Frequency-Sensitive Switching Circuit

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(Received June 18, 1956; revised version received September 11, 1956)

A frequency-sensitive switching circuit is described which automatically selects between two ranges of an otherwise conventional frequency meter by switching the leak resistor of the integrating circuit. The basis of the unit is a diode pump integrator ope-ated as a nonlinear device to obtain maximum sensitivity. The sensitivity is such that the "on-off" backlash of the switching relay corresponds to less than 1% of the switching frequency.

I. INTRODUCTION

FOR investigation of atmospheric potential gradients by a radiosonde method,[1–3] a frequency meter with automatic range selection was required for accurate telemetering of the sonde signals. In its present application the frequency-sensitive switching circuit selects between two ranges, 0 to 500 cps and 0 to 5000 cps, of an otherwise conventional frequency meter. The range selection is achieved by switching the leak resistor in the integrating network of the frequency meter.

The basis of both the switching circuit and the frequency meter is the diode pump integrator.[4] This integrator provides a most convenient way of obtaining an average dc potential from a train of voltage pulses. With certain limitations (normally applied) this potential is approximately a linear function of frequency, enabling the integrator to be used as a frequency meter,[5] random counting rate meter,[6] or as a device for the demodulation of frequency modulated radio waves.[7] However, the utility of the circuit is not limited entirely to linear applications and it is the nonlinear characteristic which is utilized in the switching circuit.

II. DIODE PUMP INTEGRATOR

Figure 1 shows the circuit of the integrator and it is assumed that each incoming pulse is a rectangular pulse as depicted. The amplitude E and the duration T of these pulses are standardized.

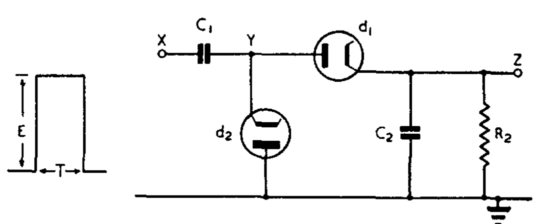

FIG. 1. Circuit of the diode pump integrator.

With a constant pulse repetition rate a quasi-steady state must eventually be reached when the charge conducted to C_2 by d_1 just replaces that leaking away through R_2. When this occurs the output voltage wave form at Z is approximately sawtooth about an average dc potential v.

If $1/n$ is the interval between pulses and $C_2 \gg C_1$, the nonlinear relationship between v and n is

$$v = E \cdot \frac{nC_1R_2}{1+nC_1R_2}. \qquad (1)$$

Normally[8,9] the circuit is operated with $v \ll E$ so that a linear relationship exists where

$$v = EnC_1R_2. \qquad (2)$$

In both cases the peak to peak ripple voltage is

$$(E-v)C_1/C_2. \qquad (3)$$

The nonlinear relationship between v and n, formed by Eq. (1) is shown graphically in Fig. 2. As the values of C_2 and R_2 can be chosen arbitrarily a family of curves can be drawn with the product R_2C_1 as parameter.

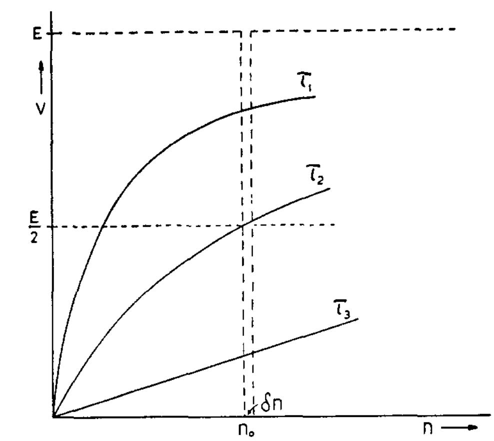

FIG. 2. Family of curves showing $v = En\tau/(1+n\tau)$ for different values of the parameter τ.

[1] K. Kreielsheimer and B. Belin, Nature **157**, 227 (1946).
[2] K. Kreielsheimer, Australian J. Sci. **9**, 95 (1946).
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[6] W. C. Elmore, Nucleonics **2**, No. 4, 43 (1948).
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[9] W. C. Elmore and M. Sands, *Electronics* (McGraw-Hill Book Company, Inc., New York, 1949).

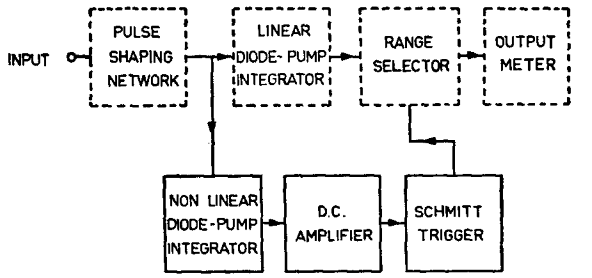

FIG. 3. Block diagram of the frequency meter incorporating the frequency-sensitive switch as a range selector.

