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EVALUATION OF PHASE DELAY IN A CRYSTAL MIXER

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EVALUATION OF PHASE DELAY IN A CRYSTAL MIXER

J. C. Sanderlin and J. W. Cook

Introduction:

In systems where the phase delay is the measured variable, any phase shift through the system appears as a fixed error in the measured variable, and any variations in phase shift through the system appear as uncertainties in the measured variable. The phase delay characteristics of a mixer when used in such a system are thus of considerable interest. This report presents the results of an analytical and empirical investigation of the phase delay characteristics of a crystal mixer.

The phase shift incurred by a sinusoidal wave in passing through a crystal mixer is treated from a casual viewpoint both analytically and empirically. The results obtained from these two approaches are in good agreement with respect to the form of phase shift as a function of the controlled variables, while fair quantitative agreement between the two approaches is obtained.

From the results of this investigation, an operating condition is obtained that minimizes the change in phase shift through a crystal mixer for reasonable variations in the controlled variable.

Statement of the Problem:

The crystal mixer is a three port device consisting of a set of linear circuit elements plus the non-linear crystal diode. The crystal mixer to be studied is self-biased, that is, the d-c bias for the crystal diode is derived from the input signals. For the purpose of analysis, it is assumed that the signal input level is very much less than the local oscillator input level. (This is also generally true in practice.) Under this assumption, it follows that the d-c bias for the crystal diode is derived almost entirely from the local oscillator input.

The operating point of the diode is defined in terms of the d-c bias current. By the argument of the preceeding paragraph, the operating point may also be expressed in terms of the local oscillator input level. The phase shift incurred by a signal in passing through a circuit is determined by the signal frequency, the circuit arrangement, and the values of resistance and reactance in each mesh of the circuit. In the case of a crystal mixer, all of the circuit values are constant with the exception of the diode. The equivalent impedance of the diode varies with d-c bias. The bias is derived from a sinusoidal source, i.e., the local oscillator input, hence the d-c bias, and thus the diode impedance would be expected to vary sinusoidally, or at worst periodically. Under these conditions, an average diode impedance can be defined, such that, for constant frequency and amplitude of local oscillator input level, the average diode impedance is a constant. The operating point of a mixer is then defined in terms of the steady-state, or average, value of the diode resistance.

For a given diode circuit, constant signal and local oscillator frequencies, as well as constant amplitude of the local oscillator input to the mixer, the phase shift should be constant and hence, its effect may be compensated by proper input and/or output matching networks. Note, however, that any variation in the local oscillator input level to the mixer results in a change in the diode impedance and hence changes the phase shift so that it is no longer compensated by the matching networks.

Relative Phase of Signals of Different Frequencies:

The mixer is a three port device in which r-f energy is applied to two ports and r-f energy is extracted at the third port. A characteristic feature of the mixer is that the frequency of the output energy is equal to the sum and/or difference of the frequencies of the input energy. Let the input signals to the mixer $E_1(\omega_1,t)$ and $E_2(\omega_2,t)$ be defined by

1) $$E_1(\omega_1,t) = E_1 \cos (\omega_1 t + \phi_1)$$

and

2) $$E_2(\omega_2,t) = E_2 \cos (\omega_2 t + \phi_2)$$

where $\omega_1 \neq \omega_2$.

Since the mixer is a product device, the output signal $E_3(\omega_3,t)$ is defined by

3) $$E_3(\omega_3,t) = E_1(\omega_1,t) \cdot E_2(\omega_2,t) \; .$$

Substitute (1) and (2) into (3) and use the trigonometric identity

4) $$\cos A \cos B = \tfrac{1}{2} \Big[\cos (A + B) + \cos (A - B) \Big] \; ,$$

to obtain

$$5)\qquad E_3(\omega_3,t) = (E_1E_2/2)\Big\{\cos\Big[(\omega_1 + \omega_2)t + (\phi_1 + \phi_2)\Big] + \cos\Big[(\omega_1 + \omega_2)t + (\phi_1 + \phi_2)\Big]\Big\}.$$

Use of a high-pass or low-pass filter permits the selection either of the frequency sum term or the frequency difference term respectively as the desired output signal.

In considering the phase shift through a mixer the quantity of interest is the difference in phase between the input and output signals. These signals are of different frequencies, hence it is of interest to consider the conditions under which the phase difference between signals of different frequencies has significance.

