The Single-Ended Diode Phase-Sensitive Detector

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The operation of the single-ended diode phase-sensitive detector with load is investigated. As in the case of the simple push-pull detector, the transfer ratios for the two diodes are found to vary considerably with the signal. This introduces a non-linearity in the output which is evaluated and a table is given from which the performance of a given detector of this type may be judged immediately. A comparison is made between this detector and the simple push-pull detector and the loading conditions, under which one is superior to the other from the point of view of linearity, are discussed.

(Voir page 192 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 200)

IN a recent article[1] (hereafter referred to as (I)), the authors investigated the operation of the simple push-pull phase-sensitive detector under various degrees of loading. In this article, a similar investigation is carried out of another type of detector which, in spite of the important advantage of providing a single-ended output, does not appear to be well-known.

The Single-Ended Phase-Sensitive Detector

Fig. 1 shows the circuit of the detector. E_1 and E_2 are the amplitudes of the reference and signal voltages respectively, assumed to be in phase. The sum of the two voltages is applied to one of the diodes, MR_1, and the difference to the other diode, MR_2. MR_1 and MR_2 are connected in opposite sense. R_1's are the effective series resistances in the two diode circuits. If the value of CR_2 is very large compared to the period of the applied sine waves, the voltages E_A and E_B developed at the nodes A and B may be assumed to be steady.

If the transfer ratios of the two diodes are k_1 and k_2, then:

$$E_A = k_1(E_1 + E_2) \text{ and } E_B = -k_2(E_1 - E_2).$$

Suppose that the diodes conduct respectively over angles $2\theta_1$ and $2\theta_2$ during a cycle; their transfer ratios can be written as:

$$k_1 = \cos\theta_1,\ k_2 = \cos\theta_2 \qquad \text{(1)}$$

The output voltage is given by:

$$E_o = \frac{R}{R_2 + 2R} \cdot \{k_1(E_1 + E_2) - k_2(E_1 - E_2)\} \qquad \text{(2)}$$

As will be shown below, k_1 and k_2 are in general unequal so that the output is not proportional to the signal amplitude E_2.

The average current flowing through MR_1 into node A is

$$\frac{\pi R_1}{E_1 + E_2}\{\sin\theta_1 - k_1\theta_1\}.$$

This is equal to the current flowing through R_2.

$$\frac{E_1 + E_2}{\pi R_1}\{\sin\theta_1 - k_1\theta_1\} = \frac{k_1(E_1 + E_2) - E_o}{R_2} \qquad \text{(3)}$$

Using equations (1) and (2), the above equation and the corresponding one for node B can be rewritten as:

$$n_1/\pi\ (\sin\theta_1 - \theta_1\cos\theta_1) = \cos\theta_1(1 - n_3) + wn_3\cos\theta_2 \qquad \text{(4)}$$

$$n_1/\pi\ (\sin\theta_2 - \theta_2\cos\theta_2 = \cos\theta_2(1 - n_3) + (n_3/w)\cos\theta_1,\ w \neq 0 \qquad \text{(5)}$$

where:

$$n_1 \equiv (R_2/R_1),\ n_3 \equiv \frac{1}{2 + n_2},\ n_2 \equiv (R_2/R),\ w \equiv \frac{1 - x}{1 + x},\ x \equiv (E_2/E_1) \qquad \text{(6)}$$

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As in (I), n_1 and n_2 may be called the *source impedance factor* and the *load factor* respectively. Equations (4) and (5) may be solved by the method described in the Appendix of (I). (The corresponding equations in (I) are equations (9) and (10).

Behaviour of the Transfer Ratios k_1 and k_2

As in (I), the calculations have been made for the source impedance factor $n_1 = 1{\cdot}46\pi$. Fig. 2 gives the variations of k_1 and k_2 against the signal-to-reference voltage ratio x for different values of the load factor n_2. It will be observed that in the absence of the signal (i.e., $x = 0$), k_1 and k_2 are equal for all values of the load factor n_2. This follows from symmetry considerations, the voltages at the nodes A and B in the absence of the signal being equal in magnitude and opposite in sign leading to an output voltage E_o equal to zero. The common value, say $k_o \equiv \cos\theta_o$, of the transfer ratios k_1 and k_2 corresponding to $x = 0$ can be obtained from equation (3) by putting $E_2 = E_o = 0$. θ_o, the semi-angle of conduction of either diode in the absence of the signal, is consequently the root of the equation:

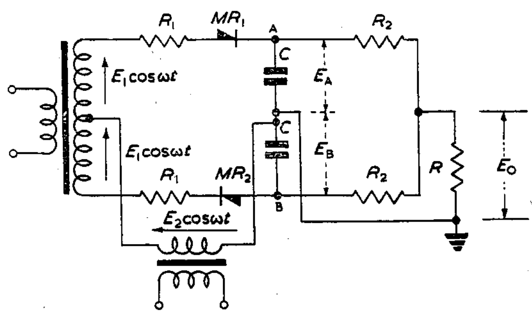

Fig. 1. Circuit of the single-ended phase-sensitive detector

$$\frac{\pi\cos\theta}{\sin\theta - \theta\cos\theta} = (R_2/R_1) = n_1 \qquad \text{(7)}$$

For any value of the load factor n_2, as the signal voltage (or x) is increased, k_1 increases and k_2 decreases continuously. This means that the angle of conduction of the diode MR_1 to which the sum of the reference and signal voltages is applied decreases and the angle of conduction of the other diode MR_2 increases. As the latter becomes equal to 180°, the transfer ratio k_2 becomes equal to zero. For higher values of x, the conduction angle increases further leading to a negative value for k_2. When k_2 becomes equal to -1, the diode MR_2 begins to conduct over the entire cycle. Beyond this k_2 loses its significance in terms of the conduction angle of MR_2. The deviations of k_1 and k_2 from the no-signal value k_o are more pronounced for lower values of n_2 (i.e., larger load resistances).

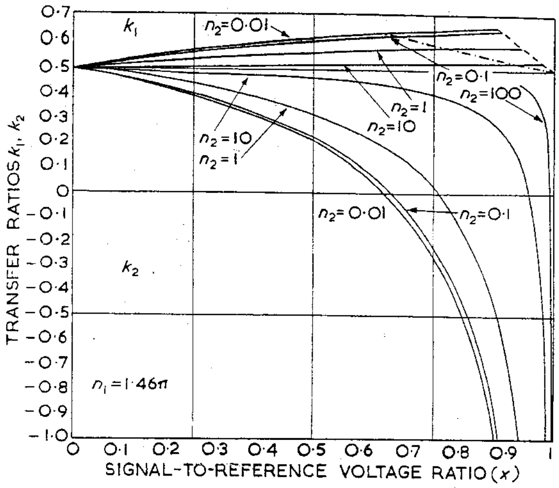

Fig. 2. Variation of the transfer ratios of the two diodes with the signal

The chain and the broken lines are the loci of k corresponding respectively to $k_2 = 0$ and -1 ($\theta_2 = 90°$ and $180°$)

Non-Linearity of the Detector

From equations (2) and (6), it can be seen that the ratio of the output d.c. voltage E_o to the signal voltage amplitude E_2, which may be called the *sensitivity* of the detector (as in (I)), is given by:

$$\alpha \equiv (E_o/E_2) = \frac{1}{n_2+2}\left\{k_1 + k_2 + \frac{k_1 - k_2}{x}\right\} \quad \ldots (8)$$

As a criterion of the non-linearity of the detector, one may consider the percentage change in the sensitivity α in the range $0 < x < x_{crit}$, where x_{crit} is the point where the conduction angle of MR_2 becomes 180° or k_2 becomes equal to zero. The total non-linearity of the detector, N_o, may be defined as:

$$N_o \equiv 100 \cdot \frac{\alpha_o - \alpha_{crit}}{\alpha_o} \text{ per cent} \quad \ldots\ldots\ldots\ldots (9)$$

where α_o is the limiting sensitivity obtained as $x \to 0$ and α_{crit} is the sensitivity at $x = x_{crit}$.

Even though $k_1, k_2 \to k_o$ as $x \to 0$, α_o is not equal to $2k_o/(n_2+2)$ because $(k_1-k_2)/x$ tends to a finite positive limit. This limit can be evaluated theoretically by the method described in the Appendix of (1) to give:

$$\alpha_o = \frac{2k_o}{n_2 + 2 - \dfrac{2}{((n_1\theta_o/\pi) + 1)}} = \frac{2k_o}{n_2 + 2 - (2\pi/n_1)\cot\theta_o} \quad \ldots\ldots\ldots\ldots (10)$$

where, as mentioned previously, θ_o is the semi-angle of conduction of either diode in the absence of the signal and is the root of the equation (7).

