplifier whereas the user can within reasonable limits select the capacitor C which for various computers ranges from 0.01 μf to 10 μf. According to eq. 24, one should be able to reduce the drift rate by a factor of 1,000 if he replaces a 0.01 μf capacitor with a 10 μf.

The important conclusion to be drawn at this point is that the user of memory must specify C such that it will satisfy his drift rate requirements without unnecessarily increasing mRC. If the input resistor is grounded during hold, then this resistor should be as large as possible but still compatible with mRC requirements.

7.0 ANALYSIS OF THE ASSISTED TRACKING OPERATIONS

If one specifies m and C based on previous error considerations, he may be confronted with a minimum pulse width-pulse space, mRC, that is too large for the counting speed required because R is determined by the permissible loading of the amplifier. Standard values of R for a $\pm$ 10 volt computer are 0.1 M and 0.01 M and for a $\pm$ 100 volt computer they are 1 M and 0.1 M. If C = 10 μf and m = 10 on a $\mp$ 100 volt computer, the shortest mRC time would be

$$mRC = 10 \cdot 0.1 \cdot 10 = 10 \text{ seconds}$$

In the application at hand, something less than 1 second is desired and the question arises as to what can be done to meet this final specification.

There are at least two methods for reducing the memory time constant. The simplest which we will discuss first involves inserting at point D in Fig. 3 the two amplifiers, i.e. buffer amplifier, with resistor R_T as shown in Fig. 1. The junction of the two R resistors in Fig. 3 become the summing junction of the buffer amplifier which has +1 gain. Thus, the voltage $e_o + e_i$ appears at the output of the buffer amplifier. Hence, one may write the Kirchoff nodal equation for current flow through the summing junction of the main amplifier in Fig. 3 as

$$\frac{e_o + e_i}{R_T} + C\,\frac{de_o}{dt} = 0 \tag{25}$$

which in transfer function form becomes

$$-\frac{\bar{e}_o}{\bar{e}_i} = \frac{1}{R_T CS + 1} \tag{26}$$

Thus the new time constant, τ, where

$$\tau = R_T C \tag{27}$$

replaces the old one, i.e. RC.

At first sight, it would appear that τ can be reduced to zero by reducing R_T to zero. This isn't so because

$$R_T = R_o + R_1 \tag{28}$$

where

R_o = open loop output impedance of the buffer amplifier

R_1 = any additional resistance inserted after the buffer amplifier and in series with its output

Now since the instantaneous current, i, flowing through R_T is

$$|i| = \frac{e_o + e_i}{R_T} \tag{29}$$

we have

$$R_T = \frac{e_o + e_i}{|i|} \tag{30}$$

The current rating of standard $\pm$ 100 volt operational amplifiers is about 25 ma. Now in the worst case of the R-M wherein e_o builds down to -100 and reset sets in causing e_i to become -100 we would need

$$R_T = \frac{200}{0.025} = 8 \text{ K } \Omega \tag{31}$$

Thus, if C = 10, m = 10, then

$$m\, R_T C = 10 \cdot 0.008 \cdot 10 = 0.8 \text{ secs} \tag{32}$$

which represents a definite improvement.

Typical values of R_o for standard operational amplifiers range from less than 500 ohms to 1-5 K ohms. Thus, the value R_T = 8 K ohms is compatible with R_o, i.e. $R_T \geq R_o$.

At least one manufacturer that the author knows of manufactures analog memory equipment and a buffer amplifier as such which has an R_o = 20 ohms and a high current rating so that R_T can essentially be reduced to 20 to 50 ohms.

In "beefing up" the track circuit one must keep in mind that the main amplifier output is on the other end of the capacitor and any current that flows through the capacitor must also be accommodated at the output of the main amplifier. A typical response for a main amplifier is plotted in Fig. 6. Thus, so long as $C \leq 0.01$ μf the response time is independent of C, and the output voltage will rise or fall linearly with time instead of following the exponential decay of eq. 25 or 26. With the 0.01 μf capacitor, an amplifier would rise 100 volts in 80 μs.

If C is beyond 0.01, say 10 μf, then the response curve would be expontential in character and to rise 100 volts according to Fig. 6 would require

$$\frac{10 \cdot 10^4}{10^3} = 100 \text{ ms}$$

If R_1 were zero and $R_o = 20\ \Omega$ for a buffer amplifier, the time according to eq. 27 is

$$10\tau = 10 \cdot 20 \cdot 10^{-6} \cdot 10 = 2 \cdot 10^{-3} \text{ secs}$$

$$= 2 \text{ ms}$$

The reason for this discrepancy is that with the large capacitor and the small buffer amplifier open loop output impedance of 20 Ω the main amplifier becomes the current controlling factor. The open loop output impedance of the main amplifier is about 1,000 ohms so that

$$10\tau = 10 \cdot 10 \cdot 10^{-3} \text{ secs}$$

$$= 100 \text{ ms}$$

The second method for reducing memory time is based on the circuit shown in Fig. 7. The potentiometer R_2 may be set so that

$$e_2 = -G\, e_o \tag{33}$$

Also

$$\frac{e_2 - e_1}{R_T} = C_1 \frac{de_1}{dt} \tag{34}$$

and

$$C_1 \frac{de_1}{dt} + \frac{e_o}{R} + \frac{e_i}{R} + C \frac{de_o}{dt} = 0 \tag{35}$$

Thus

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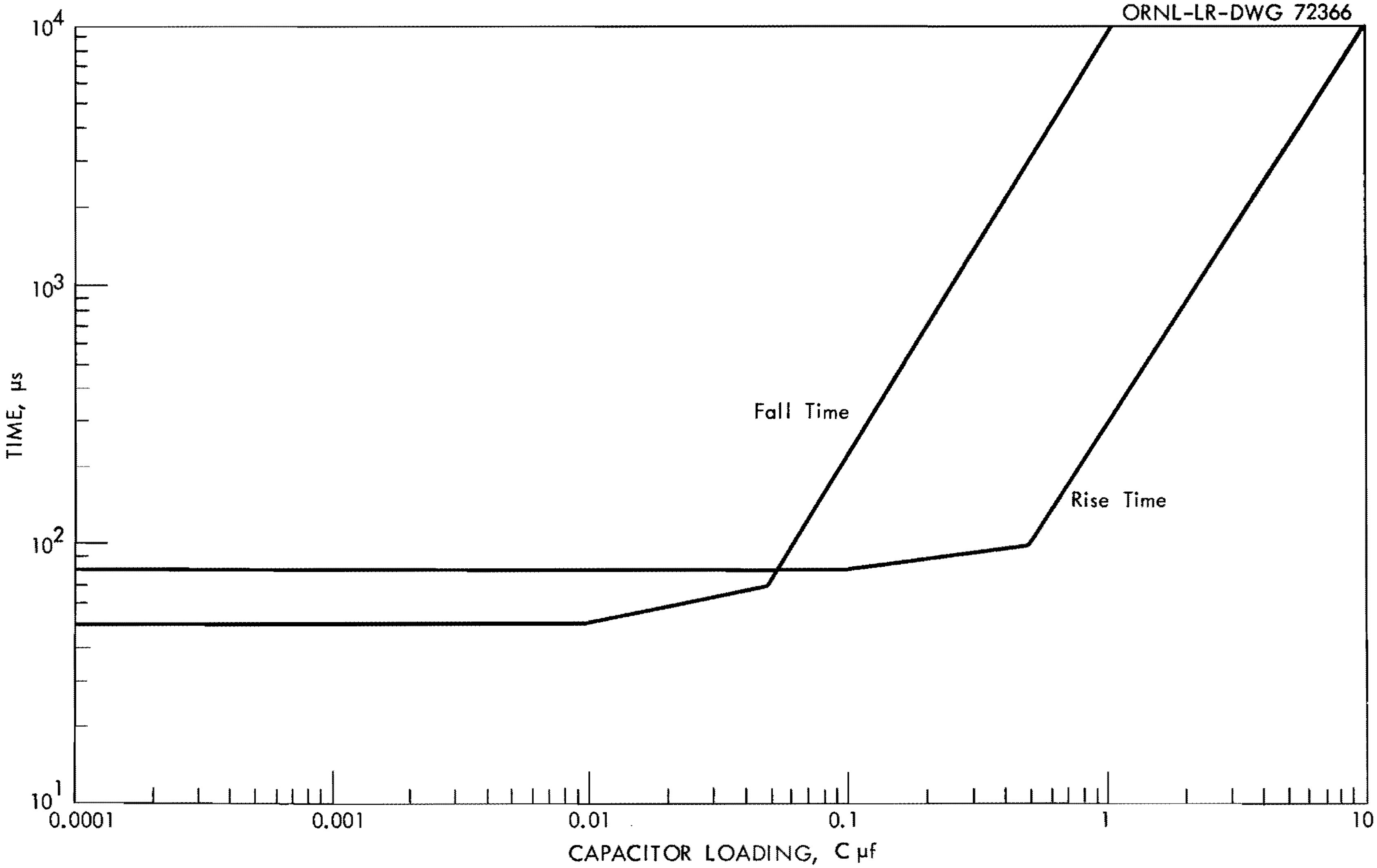

Fig. 6. Operational amplifier response to 100 V.P.P square wave input.