Consider two signals $E_A(\omega,t)$ and $E_B(\omega,t)$ where

$$6)\qquad E_A(\omega,t) = E_a \sin(\omega_a t - \phi_A)\ , \quad -\pi < \phi_A < \pi\ ,$$

and

$$7)\qquad E_B(\omega,t) = E_b \sin(\omega_b t - \phi_B)\ , \quad -\pi < \phi_B < \pi\ , \quad \omega_b > \omega_a\ .$$

The term "phase difference" between signals of different frequencies is ambiguous, since a phase implies a frequency and in the case of two frequencies the phase difference may be referred to either of the frequencies. To eliminate this ambiguity consider a "time difference" between the two waves. In considering a time difference no frequency is implied, hence no ambiguity results. The time delay between two periodic waves may be described in terms of the occurence times of any two like points on the two waves. Select the positive-going zero crossings of the waves as reference points.

In the (ω,t) plane, equation (6) represents a sinusoid displaced from the origin by an amount $\omega_a t = \phi_A$. Make a transformation to the (ω,T_a) plane where T_a is such that $E_a(\omega,T_a)$ is not displaced from the origin. In the (ω,T) plane, equations (6) and (7) are of the form

8) $$E_A(\omega,T_a) = E_a \sin \omega_a(t - \phi_A/\omega_A) = E_a \sin \omega_a T_a \quad ,$$

and

9) $$E_B(\omega,T_b) = E_b \sin \omega_b(t - \phi_B/\omega_b) = E_b \sin \omega_b T_b$$

where $T_a = (t - \phi_A/\omega_a)$ and $T_b = (t - \phi_B/\omega_b)$.

The positive-going zero crossings of $E_A(\omega,T)$ and $E_B(\omega,T)$ occur for

10) $$\omega_a T_a = (2n)\,\pi, \quad n = 0, 1, 2, 3, \ldots ,$$

and

11) $$\omega_b T_b = (2m)\,\pi, \quad m = 0, 1, 2, 3, \ldots$$

respectively.

In the (ω,T) plane, the initial time delay between E_A and E_b is zero. Also, the time delay is defined only at those times when E_A and E_B have simultaneous, positive-going zero crossings. That is, for $T_a = T_b$, hence from (10) and (11), the statement that $T_a = T_b$ implies

12) $$n/m = \omega_a/\omega_b \quad .$$

From the definition of T_a and T_b (equations 8 and 9), the statement that $T_a = T_b$ implies

13) $$T_a/\omega_a - T_b/\omega_b = t_a - t_b$$

where t_a and t_b are the occurrence times of the positive-going zero crossings of $E_A(\omega,t)$ and $E_b(\omega,t)$ respectively in the (ω,t) plane. Hence, $t_a - t_b$ is just the time delay in question. Equation (13) may be written in the form

14) $$\Delta t = T_a/\omega_a - T_b/\omega_b$$

The first coincident zero crossings in the $E(\omega,T)$ plane occur for m and n = 0. The second coincident zero crossings occur for the smallest integer m such that $m(\omega_a/\omega_b)$ is integral. For example, if $\omega_a/\omega_b = 1.2$, then the second coincidence of zero crossings occurs for n = 6 and m = 5. It should be clear that for equation (12) to be defined, ω_a/ω_b must be a rational fraction. From (12) it follows that

15) $$d\phi_a(t)/d\phi_b(t) = n/m$$

Hence, for any two signals of different frequencies, where the frequencies satisfy (12), there is defined a time delay between the two waves having the form of (13).

Refer to equation (5) in which $E_3(\omega,t)$ is the mixer output signal, and $E_1(\omega,t)$ is the mixer signal input, as defined in (1). From (1), (5) and (12), the time delay through the mixer is defined only if ω_1 and ω_2 are related by

16) $$\omega_2/\omega_1 = (m - n)/m$$

It should be noted that (12) and (15) are the conditions for coherence between two waves having different frequencies.

Analysis of a Crystal Diode Mixer:

The d-c volt-ampere characteristic for a crystal diode is presented in Figure 1. where the r-f signal is shown superimposed on the local oscillator signal. From the diode characteristic it is clear that the diode impedance is a function of the applied signal voltage. The applied signal voltage is of the form

17) $$e(t) = V_L \cos \omega_L t + V_s \cos \omega_s t$$

where $e_L(t) = V_L \cos \omega_L t$,

$e_s(t) = V_s \cos \omega_s t$,

V_L = amplitude of the local oscillator signal,

V_s = amplitude of the r-f signal,

ω_L = angular frequency of the local oscillator signal,

and ω_s = angular frequency of the r-f signal.