To evaluate α_{crit}, it may be noted that at $x = x_{crit}$, the node B is at ground potential, so that the node A can be considered connected to ground through a resistance R_{eq}, where

$$R_{eq} = R_2 \cdot \frac{R_2 + 2R}{R_2 + R} = R_2 \cdot \frac{n_2 + 2}{n_2 + 1}$$

The semi-angle of conduction $\theta_{1(crit)}$ of the diode MR_1 at $x = x_{(crit)}$ can therefore be determined from equation (7) after substituting R_{eq} for R_2. The corresponding value of the transfer ratio, $k_{1(crit)}$, is $\cos\theta_{1(crit)}$.

Now, from equation (5), by putting $\theta_2 = \pi/2$,

$$w_{crit} = \frac{1 - x_{crit}}{1 + x_{crit}} = \frac{\pi}{n_1(n_2+2)} k_{1(crit)}$$

Hence, from equation (8), remembering that $k_{2(crit)} = 0$,

$$\alpha_{crit} = \frac{2k_{1(crit)}}{n_2 + 2 - (\pi k_{1(crit)}/n_1)} \quad \ldots\ldots\ldots (11)$$

To determine the semi-angles of conduction, θ_o and $\theta_{(crit)}$, from equation (7), a graphical method is used and the approximate values so obtained are refined using Taylor's theorem. Values of α_o, x_{crit} and N_o are given in Table 1 for various values of n_1 and n_2 and the suitability of a detector may be judged immediately from this table.

Comparison of the Single-Ended and the Simple Push-Pull Detectors

On comparing the table given in this article with the table given in (I), it is observed that for a given value of n_1, the non-linearity of the single-ended detector increases as the load resistance is increased while the reverse is the case for the simple push-pull detector. In fact, the single-

TABLE 1

n_1 \ n_2	100	30	10	3	1	0·3	0·1	0·03	0·01
1 ..	0·00432 0·987 —	0·0142 0·957 0·05%	0·0410 0·886 0·15%	0·121 0·726 0·40%	0·271 0·542 0·70%	0·482 0·400 0·95%	0·619 0·350 1·10%	0·688 0·319 1·10%	0·710 0·312 1·15%
1·46π ..	0·00988 0·994 0·05%	0·0320 0·979 0·15%	0·0892 0·943 0·40%	0·238 0·861 1·10%	0·453 0·767 1·85%	0·663 0·693 2·45%	0·764 0·663 2·65%	0·807 0·650 2·75%	0·820 0·645 2·75%
10 ..	0·0127 0·996 0·05%	0·0410 0·987 0·20%	0·133 0·964 0·55%	0·289 0·918 1·25%	0·523 0·862 2·10%	0·730 0·818 2·70%	0·823 0·800 2·85%	0·861 0·792 2·95%	0·873 0·790 3·00%
30 ..	0·0158 0·998 0·05%	0·0506 0·995 0·20%	0·137 0·986 0·45%	0·340 0·966 1·10%	0·590 0·943 1·80%	0·794 0·925 2·25%	0·882 0·917 2·45%	0·917 0·914 2·45%	0·928 0·914 2·50%
100 ..	0·0177 1·000 0·05%	0·0567 0·998 0·10%	0·152 0·995 0·30%	0·371 0·989 0·75%	0·630 0·981 1·15%	0·834 0·975 1·50%	0·918 0·972 1·60%	0·952 0·971 1·65%	0·962 0·971 1·65%

The set of three figures in each square gives in that order the limiting sensitivity α_o, rounded off to three significant figures, the value of x, x_{crit}, where the diode MR_2 starts conducting over an angle 180° and the total non-linearity in the range, $0 < x < x_{crit}$

ended detector is absolutely linear when the output terminal is shorted to ground, while the simple push-pull detector is absolutely linear when the output terminals are open-circuited. This leads to the important design criterion that the simple push-pull detector should be used when the output feeds, for example, a high input impedance valve-voltmeter. On the other hand when it is a low impedance current measuring instrument that is available, the single-ended detector should be chosen.

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1. Chidambaram, R., Krishnan, S. Diode Phase-Sensitive Detectors with Load, *Electronic Engng. 31*, 613 (1959).