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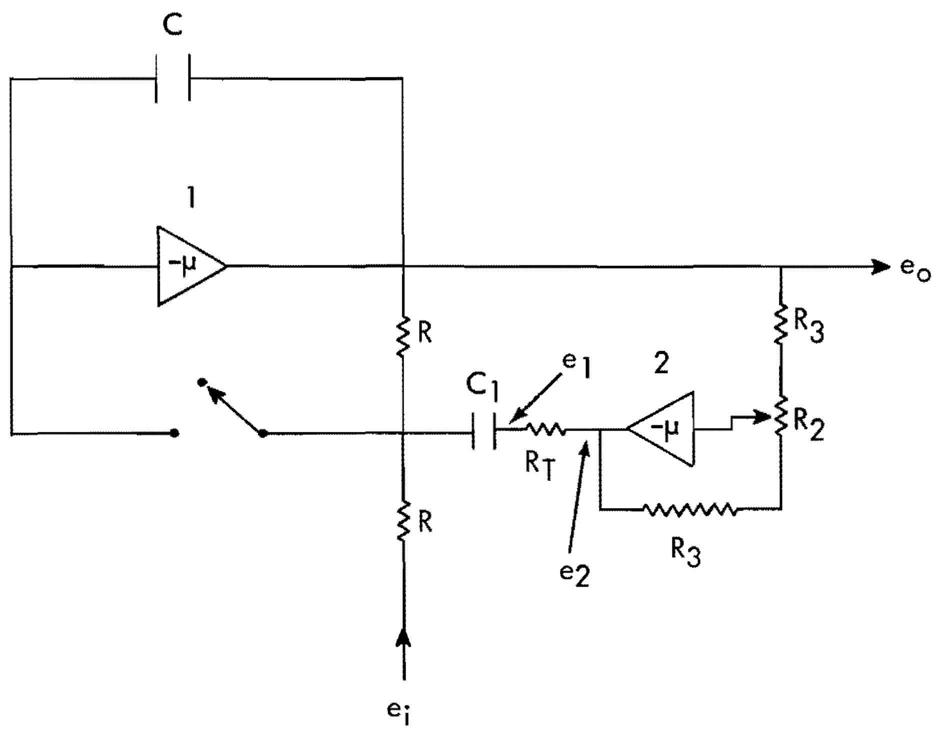

Fig. 7. An alternate reset assist.

$$-\frac{G\bar{e}_o + \bar{e}_1}{R_T} = C_1 \, S \, \bar{e}_1 \qquad (36)$$

or

$$\bar{e}_1 = -\frac{G\bar{e}_o}{R_T C_1 S + 1} \qquad (37)$$

Transforming eq. 35 and eliminating $\bar{e}_1$ between it and eq. 37 leads to

$$\left[\left(-\frac{GRC_1}{R_T C_1 S + 1} + RC\right) S + 1\right] \bar{e}_o + \bar{e}_i = 0 \qquad (38)$$

or the transfer function is

$$-\frac{\bar{e}_o}{\bar{e}_i} = \frac{1}{\left[RC - \frac{GRC_1}{R_T C_1 S + 1}\right] S + 1} \qquad (39)$$

To avoid regeneration

$$RC \geq \frac{GRC_1}{R_T C_1 S + 1}$$

The resistance R_T which includes amplifier 2 output impedance is required because C_1 may be larger, i.e. > 500 μμf, and amplifier 2 would tend to go into oscillation if $R_T = R_o$. To reduce distortion and transfer maximum power at the output of amplifier 2 one should choose R_1 in eq. 28 so that

$$R_1 = R_o \qquad (40)$$

and

$$R_T = 2R_o \qquad (41)$$

In general, C_1 should be chosen so that

$$R_T C_1 \ll RC$$

Under these circumstances the $R_T C_1 S$ term dies out shortly after tracking starts and eq. 39 becomes

$$-\frac{\bar{e}_o}{\bar{e}_i} = \frac{1}{RC\left[1 - \frac{C_1}{C} G\right] S + 1} \tag{42}$$

where G is chosen so that

$$\frac{C_1}{C} G \leq 1 \tag{43}$$

and the time constant now is

$$\tau = RC\left[1 - \frac{C_1}{C} G\right] \tag{44}$$

and the minimum track or hold time is

$$t = mRC\left[1 - \frac{C_1}{C} G\right] \tag{45}$$

If we wanted

$$t = 0.1 \text{ sec}$$

and

$$m = 10$$

$$R = 0.1 \text{ M}$$

$$C = 10 \ \mu f$$

then G must be set such that

$$\left[1 - \frac{C_1}{C} G\right] = \frac{0.1}{10 \cdot 0.1 \cdot 10} = 0.01 \tag{46}$$

8.0 SIMPLIFIED ACCOUNTING CIRCUIT BASED ON TRANSFER FUNCTION SIMULATION

It has been seen that to obtain a counting speed of 5 pulses or cycles per second requires per channel about $1,500.00 worth of power supplies, amplifiers, comparators, and associated equipment. This derives basically from the long term drift stability requirement and the use of conventional analog accounting techniques.

If electronic accounting is to be used in a multitude of applications, it is of course desirable to have some flexibility in counting speed at minimum cost. Since time is one of the variables logged, we might for example want to count the 60 cycle line frequency which is out of the range of the circuits previously discussed.

On technical grounds conventional analog accounting operates at a disadvantage because among other things the voltage increment to be remembered decreases as the total count between start and reset increases. We might say that the "driving force" for learning diminishes as the total count increases. Thus, in searching for a more flexible accounting circuit one should concentrate his efforts on raising this voltage level as close to the reference limits available at the analog equipment. It should, if possible, also be independent of the total count. Reset over large voltage intervals must also be avoided.

The basic requirement of any analog memory circuit is that its transfer function be of the form of or a reasonable approximation to that given in eq. 3. The simplest circuit that comes to mind which avoids reset over large voltage intervals and also utilizes a constant step voltage and reference level independent of total count is the one shown in Fig. 8 and costs about $600.00. Let's for the moment omit the comparator at the left of the figure and assume that line A is attached to +100 volts directly. Line F is broken and the R, R_o resistors are omitted. Line B is connected to ground through diode D_1. When the meter switch closes a positive voltage of 100 volts is applied to capacitor C causing it to charge through R_1, C_1. Electrons will flow from C_1 through R_1 to C therefore making line B positive and the amplifier and its network integrates with its output going negative while line B is positive. After 10 R_1 C integration has essentially ceased because C is fully charged. When the meter switch is opened the electrons will flow from the right plate of C to the left if they have the opportunity to do so. Diode D_1 becomes conducting with its forward resistance being about 13 to 14 ohms. Thus line B becomes only slightly negative. If D_1 didn't conduct the electrons would flow through the grid R_1 to C_1 and line B would become just as negative during discharge as it did during charge and the integrator output voltage would return to its starting value. With the diode D_1 the integrator remembers the pulse leading edge effect no matter how broad the pulse may be just so it is not narrower than 10 R_1 C_1. While the meter switch is open and line B is essentially at zero volts the integrator simply holds (remembers) at its output the effect of one pulse. This effect is additive for succeeding pulses.

When the integrator output reaches -100 volts we need the comparator to switch line A to -100 volts and line B from diode D_1 to diode D_2. The resistors R and R_o determine the sum voltage between line A and the integrator output and divide it down to millivolt level for direct feed to the digital data logging equipment. The voltage sum is used instead of the integrator output alone so that in data logging one can distinguish, by sign, voltages logged from one swing in contrast to those logged from a previous or succeeding swing of the integrator output.