In general $V_L \gg V_s$, so that (17) may be written

18) $$e(t) \cong V_L \cos \omega_L t \ .$$

For a periodic applied voltage, the functional relationship between the diode admittance and applied voltage implies that the diode admittance is also periodic. Hence, the diode admittance may be represented by the Fourier series

19) $$Y = a_o/2 + \sum_{n=1}^{\infty} a_n \cos (n \, \omega_L t)$$

where the a_n are determined by the diode characteristics.

Suppose in Figure 2a the diode admittance is very much less than the other circuit admittances. Then, the signal current is determined by the diode admittance, and the diode current is given by

20) $$i_d = Y \, e_s(t) \ .$$

Substitute from (17) and (19) into (20) to obtain

$$21)\qquad i_d = (a_o/2)\ V_s \sin(\omega_s t) + \sum_{n=1}^{\infty} a_n V_s \cos(n\ \omega_L t) \sin(\omega_s t)\ .$$

From trigonometry

$$22)\qquad \cos A \sin B = (1/2)\left[\ \sin(A+B) + \sin(A-B)\ \right]\ .$$

From (22), equation (21) may be written

$$23)\qquad i_d = (a_o/2)\ V_s \sin\omega_s t + V_s \sum_{n=1}^{\infty} (a_n/2)\Big[\sin(\omega_s + n\omega_L)t + \sin(\omega_s - n\omega_L)t\ \Big]\ .$$

Write (23) in the form

$$24)\qquad i_d = I_o + \sum_{n=1}^{\infty} (I_n^{+} + I_n^{-})$$

where $I_o = (a_o/2)V \sin\omega_s t$,

$$I_n^{+} = \frac{(a_n V_s/2)}{2} \sin(\omega_s + n\omega_L)t\ ,$$

and

$$I_n^- = \frac{a_n V_s}{2} \sin (\omega_s - n\omega_L)t.$$

The currents $I_n^{\pm}$ can be separated by filters, in which case the conversion admittance $Y_n^{\pm}$ corresponding to each $I_n^{\pm}$ can be written in the form

$$25) \qquad Y_n^{\pm} = \frac{|I_n^{\pm}|}{|V_s|} = \frac{a_n}{2} .$$

The value of the coefficients a_n of the Fourier series are given by

$$26) \qquad a_n = \frac{1}{\pi} \int_0^{2\pi} Y \cos n\omega_L t \quad d(\omega_L t) .$$

There exists, on the diode i-v characteristic, Figure 1, a point (i_o, v_o) that corresponds to optimum performance for the diode. In a mixer, the criterion for optimum performance is generally minimum conversion loss. This point is thus termed the optimum operating point for the diode. It should be noted that the location of this point depends upon the characteristics of the particular diode in question. Let $e_L(t)$ be such that the average value of $V_L \cos \omega_L t = v_o$. That is the diode is to be operated at its optimum point. It is not important here to evaluate the constants in (26); however, it should be noted that from (25), equation (19) may be written in the form

27) $$Y = Y_o + \sum_{n=1}^{\infty} Y_n \cos n\omega_L t$$

where the Y_n depend solely upon the diode characteristics. The diode is assumed to be operated at the optimum point.

Assume the load admittance is matched to the diode output admittance (at the optimum operating point) and that included in the load are lossless filters such that only the currents I_o and I_1^+, as defined in (24), are allowed to flow. Corresponding to I_o assume a generator

28) $$e_a(t) = E_a \sin \omega_s t \ ,$$

and corresponding to I_1^+ assume a generator

29) $$e_b(t) = E_b \sin (\omega_s + \omega_L)t \ .$$

The total diode current is thus $I_o + I_1^+$ and the applied signal is $e_a(t) + e_b(t)$. Hence, from (20) write

30) $$i_d = I_o + I_o = Y e_s(t) = Y \left[e_a(t) + e_b(t) \right] \ .$$

Inclusion in (25) of the currents in the source admittances Y_s and Y_1^+ of the generators $e_a(t)$ and $e_b(t)$ respectively, yields the total current i_T. Hence,