An Events Per Unit Time Meter (EPUT Meter)

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This article describes a simplified form of Events Per Unit Time Meter (EPUT meter), which gives a read-out in cycles per second up to 9 999. The 1sec gating waveform is derived from a 1kc/s crystal oscillator having a long term stability not worse than one part in 10^4. A separate output is provided from this oscillator for checking the accuracy of count, and for use as an external calibrator. Both manual and automatic resets are provided, enabling the instruments to be used as a straight counter if required.

(Voir page 192 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 200)

THE Events Per Unit Time Meter (EPUT meter) to be described was designed to meet the need for a cheap instrument counting up to 9 999 impulses per second. To give a direct read-out in cycles per second the sampling period is fixed at one second, after which there is a read-out period of three quarters of a second before the automatic reset operates and the cycle is repeated. The 1sec sampling waveform is derived from a built-in 1kc/s oscillator having a stability not worse than 1 part in 10^4, and the count is therefore accurate to within one cycle per second. A sinusoidal output from the crystal oscillator is brought out on a jack for external use, this can be linked to the input jack to check the calibration of the counters. A manual reset is provided as an alternative to the automatic one, and the 1sec gate can be switched out enabling the instrument to be used as a straight counter. A sinusoidal input of not less than 1V r.m.s. is required.

General Description

A schematic diagram of the unit is given in Fig. 1, and the circuit in Fig. 2.

The crystal oscillator uses a G.E.C. 1kc/s bimorph crystal having a low temperature coefficient as is shown by the typical frequency temperature characteristic, Fig. 3.

The crystal operates as a three terminal network with the common point taken to earth via C_1. Variation of this capacitor enables the crystal to be trimmed over about 50 parts in 10^6. The two remaining terminations are taken to the grid and anode circuits of the pentode oscillator V_1. Approximately one fifth of the total output is developed across R_2 and applied as a drive to the crystal, the full output is via C_5 to the amplifier stage V_2, automatic gain control being effected by rectifying a portion of this output and applying the filtered d.c. to the grid of V_1. A tuned circuit is included in the cathode of V_2 from which a sinusoidal output is taken, the amplified signal developed across R_8 is fed via a phase shifting network to the guides of the Dekatron V_3.

V_3 to V_7 comprise a conventional Dekatron dividing chain having a divide factor of 1 000 thus producing at the anode of V_{6b} a one per second impulse. This is applied simultaneously to each grid of the double triode V_8 which is connected to form a binary cell. The circuit is balanced by RV_{41} so that when a pulse is applied to the grid of the triode which is conducting it becomes momentarily less conducting, amplification occurs in the triode which was cut off; and the system jumps to its other stable condition. This change takes place almost instantly resulting in a square wave output voltage in the anode circuits the polarity of which reverses with each triggering pulse. This square wave of one second duration is fed via C_{23} to the buffer amplifier V_{9a} and also via R_{46} to the reset pulse generator V_{10}.

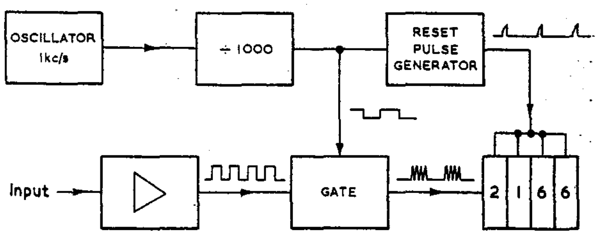

Fig. 1. General arrangement of EPUT meter

The square wave developed across R_{47}, the anode load of V_{9a}, is applied to the suppressor grid of the gating valve V_{12} via a small neon lamp V_{13}. The suppressor is also connected via R_{66} to a negative voltage, which prevents the valve from conducting. During the 1sec when the square wave at the anode of V_{9a} is positive, the neon strikes and the suppressor is driven positive allowing the valve to pass impulses to the counters.

In order to reset the counters it is necessary to produce a reset impulse every 2sec phased as shown in Fig. 4, allowing a reading time of approximately 0·75sec before resetting to zero ready for the next counting period.

This is achieved by feeding the square wave developed across R_{43} to the integrating circuit R_{49}, C_{24} and thence to the grid of V_{10a}. V_{10} is a biased flip-flop the operating level of which—and hence the phasing of the ouput pulse—is controlled by a variable cathode resistance, the effect of which