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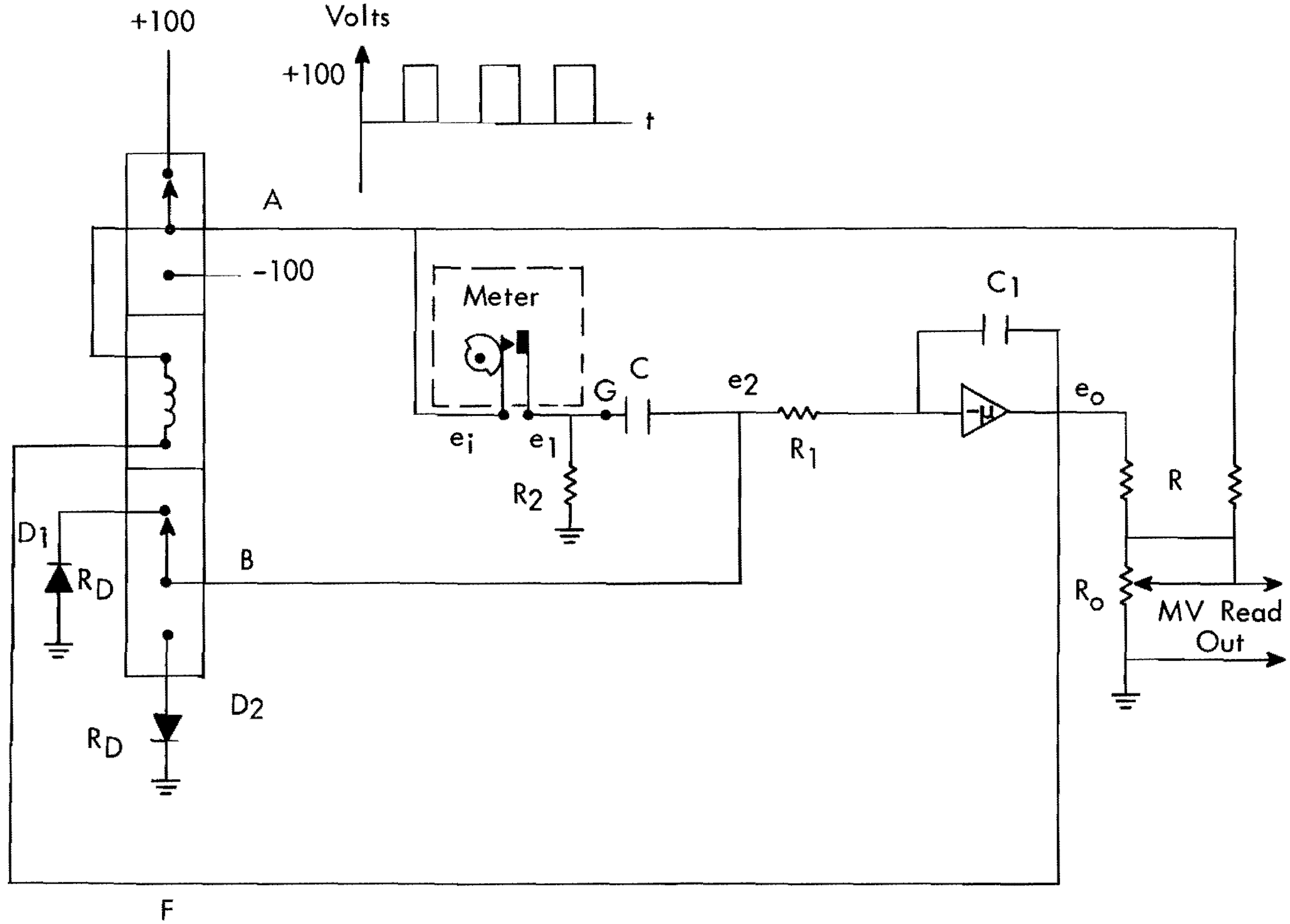

Fig. 8. Simplified approach to pulse accounting.

9.0 ANALYSIS OF THE SIMPLIFIED ACCOUNTING CIRCUIT

During charge $e_1 = e_i$ and the nodal equation for current at line B is

$$C \frac{d(e_2 - e_i)}{dt} + \frac{e_2}{R_1} + \frac{e_2}{R_D} = 0 \qquad (47)$$

or

$$\tau S \ (\bar{e}_2 - \bar{e}_i) + \bar{e}_2 = 0 \qquad (48)$$

and

$$\frac{\bar{e}_2}{\bar{e}_i} = \frac{\tau S}{\tau S + 1} \qquad (49)$$

where

$$\tau = \frac{C}{\frac{1}{R_1} + \frac{1}{R_D}} \qquad (50)$$

$\simeq R_1 C$ during charge because $(R_D)_{back} \gg R_1$, i.e. $R_D > 2M$

For the integrator, of course

$$\frac{\bar{e}_o}{\bar{e}_2} = - \frac{1}{R_1 C_1 S} \qquad (51)$$

Thus

$$\frac{\bar{e}_o}{\bar{e}_i} = - \frac{\tau}{R_1 C_1} \left[\frac{1}{\tau S + 1}\right] \qquad (52)$$

$$\simeq - \frac{C}{C_1} \left[\frac{1}{\tau S + 1}\right] \qquad (53)$$

During discharge the meter switch is open. The line connecting the switch to the accounting circuit has capacitance say C_2 which is not shown in Fig. 8 but which is shunted across the open switch. Assuming $R_2 = R_1$ the nodal equations are

$$\frac{C}{C_1} \; e_i$$

As illustration, lets consider the TR-10 analog computer manufactured by Electronic Associates. Its reference supply voltage is $\pm$ 10 volts. The standard patching components are 10 μf, 0.1 M and 0.01 M. To keep the time constant, $\tau = R_1C$, short lets choose

$$R_1 = 0.01 \text{ M}$$

To keep memory drift low, the 10 μf capacitor will be used so that

$$C_1 = 10 \ \mu\text{f}$$

Next let's assume that we want to count pulses up to 500 per sec. The pulse frequency, f, then is

$$f = 500 \text{ cycles/sec}$$

with period T being given by

$$T = \frac{1}{f} = 0.002 \text{ seconds}$$

The pulse width-pulse space is then 1 ms so that

$$10 \ \tau = 10 \cdot 0.01 \cdot C = 0.001$$

or

$$C = 0.01 \ \mu\text{f}$$

which is a standard size capacitor and readily available on the market. Thus in our swing of e_i from $-e_i$ to $+e_i$, we can count

$$2 \ \frac{10}{0.01} = 2{,}000 \text{ pulses}$$

If C is increased to 0.1 μf, then in one swing we can count 200 pulses at a rate of 50 pulses per second. These computations apply to any reference supply voltage. In either case e_o would, for example, simply register a pulse height 10 times greater on a $\pm$ 100 volt computer as on the $\pm$ 10 volt TR-10 and this to take into account automatically the differences in reference level. The output voltage divider in Fig. 8 would on a $\pm$ 100 volt computer be simply set to 10% of that on the $\pm$ 10 volt TR-10.

Thus the development of a simple analog accounting circuit based on transfer function simulation has far exceeded all expectation. The equipment cost is about half that to obtain 5 cps with conventional circuitry. At the same time the frequency limitation has for all practical applications been removed. The one limitation is the amplifier pass band or break point frequency which is 25 K to 1 M depending on quality of amplifier. The other is the smallest practical value of C. According to eq. 64, we certainly would not want to take C smaller than 10^{-5} to 10^{-6} which is in the range of 1-10 μμf. This value is dangerously low because of inherent circuit capacitances. If we limit C to ≥ 100 μμf with low capacitance leads to meter switch then

$$C \geq 10^{-4} \ \mu f$$

which means that

$$f \leq 50 \text{ K cps}$$

probably approximates some sort of reasonable limit. To go this high one, of course, would have to replace the comparator with solid state switches in Fig. 8 because high quality mechanical comparators have a throw time of about 1 ms.

It should be mentioned in passing that the circuit of Fig. 8 is the one that one intuitively would suspect as the likely candidate for the job at hand and that there is available a much more general approach to network synthesis. The simplest such approach is to recognize that for a single amplifier

$$\frac{\bar{e}_o}{\bar{e}_i} = - \frac{Z_f}{Z_i}$$

where

Z_f = feedback network transfer impedance

Z_i = input network transfer impedance

and then to resort to a table of transfer impedances and networks to choose Z_f and Z_i such that the requirements are satisfied. Thus, there are many combinations of Z_f and Z_i that could be used with that of Fig. 8 being one of if not the simplest one. Unless one could find a substantial reason for doing so, it would be difficult to justify using a more expensive network. Also, in a complex network which usually involves more than one time constant "learning" calculations must be based on the longest time constant.

One substantial reason for a refined network can be anticipated in the case of "contact bounce" in a snap action switch. In this case, when the switch snaps on the contacts close but only momentarily so. The impact causes the switch contacts to open and close several times before finally staying closed. This produces serrations in the leading end of what otherwise would be a solid pulse. To prevent these serrations from being counted, one would have to insert a filter at point G in Fig. 8. The filter would consist of a resistor R_3 inserted at point G followed by a capacitor C_3 which is grounded at one end and connected between R_3 and C at the other.