Maximum sensitivity at a specified switching frequency corresponds to a maximum voltage change per cycle at that frequency. As the components C_1 and R_2 occur in product form in Eq. (1) they can be replaced by τ where $\tau = R_2C_1$. Hence Eq. (1) becomes

$$v = E\frac{n\tau}{1+n\tau}. \tag{4}$$

The value of τ which makes $\partial v/\partial n$ a maximum for a given value of n, say n_0, is required and since $\partial v/\partial n = g(\tau)$ this maximum will occur when

$$\partial^2 v/\partial\tau\partial n = 0. \tag{5}$$

Equation (5) is satisfied by the solution

$$\tau = 1/n_0. \tag{6}$$

Therefore the maximum value of $\partial v/\partial n$ for a specified frequency is $E/4n_0$ and this maximum exists when the equilibrium voltage v is equal to half the incoming pulse amplitude.

III. RANGE SWITCHING CIRCUIT

Figure 3 shows a block diagram of the switching circuit designed for use as a range selector in an otherwise conventional frequency meter. The switching circuit consists of a diode pump integrator whose amplified output controls a Schmitt trigger[10] in which the actual switching relay is incorporated.

With 70-v pulses from the pulse shaping network the maximum value of $\partial v/\partial n$ obtainable when $n_0 = 500$ cps is $70/(4\times 500)$ v/cy, so that for a frequency change of 3 cps the integrator output variation is approximately 0.1 v. Because this change is too small to operate a Schmitt trigger directly a dc amplifier (gain factor 60) is connected between the integrator and the trigger.

The trigger design is such that the relay is "on" whenever the input to V_2 (Fig. 4) falls below 100 v and "off" when the input rises above 106 v. Backlash in the trigger is essential to prevent hunting and this value of 6 v, corresponding to a 3-cps change of input frequency, i.e., to a variation of less than 1% of the switching frequency, is the minimum that can be employed for satisfactory operation.

[10] O. H. Schmitt, J. Sci. Instr. 15, 24 (1938).

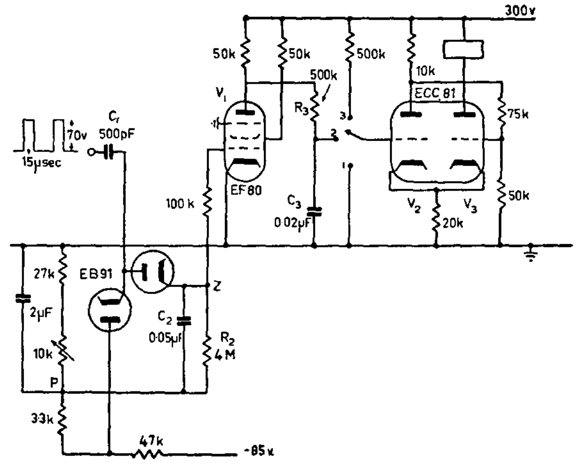

FIG. 4. Practical circuit of frequency-sensitive switch (values for switching frequency of 500 cps).

The diode pump integrator is connected to the negative supply line at P where the potential of P is such that, with the addition of 35 v across C_2 ($E/2$ at 500 cps), the dc amplifier V_1 is biased to amplify in its linear region. The anode voltage of V_1 can be adjusted, by the potentiometer in the bias line, so that with an input of 500 cps the trigger "fires." A resistor R_3, between V_1 anode and V_2 grid ensures that when V_1 is cut off the grid current in V_2 is not excessive. Manual control of the trigger can be achieved by the selector switch at V_2 grid. The three positions are (1) on, (2) automatic, (3) off.

In the nonlinear integrator itself the value of C_1 is determined by the loading on the pulse shaping network and R_2 is then determined by Eq. (6) and the switching frequency (500 cps). With $C_2 = 100C_1$ the ripple voltage is 0.35 v at Z [(see Eq. (3)] and 21 v after amplification; so that it is more than three times the backlash of the trigger. To decrease the ripple by increasing C_2 means increasing the response time R_2C_2 by the same factor. However, as the ripple frequency is 500 cps its magnitude can be reduced to 0.6 v at the input to the trigger by C_3 connected between V_2 grid and earth. The presence of this capacitor introduces another time-constant R_3C_3 (0.01 sec) but this is negligible compared with R_2C_2 (0.2 sec).