31) $$i_T = e_a(t)\ (Y + Y_s) + e_b(t)\ (Y + Y_1^+)\ .$$

Substitute (27), (28) and (29) into (31) obtaining,

32) $$i_T = E_a \sin \omega_s t\ (Y_s + Y_o + \sum_{n=1}^{\infty} Y_n \cos n\omega_L t)$$

$$+ E_b \sin (\omega_s + \omega_L)t\ (Y_1^+ + Y_o + \sum_{n=1}^{\infty} Y_n \cos n\omega_L t)\ .$$

Expand (32) obtaining,

33) $$i_T = (Y_s + Y_o)\ E_a \sin \omega_s t + Y_1 E_a \sin \omega_s t \cos \omega_L t$$

$$+ (Y_1\ \ + Y_o)E_b \sin (\omega_s + \omega_L)t + Y_1 E_b \sin (\omega_s + \omega_L)t \cos \omega_L t + \ldots\ .$$

Use (22) to expand (33), thus obtaining

34) $$i_T = (Y_s + Y_o)E_a \sin \omega_s t + Y_1 E_a \sin (\omega_s + \omega_L)t + Y_1 E_a \sin (\omega_s - \omega_L)t$$

$$+ (Y_1^+ + Y_o)E_b \sin (\omega_s + \omega_L)t + Y_1 E_b \sin \omega_s t$$

$$+ Y_1 E_b \sin (\omega_s + 2\omega_L)t + \ldots .\ .$$

The only allowed terms in (34) are those having frequencies ω_s or $(\omega_s + \omega_L)$. Thus, the terms in (34) of frequency $(\omega_s - \omega_L)$ and $(\omega_s + 2\omega_L)$ can be deleted, it should also be clear that expansion of Y for $n > 1$ yields no additional useful terms.

Equation (34) may be simplified to obtain

$$35)\qquad i_T = \left[(Y_s + Y_o)E_a + Y_1E_b \right] \sin \omega_s t$$

$$+ \left[Y_1E_a + (Y_1^+ + Y_o)E_b \right] \sin (\omega_s + \omega_L)t \quad .$$

It was necessary to assume a generator $e_b(t)$ in order to establish the desired currents. At this point remove the generator $e_b(t)$ from the circuit leaving its source admittance Y_1^+. It is clear that (35) represents a nodal equation for a two node circuit. The nodal equations for a general two node circuit, with a generator at the first node are

$$36)\qquad i_1 = Y_{11}E_1 + Y_{12}(E_1 - E_2)$$

and

$$37)\qquad 0 = Y_{12}(E_2 - E_1) + Y_{22}E_2$$

where Y_{ii} = the total admittance between node i and the reference node,

$Y_{ij} = Y_{ji}$ = the admittance between nodes i and j.

To put (35) into the form of (36) and (37), equate currents and sources of like frequency in (35), recalling that $e_b(t)$ was removed. Hence,

38) $$i_s = (Y_s + Y_o)E_a + Y_1E_b$$

and

39) $$0 = Y_1E_a + (Y_1^+ + Y_o)E_b \quad .$$

Collect like terms in (36) and (37) to obtain

40) $$i_1 = (Y_{11} + Y_{12})E_1 - Y_{12}E_2$$

and

41) $$0 = - Y_{12}E_1 + (Y_{22} + Y_{12})E_2$$

Let $i_s = i_1$, $E_a = E_1$, $E_b = E_2$ and equate (38) to (40) and (39) to (41). Hence,

42) $$(Y_{11} + Y_{12} + Y_s + Y_o)E_1 - (Y_1 + Y_{12})E_2 = 0$$

and

43) $$- (Y_{12} + Y_1)E_1 + (Y_{22} + Y_{12} + Y_1^+ + Y_o)E_2 = 0 \quad .$$

Equations (42) and (43) are two equations in three unknowns, which as a rule are not solvable. This difficulty may be eliminated either by solving for two of the unknowns in terms of the third or by assuming a solution for one unknown and then solving for the other two. Hence, let