The transfer function for Fig. 8 with the filter added is

$$\frac{\bar{e}_o}{\bar{e}_i} = \frac{C}{C_1} \frac{1}{(1 + S\tau_1)(1 + S\tau_2)} \tag{69}$$

where the τ are given by

$$\tau = \left[R_3C_3 + R_1C + R_3C\right] \frac{1 \pm \sqrt{1 - 4\dfrac{R_3C_3\ R_1C}{\left[R_3C_3 + R_1C + R_3C\right]^2}}}{2} \tag{70}$$

If, for good filtering action, one chooses

$$R_1 = R_3$$

$$C_3 \gg C$$

then

$$\tau_1 \simeq R_3C_3$$

$$\tau_2 \simeq 0$$

During discharge of C, line B is essentially grounded. The transfer impedance of the network consisting of R_3, C_3, R_2, and C is

$$R_2 \frac{1 + S\theta\tau}{1 + S\tau} \tag{71}$$

where the time constant, τ, for discharge is

$$\tau = (R_2 + R_3)(C_3 + C) \tag{72}$$

and

$$\theta = \frac{R_3}{R_3 + R_2} < 1 \tag{73}$$

A refined version of the serration filter for positive pulses is shown in Fig. 9. For negative pulses, the polarity of the diodes and reference voltage would have to be reversed via a comparator.

In Fig. 9, the voltage e_4 is -10 volts until the meter switch closes. The voltage e_3 is zero and so the diode D_2 is non-conducting thus causing

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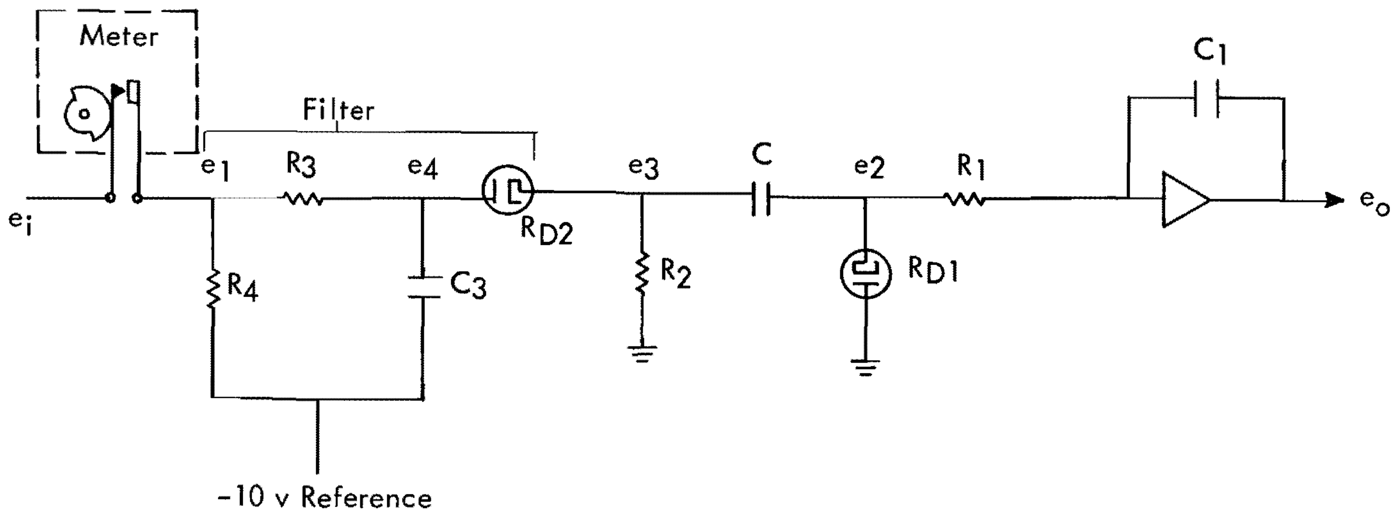

Fig. 9. A refined serration filter and accounting network.

isolation of the integrator from voltage e_4. When the switch closes, capacitor C_3 begins to charge through R_3 and e_4 rises from -10 volts. If the switch opens, i.e. the first serration is completed, before e_4 reaches zero diode D_2 still doesn't conduct and the serration is not transmitted to the integrator. While the meter switch is open e_4 dies back toward -10 volts due to the discharge of C_3 through R_3 and R_4. Subsequent serrations will also expend themselves in R_3 and R_4 if the filter time constant is selected properly. Only the solid portion of the pulses will be substantial enough, i.e. of long enough duration, to cause e_4 to rise above zero and only then will that portion of the pulse be transmitted to the integrator. Under these circumstances if the solid part of the pulse is long enough e_3 will rise until

$$e_3 \simeq e_4 \simeq \frac{R_2}{R_2 + R_3} e_1 \tag{74}$$

Electrons will be "pumped" out of C_1 via R_1 and stored on the right hand plate of C. When the meter switch opens e_4 dies away to -10 volts and e_3 to zero volts. The electrons leave C via diode D_1.

While diode D_2 is non-conducting and the meter switch is closed, we have

$$e_4 - e_1 + R_3C_3 \frac{de_4}{dt} = 0 \tag{75}$$

or

$$-\int_{e_4 = -10}^{e_4 = 0} d \ln (e_4 - e_1) = \frac{1}{R_3C_3} \int_0^t dt \tag{76}$$

where

$$e_1 = +10$$

Thus

$$R_3C_3 = \frac{t}{\ln 2} \tag{77}$$

is the worst case time constant, i.e. the one wherein the combined serrations last for t seconds and are spaced by infinitely small spaces. If t = 5 ms, then

$$R_3C_3 = \frac{0.005}{\ln 2} = 0.0072 \text{ secs} \tag{78}$$

Such a filter will stop any combination of +10 volt serrations the total time of which does not exceed 5 ms duration.

After e_4 reaches zero and the solid portion of the pulse sets in, then

$$e_3 \simeq e_4$$

because

$$R_{D2} = 14 \cdot 10^{-6} \text{ M}$$

Also

$$R_{D1} > 2 \text{ M}$$

Therefore

$$\frac{\bar{e}_o}{\bar{e}_1} \simeq - \frac{R_2}{R_2 + R_3} \frac{C}{C_1} \frac{1}{\tau_1 \ \tau_2 \ S^2 + (\tau + \tau_2) \ S + 1} \tag{79}$$

where

$$\tau = \left[\frac{R_2 R_3}{R_2 + R_3} (C_3 + C) + CR_1\right] \frac{1 \pm \sqrt{1 - 4 \ \dfrac{R_2}{R_2 + R_3} \dfrac{R_3 C_3 \ R_1 C}{\left[\dfrac{R_2}{R_2 + R_3} R_3 (C_3 + C) + CR_1\right]^2}}}{2} \tag{80}$$

If, for example,

$$R_1 = R_2 = R_3$$

$$C = C_3$$

Then

$$\tau = R_1 C \ (1 \pm 0.707) \tag{81}$$

If, on the other hand,

$$R_1 = R_2 = R_3$$

$$C_3 \gg C$$

then

$$\tau_1 \simeq \frac{R_3C_3}{2} \tag{82}$$

$$\tau_2 \simeq 0$$

At the end of the pulse, the electrons on C return to ground via $R_{D1} \simeq 14 \cdot 10^{-6}$ M leaving e_2 essentially at ground. Eventually $R_{D2} \gg 2$ M and e_4 decays from zero down to -10 volts. This latter time constant is

$$\tau = (R_3 + R_4)C_3 \tag{83}$$

In a conservative design of the type just discussed, the pulse frequency f would be related to the design such that

$$\frac{1}{f} = 2 \cdot 10 \cdot \tau_{largest} \tag{84}$$

where

$\tau_{largest}$ = longest time constant involved in the overall design

10.0 RESULTS

The first portion of the experimental investigation was carried out on the TR-10 computer. A finger pump with a snap action micro-switch attached to one of the fingers was used to simulate a meter switch. The pulse rate could be adjusted by simply setting the speed reducer.

The circuit of Fig. 1 was patched first. The buffer amplifiers were not used and C = 0.1 μf. This circuit accommodated the 9 cps limit of the pump very nicely. However, drift presented a problem. A series of store tests were run over a period of one hour. The hold voltage was read out on the DVM (Digital Volt Meter) of the logging equipment where 50 m volts corresponded to 10 volts at the TR-10. The voltage read out on the DVM would change up to ± 3 mv in one hour depending on the zeroing of the balance adjustment on the amplifiers. A differentiator followed by a variable high gain amplifier was hooked to the hold memory and the output fed back to its input to increase the drift stability. The drift was reduced to about ± 1 mv per hour but we couldn't be certain whether this was due to drift rate compensation or multiple amplifier drift cancellation and so this idea was abandonded.