Once the conditions $C_2 \gg C_1$ and $R_2C_2 \gg 1/n_0$ have been satisfied the actual value of C_2 is determined by a compromise between the ripple that can be tolerated and the transient response required. In the application for which the present switching circuit was designed the transient response is such that for any suddenly impressed frequency between 0 and 5000 cps the needle on the frequency meter never "hits the stop" while still on the 0 to 500 cps range.

If the time constants of the linear integrator and recording meter are t_1 and t_2, respectively, this means that $R_2C_2 < (t_1^2 + t_2^2)^{\frac{1}{2}}$ thus specifying an upper limit for C_2.

Any drift in the dc amplifier output causes a variation of the switching point. In the present application the actual switching point is not too critical. However after an initial warming up period the drift is less than 5 cps at any time during a 5-hr interval. For more critical conditions the amplifier can be compensated by any of the usual methods for eliminating drift.[11]

Further applications of the frequency-sensitive switch are contemplated, one of them using a series of switching circuits set at different frequencies to produce an automatic multirange counting rate meter for particle counting.

[11] J. B. Earnshaw, Electronic Eng. (to be published).

IV. ACKNOWLEDGMENTS

The above work was carried out as part of a project for measuring potential gradients in the earth's atmosphere by radiosonde methods. The author would like to thank the University of New Zealand Research Grants Committee who supported the work and to acknowledge the helpful criticism and encouragement given by Dr. K. Kreielsheimer, whilst the work was in progress.

THE REVIEW OF SCIENTIFIC INSTRUMENTS VOLUME 27, NUMBER 12 DECEMBER, 1956

Slide Rule for Radiographic Analysis

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(Received June 22, 1956; revised version received August 24, 1956)

A simple slide rule has been developed which will permit a rapid and accurate determination of radiographic sample transmission, of absorption index, μm, and of the photographic film density corresponding to the direct x-ray beam by means of a single setting of the sliding scale.

In this laboratory, as in many others where microradiographic techniques are employed, a considerable amount of time is often required for the determination of film densities, $D_{\text{direct beam}}$ and D_{sample}; of sample transmissions, t; and of the absorption index, μm, as calculated from the microphotometer readings on the radiogram, $T_{\text{background}}$, T_{sample}, and $T_{\text{direct beam}}$. With the usual assumptions that the reciprocity law holds and that the film densities are proportional to exposure for the given density range, the mathematical relationships which are involved in these calculations can be written as follows:

$$D_s/D_d = I_s/I_d = e^{-\mu m} = t \qquad (1)$$

in which I and I_d are incident and transmitted x-ray intensities, respectively, and in which, by definition,

$$D_s = \log_{10}(T_b/T_s) \quad \text{and} \quad D_d = \log_{10}(T_b/T_d). \qquad (2)$$

Combining these equations, one obtains

$$t = \frac{\log_{10}T_b - \log_{10}T_s}{\log_{10}T_b - \log_{10}T_d} = e^{-\mu m}. \qquad (3)$$

In these equations μ is the mass absorption coefficient and m is the mass per unit area of the sample.

A slide rule has been designed that permits an easy and rapid determination of the values of μm, t, and D_d by means of a single setting of a sliding scale which is constrained to translate only along its length or perpendicular to this direction over a two-dimensional, stationary scale. A photograph of the instrument is shown in Fig. 1. It has been mounted on a standard two-by-three-foot drafting board of the type which has a straight-edge that is constrained by a string-and-pulley system to remain parallel to the bottom edge of the board.

The design of the scales is illustrated in Fig. 2. The transparent sliding scale b, on which the photometer readings are entered, is logarithmic to base 10. Consequently, the length XY is proportional to D_s since

$$D_s = \log_{10}T_b - \log_{10}T_s$$

and similarly, the length XZ is proportional to D_d since

$$D_d = \log_{10}T_b - \log_{10}T_d.$$

Scale c is linear and is of range 0 to 1. Hence, by projecting the lengths XY and XZ onto scale c, as shown in Fig. 2, the ratio (XY/XZ) is read directly and, by Eq. (3), is equal to the sample transmission, t.

Scale a is constructed alongside scale c in such a manner as to give the value of the absorption index, μm, that corresponds to the particular value of sample transmission, t, as read directly above on scale c. Hence, scale lengths measured left-to-right on scale a must be proportional to the value, $\exp(-\mu m)$, which is equal to t as defined in Eq. (3).

The lines which project the lengths XY and XZ onto scale c emanate from a point, 0′, directly below zero

* Department of Physics, Pomona College, Claremont, California. This work was part of a program conducted while on sabbatical leave at the Karolinska Institutet as a Guggenheim Fellow.