The capacitor C in Fig. 1 was increased to 10 μf and in one hour the DVM which could be read to 0.1 mv with 0.01 mv being estimated (due to DVM equipment background noise) showed no detectable voltage change in one hour. For this reason the 10 μf capacitor was used in all subsequent work. Under these circumstances, however, the pump could not be run much faster than the theoretical 1 cps otherwise at reset the memories would not reset to the full -10 volts.

A buffer amplifier consisting of a pair of inverters was added as shown in Fig. 1 and this enabled us to attain a little better than 5 cps before the memories would not completely reset to -10 volts. The resistance R_1 in R_T was varied between 50 to 250 ohms with but little effect. If R_1 was reduced below 50 ohms the amplifiers eventually overloaded because the output of the buffer was approaching the ground condition.

The circuit of Fig. 7 was investigated with

$$
\begin{aligned}
R &= 0.01\ \mathrm{M} \\
R_3 &= 0.1\ \mathrm{M} \\
R_3 &= 5\ \mathrm{K} \\
C &= 10\ \mu\mathrm{f} \\
C_1 &= 10\ \mu\mathrm{f}
\end{aligned}
$$

With R_T reduced so that $R_1 = 0$ amplifier 2 overloaded and this was true for R_1 up to about 10 ohms. The pot R_2 was set by noting its effect on e_o. As soon as R_2 was set so 1-G became even slightly negative e_o swung wildly thus indicating regeneration. In our tests R_2 was set so as to barely avoid regeneration. The results showed that we could successfully count the 5 cps or better but that we could not cover the maximum of 9 cps output of our pump because the main amplifier wouldn't reset completely to -10 volts. The effect of increasing R_T is quite pronounced as one can appreciate from eq. 39. The limit of 5 cps was due to the capacitor C and the open loop output impedance of the main amplifier rather than limitations due to the assisting circuit. Figure 7 has possibilities that actually were not investigated thoroughly primarily due to the high amplifier requirement and limitations of the amplifiers in the TR-10.

The circuit of Fig. 1 as such or with the buffer amplifier or the assisting circuit of Fig. 7 would have a broader range of application if the A comparator were used to switch +u volts to -u volts when Z = v instead of switching $(+Z)_{max}$ to -w. This would cause the Z voltage to gradually drop from $(+Z)_{max}$ to $(-Z)_{min}$, i.e. to produce a triangular wave instead of a saw tooth wave. It would eliminate the need for resetting the memories over the maximum voltage range which demands high currents.

At this point the simplified accounting circuit of Fig. 8 was patched into the TR-10. Here we used $\pm$ 10 reference supply voltage and

$$
\begin{aligned}
R_1 = R_2 &= 0.01\ \mathrm{M} \\
C &= 0.1,\ 0.01\ \mu\mathrm{f} \\
C_1 &= 10\ \mu\mathrm{f} \\
R &= 0.1\ \mathrm{M} \\
R_o &= 5\ \mathrm{K}
\end{aligned}
$$

With the 0.1 μf capacitor, the output of the amplifier in Fig. 8 should swing from -10 volts to +10 volts or vice versa during the course of 200 counts. This figure should increase to 2,000 counts with the 0.01 μf capacitor.

The results obtained from the circuit of Fig. 8 are plotted in Figs. 10 and 11. Two different types of pulse generators were used. The one involved the pump and its switch with reference voltage source. The other involved a square wave generator the details of which are shown in Fig. 12. Here the frequency of the square wave is given by

$$f = \frac{2.5}{k}, \text{ cps}$$

where

k = fractional pot setting

The pump switch cycles, at the lower switching frequencies, were actually counted by hand, i.e. by placing the finger on the switch and hand counting for a time interval established with a stop watch. From this information the input pulse frequency could be calculated. The output frequency was calculated by counting on the TR-10 voltmeter the number of pulses required in swinging e_o of Fig. 8 from -10 volts to +10 volts and back again to -10 volts. This event was also clocked with a stop watch. The resulting data points are shown as circles in Fig. 10. At the higher pump frequencies the stroboscope was used which permitted calculating the input switching frequency and these results are shown as crosses in Fig. 10. The stroboscope itself left something to be desired because we had to work with a second harmonic at the low end of the low range. Further in standardizing the low range on the stroboscope difficulty was experienced in judging true balance based on the 900 cpm reed provided for this purpose. This is one reason the results are plotted on log-log paper. Constant fractional error in the data does not prevent comparing data for parallelism with the solid theoretical line. The stroboscope data are (in trend) essentially parallel to the theoretical straight line.

The data points in Fig. 10 are those obtained with the circuit of Fig. 12. The input cps was calculated from the formula relating frequency and pot setting k in Fig. 12. The output was determined in the same way as before.

The pulse generator data show good parallelism to theoretical up to about 60-70 cps. The circuit of course was actually designed for a maximum of 50 cps. In this regard we should remark that the 0.1 μf capacitor that we used was a stock ± 20% tolerance type and not a standard computing capacitor of ± 0.1% tolerance.

Next we changed C in Fig. 8 to 0.01 μf and then used the square wave drive as shown in Fig. 12 to obtain the data shown in Fig. 11. The input frequency f was computed from the pot setting k. The output frequency was calculated by timing several TR-10 voltmeter swings of -10 volts to +10 volts and return and a knowledge of pulses per swing. Based on theory we calculated that a swing from -10 volts to +10 volts on the meter corresponded to 2,000 pulses. It is obvious from Fig. 11 that in the vicinity of 200 cps and above the output begins to drop off. Since our design is for 500 cps this indicates a saturating effect of some sort which apparently sets in in

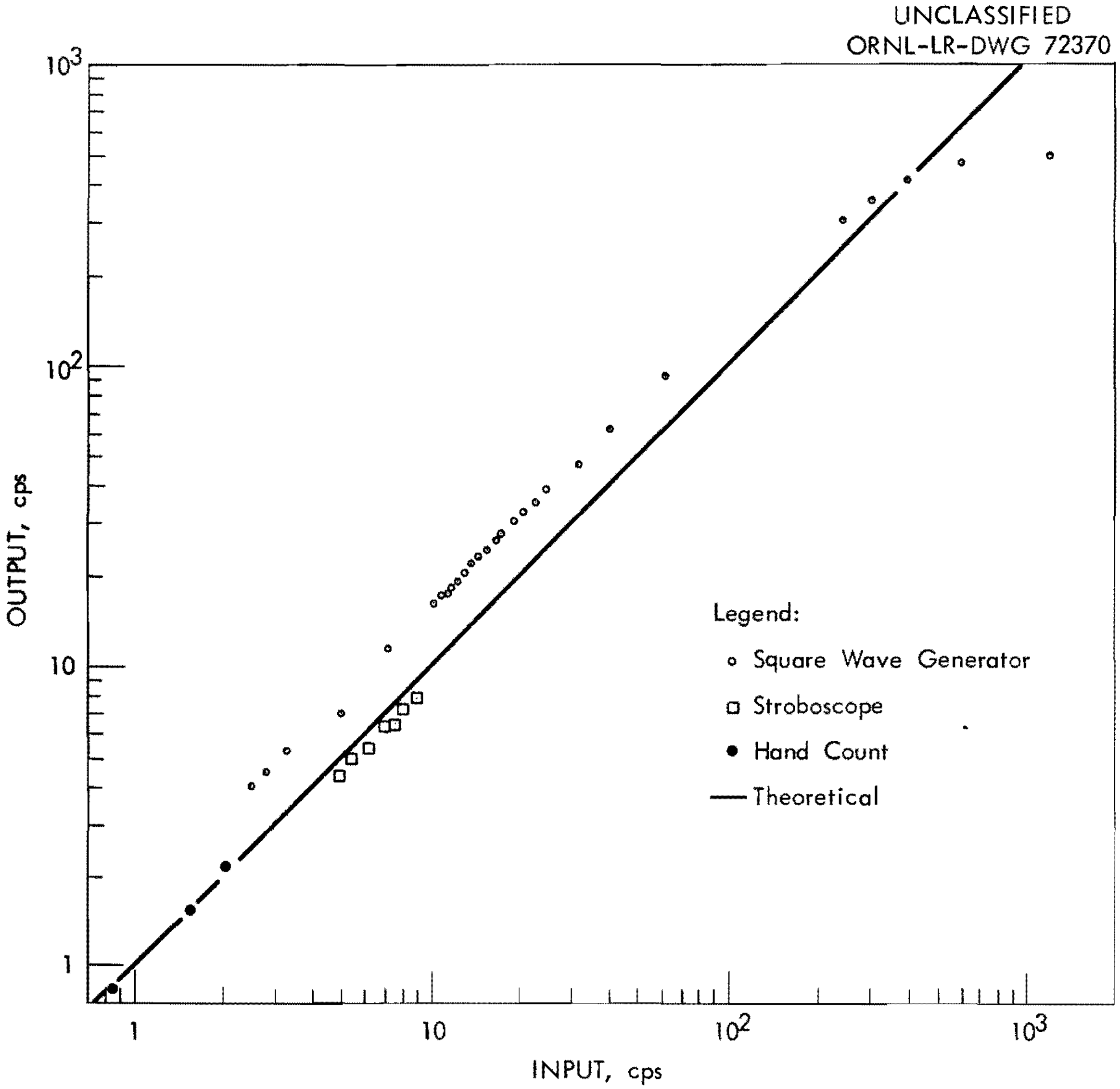

Fig. 10. Evaluation of simplified accounting circuit with C = 0.1 μf.

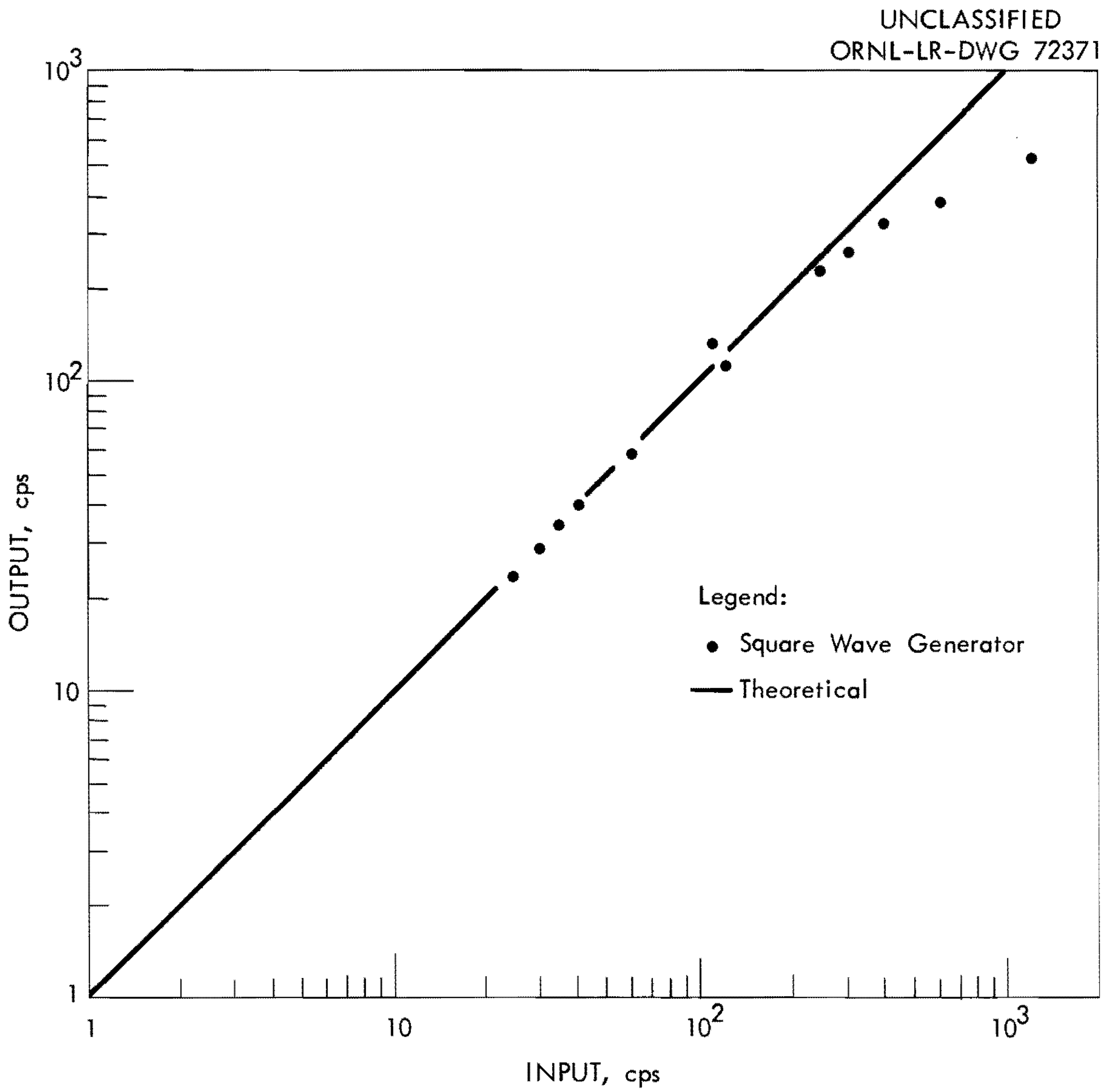

Fig. 11. Evaluation of simplified accounting circuit with C = 0.01 μf.

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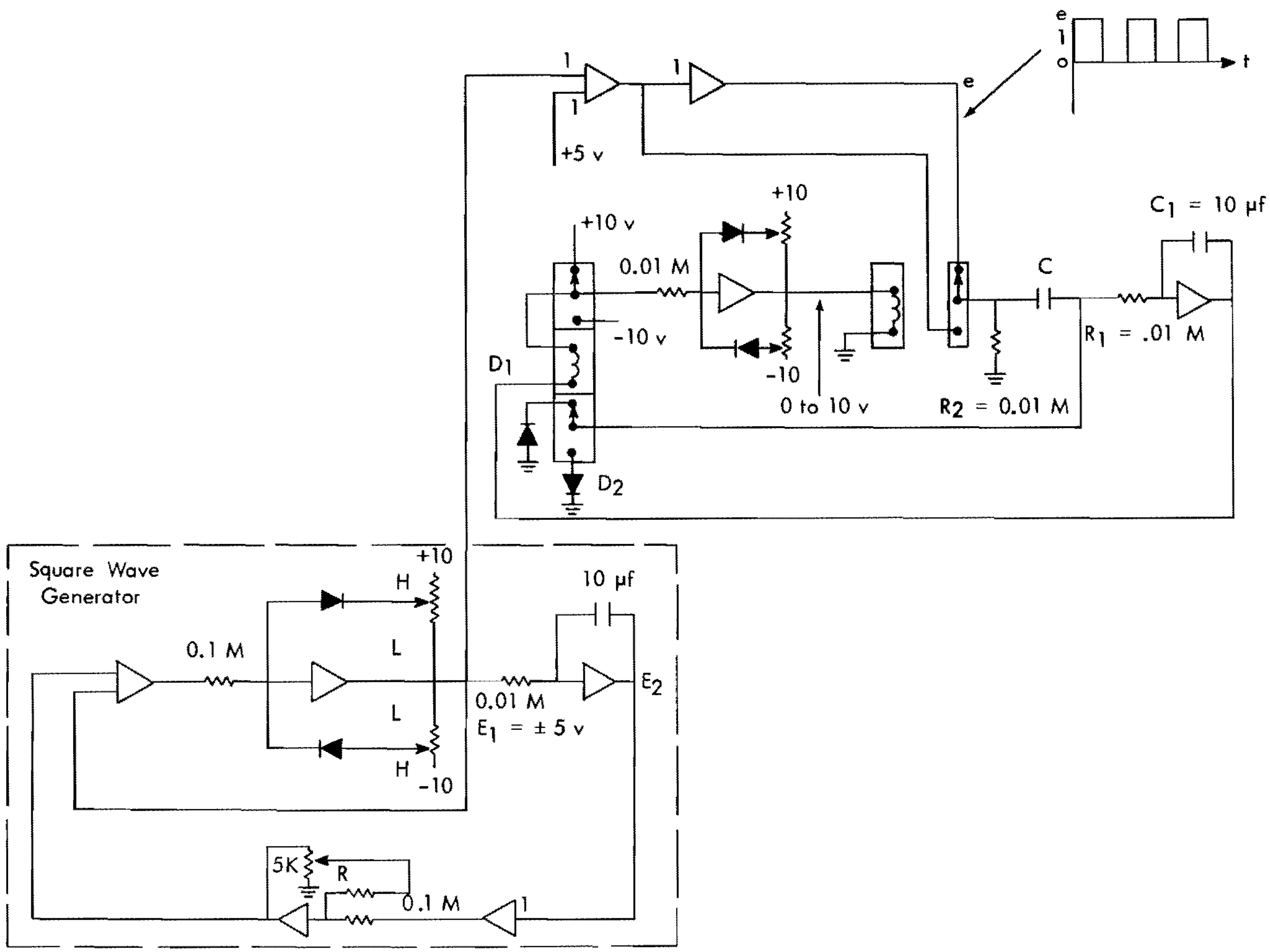

Fig. 12. Simplified accounting circuit driven by square wave generator.

the square wave generator of Fig. 12 rather than in the circuit under investigation. This was confirmed by the fact that the square wave generator overloaded in the higher frequency range. It could be that this is simply a practical limitation due to TR-10 quality components and circuitry.

It should be pointed out that in Fig. 8 the voltage e_1 is raised to e_i when the meter switch is closed and it is left to be determined by circuit parameters when it is open. The square wave generator on the other hand forces e_1 to track a true square wave. This should, however, have no bearing on the final results because when the switch is open e_2 is essentially at ground because one or the other of the diodes is conducting.

An interesting result from this work was that according to the design equations, a TR-10 meter swing from -10 v to +10 v should represent

200 pulses; 0.1 μf capacitor
2,000 pulses; 0.01 μf capacitor

This was essentially confirmed when the square wave generator was used as drive. However, when the pump snap switch was used to determine the counts in a meter swing there were found

244 pulses; 0.1 μf capacitor
845 pulses; 0.01 μf capacitor

The number of pulses is off by about 20% in the case of the 0.1 μf capacitor and this is within capacitor tolerance. This, however, must not be the case for the 0.01 μf capacitor. The obvious explanation is based on the fact that due to "contact bounce" in the pump switch there are serrations that are actually being partially counted by the lower time constant circuit (0.01 μf). In fact, to the latter circuit it appears that there are on the average at least

$$\frac{2,000}{845} - 1 = 1.38 \text{ serrations/pulse}$$

At this point we were convinced that we should confirm by independent investigation the presence of serrations in the pump switch pulse. While we were set up to do this we also decided to investigate the various switches actually used in the various meters of the calciner operation. For purposes of this investigation we hooked an oscilloscope to the switch in question as shown in Fig. 13. The scope was set to trigger on the first serration of the pulse so that the scope display was as shown in Fig. 14. The time t_s is the time during which serrations were observed. Time t_p is the total pulse width and t_o the pulse space. The ratio of t_p and t_o is constant for a given switch. Both t_p and t_o vary with meter cam speed and for this reason each process switch cam was operated at the maximum speed actually used in the process. The results are shown in Table 1.

The pulse displays for all the snap action micro-switches were consistent in repeatability and uniformity. Each such switch produced 4-8 distinguishable serrations in the front end of the pulse. The serration

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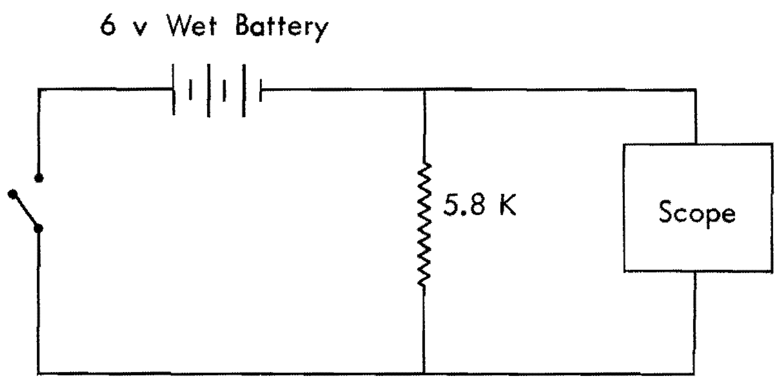

Fig. 13. Scope-switch circuit for observing character of pulses.

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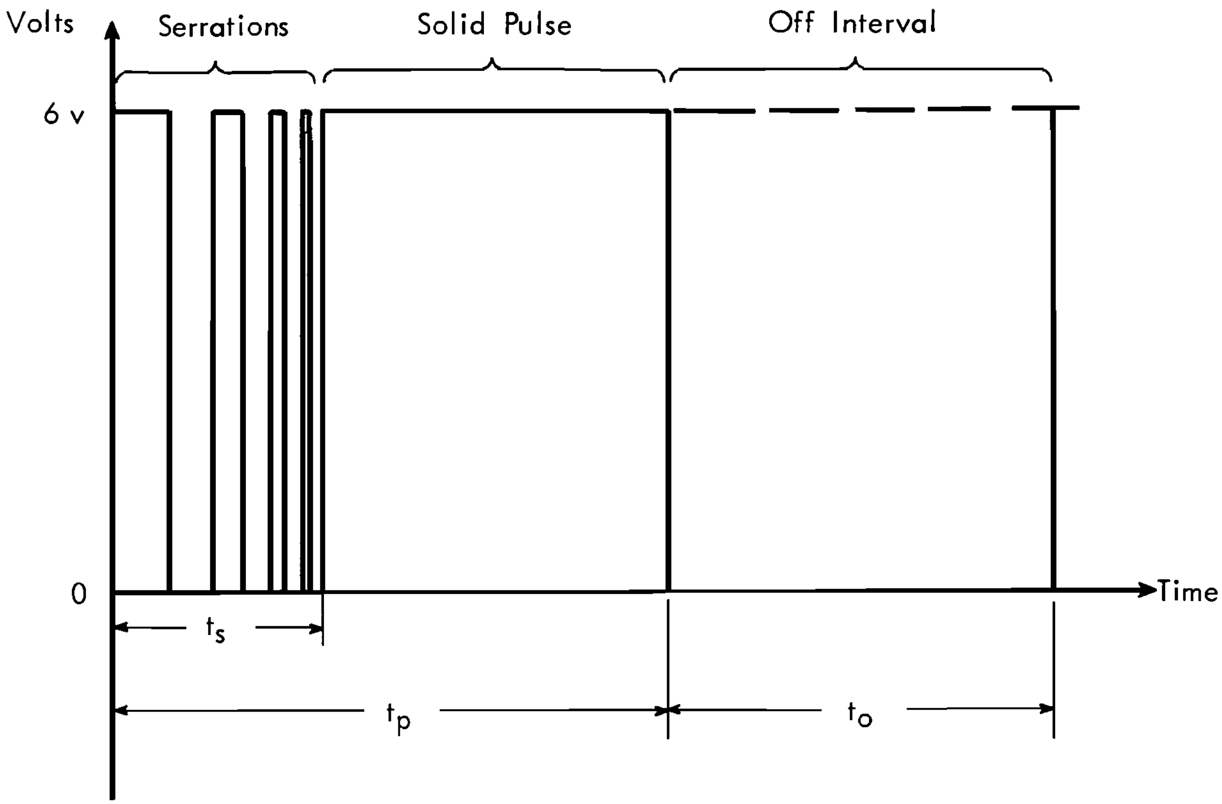

Fig. 14. Pulse analysis.

Table 1. The Pulse Composition for Various Meter Switches

Meter	t_s	t_p	t_o	Remarks
Finger pump	2.5 ms	Variable		Snap action micro-switch
Process condensate meter integrator	2-3 ms	1.8 sec	4.5 sec	Snap action micro-switch
Steam condensate meter integrator	4.5 ms	500 ms	5.5 sec	Snap action micro-switch
Wattmeter (to all segments of calciner heater)	25 ms	5 sec	9 sec	G.E. cam switch. Solid pulse not uniform. Also erratic serrations in trailing edge.
Calciner condenser water meter	2.5 ms	8 sec	60 sec	Snap action micro-switch
Process water meter (feed water to evaporator)	5.5 ms	250 ms	2 sec	Snap action micro-switch

width became narrower as the solid portion of the pulse was approached. The trailing end of each pulse was clean and solid except for the G. E. Wattmeter switch. The latter switch displayed unacceptable performance. The entire pulse was erratic and the trailing end exhibited serrations as well as other wave forms sometimes extending completely back into what otherwise would be the solid portion of the pulse. This switch also exhibited a time variable contact resistance. Repair contributed little to its overall performance.

The final experimental phase consisted of checking the circuit of Fig. 8 on the Pace 221-R analog computer at the University of Tennessee. This computer is a $\pm$ 100 v type with standard passive components consisting of 0.1 M and 1 M resistors and 1.0 μf and 0.1 μf capacitors. The comparators in this computer have a 1 ms throw time which probably represents an improvement over those in the TR-10. Also, the comparators in the TR-10 did not give a good comparison. They were off by about 2 volts and tended to go into oscillation at the comparison point. The latter difficulty, however, was eliminated by feeding to the comparison terminals through 0.1 M resistors.

On the Pace 221-R and with reference to Fig. 8, we used

$$R_1 = R_2 = 0.1 \text{ M}$$
$$C = 0.01 \ \mu\text{f}$$
$$C_1 = 10 \ \mu\text{f}$$

The R and R_o were omitted because e_o on the 221-R could be read directly on the DVM included in the computer console.

The Pace comparators are fused in lines A and B of Fig. 8 and these fuses blew as soon as the circuit was placed in operation. Investigation revealed that although on the patch board the two blades of the comparator are shown as being mechanically linked they are in fact two separate relays the coils of which are magnetically linked to the driving coil of the comparator. Because of the high speed of the comparator and absence of mechanical linkage between blades it was possible to have at the instant of throw the wrong combination of diode and reference voltage thus allowing sufficient surge current to ground which would blow a fast acting fuse.

The comparator in the circuit of Fig. 8 was replaced with an ordinary snap action relay the blades of which were mechanically linked. The relay coil was driven as shown in Fig. 15. This circuit performed excellently and without trouble. Pricewise this circuit is actually slightly less expensive than that using the comparator.

The circuit of Fig. 15 is capable of 2,000 pulses per swing at a rate up to 50 pps. Each pulse causes an incremental voltage change in e_o equal to

$$\frac{200}{2,000} = 0.1 \text{ volt or } 100 \text{ mv}$$

Thus, if one specified that not more than 1 pulse out of 2,000 shall be lost in one hour due to drift of the integrator this amounts to a drift of not more than 100 mv/hr. The manufacturer specifications on the Pace 221-R equipment specify drift not to exceed 3.6 mv/min with a 1 μf capacitor which would be 0.36 mv/min with a 10 μf capacitor. The latter amounts to 21.6 mv/hr which is of the order of 1/5 count/hr. A number of the amplifiers were actually checked for drift and it was found that their drift could be reduced to within specifications if the DVM were used to determine balance. If the light provided at each amplifier were used the drift varied from specifications up to a couple of counts out of 2,000 in one hour and at first this was thought to be due to the lower quality of the light as a null detecting device. Later it was traced to be a faulty ground connection in the computer itself.

11.0 CONCLUSIONS AND RECOMMENDATIONS

Of all the analog accounting circuits investigated, the most plausible one recommended for actual use in the calcination data logging operation is a simple one amplifier job designed based on transfer function simulation. With it one can count pulse frequencies of 200 cps with the upper limit probably being in the vicinity of 50 K cps which limit is certainly beyond the range of practical mechanical switching.

Evidence from experimentation indicated the presence of "contact bounce" in the particular switches investigated. By means of a simple RC filter network the serrations of the pulse can be omitted from the total count and it is recommended that this be done as a precautionary measure in connection with high speed counting.

In any event, the analog accounting circuits have within design limits been found completely reliable and free from "miss fires" or erratic

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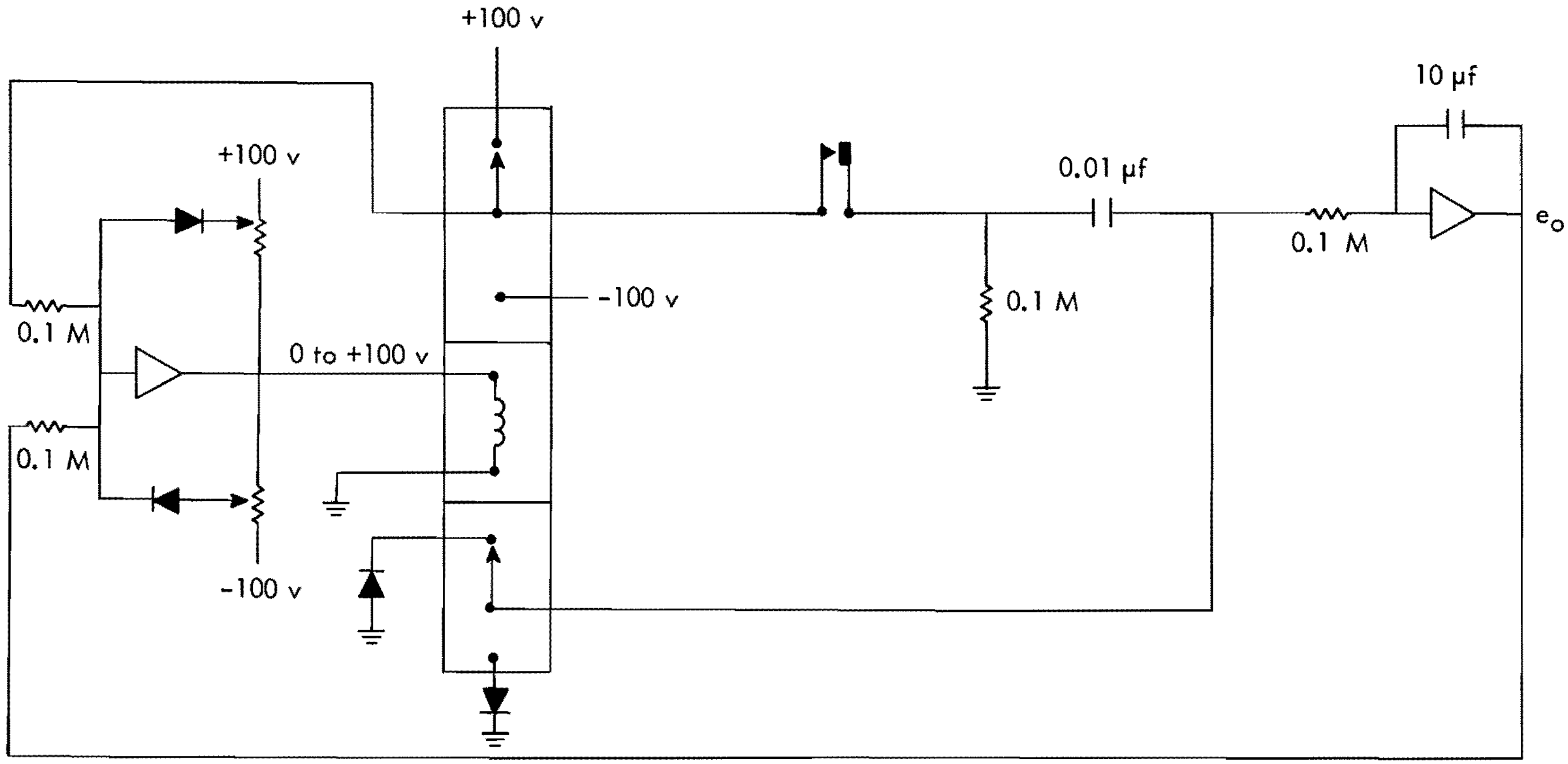

Fig. 15. Simplified accounting circuit utilizing a "Bang-Bang" with relay to replace comparators.

non-reproducible behavior. Drift requirements of less than 0.01 to 0.03 mv/hr at the data logging equipment have been met through proper choice of feedback capacitor (10 μf in the present case) around the analog accounting amplifier.

In purchasing analog equipment, it is imperative that the drift limitation on amplifiers be specified. It is also important that the blades of comparators or relays be mechanically linked to avoid overloading circuits during switching.

Whether or not one uses analog accounting in the calcination data logging operation is related to other alternatives as well as cost. In the case of fluid flow measurements, it is known that magnetic meters could be used to generate an a.c. millivoltage proportional to instantaneous flow rate. This a.c. millivoltage would have to be rectified and smoothed. With an amplifier and integrator integral flows could be transmitted for data logging. In the case of electrical power the situation is more subtle. One might, for example, use the recently developed silicon control rectifiers to replace relays associated with calciner heaters. The power supplied to the heaters would then be a continuously variable function of time. The average power, P, is

$$P = EI \cos \Theta$$

where

E = effective value of line voltage across heater load. This varies with time

I = effective value of current through heater load. This varies with time

With control rectifiers or saturable core reactors cos Θ would have to be measured and the output wave forms are such that E and I would have to be computed from the wave form at each control setting. Ultimately E, I, and cos Θ would have to be multiplied together and integrated.

From the cost standpoint alone it is obvious why analog accounting should be preferred.

Analog accounting as such does not imply the need for an analog computer since individual components can be purchased and assembled into a relay rack. However, if a small computer were being used in the calcination process for reasons other than accounting, then obviously it would be much cheaper to do the accounting on this computer.

The interesting fact is that analog computation is actually desirable in the automatic control of the evaporator in the calcination process. Here one measures boiling point and density with which to control water and feed liquor supply. What one is actually attempting to do in controlling water and feed liquid is to control the acidity and salt concentration in the evaporator. The boiling point and density are related to acidity and salt concentration in such a way that if an analog computer or components were incorporated one could prevent the evaporator operation from going out of control as it now does for excessive deviations from the set points.

Since analog computers themselves are available with digital readout, one could use such equipment to replace the present DCAT-digital logging combination in any new installation where logging is necessary.

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