44) $$Y_{12} = Y_1 \ ,$$

then

45) $$Y_{11} = Y_s + (Y_o - Y_1)$$

and

46) $$Y_{22} = Y_L + (Y_o - Y_1) \ .$$

From (40), (41), (44), (45) and (46) the equivalent circuit presented in Figure 2-b may be drawn. Transforming to impedance parameters, one obtains the equivalent circuit of Figure 3. The circuit of Figure 3 may be replaced by a Thévenin's equivalent circuit as shown in Figure 4 where

47) $$Z_T = Z_o - \frac{{Z_1}^2}{(Z_o + Z_s)} \ ,$$

and

48) $$E_T = \frac{E_s Z_1}{(Z_s + Z_o)} \ .$$

Initially it was assumed that the diode impedances were very much greater than the impedances of other circuit elements. In general the source impedance is equal to, or less than the load impedance, and the d-c diode impedance Z_o is equal to or less than Z_1. This may be written mathematically as

$$49) \qquad Z_s \leq Z_1^{+} \ll Z_o \leq Z_1 \quad .$$

Application of the inequalities (49) into (47) and (48) yields

$$50) \qquad Z_T \cong Z_o(v) = R(v) + jX(v) \quad ,$$

and

$$51) \qquad E_T \cong E_1 \frac{Z_1}{Z_o} \quad .$$

Recall that the diode impedance is a function of the local oscillator voltage. Suppose the load impedance is matched to the diode impedance at the optimum operating point. That is, as shown in Figure 5,

$$52) \qquad Z_1^{+} = Z_o(v_o)$$

Then it should be clear that the match exists at only that point, and for other operating points, corresponding to other local oscillator voltages, the source and load reactive components do not cancel, and a phase shift is thus introduced that is proportional to the function relating diode impedance and local oscillator drive level.

Empirical Investigation:

Two empirical investigations were performed with different objectives; (a) To provide data for use in computing the phase shift via the equivalent circuit derived in the previous section, and (b) to provide a measurement of the phase shift for comparison with the computed phase shift.

Crystal Mixer Output Impedance:

The output impedance of a General Radio 874 crystal mixer, using a Sylvania 1N21-C crystal diode, was measured. The measurements were performed with a General Radio 1607-A Transfer-function and Immittance Bridge. The diode was self-biased, and the output impedance, at 60 mc/sec, was measured as a function of local oscillator drive level, as indicated by d-c diode current. These data are presented in Table I.

Experimental Measurement of Mixer Phase Shift:

A block diagram of the equipment used to measure the phase shift through a mixer is presented in Figure 7. The signal and local oscillator frequencies were chosen to be 45 mc/sec and 15 mc/sec respectively, and the frequency sum component at 60 mc/sec was chosen as the mixer output signal. The selection of these particular frequencies was based on equipment availability.

In the analysis it was shown that the phase delay between signals of different frequencies is defined only if the signals are phase coherent. Hence, phase coherent signals at 45 mc/sec and 15 mc/sec were required for the phase delay measurement. The 45 mc/sec signal was obtained from a signal generator, and a phase coherent 15 mc/sec signal was synthesized from the 45 mc/sec signal by a regenerative frequency divider circuit (Figure 8) constructed for that purpose.

The mixer input, local oscillator, and output signals were displayed on two Hewlett Packard 185B sampling oscilloscopes. On one, the input and output signals were compared, while on the other, the local oscillator and input signals were compared. The latter comparison assures a constant relative phase between signal and local oscillator inputs, while the former is a visual presentation of the mixer phase delay. The load was matched to the mixer at an operating point defined by the local oscillator drive level corresponding to 1.6 ma, d-c diode current. The "input-output" oscilloscope presentation was then photographed, as the local oscillator drive level was varied over a range corresponding to d-c diode currents of 0.6 ma to 5.2 ma. These data were taken for a small input signal (-35 dbm) and also a large input signal (-15 dbm). From these photographs, the mixer phase shifts as a function of diode current were extracted and plotted (Figure 10).

The theoretical phase shift versus diode current curve is also plotted in Figure 10 where it may be seen that good agreement exists in the forms of the curves; however, the absolute values agree only approximately. It is, however, the form of the phase shift-versus-diode current curve that is of real interest, since at any diode current, the phase shift may be nulled by a proper terminating network. Thus the problem is to determine the diode current at which the phase shift is least sensitive to changes in local oscillator drive level as represented by diode current. Figure 10 clearly shows that this consideration implies a large diode current. It is of interest to observe from the data of Table II that the condition for minimum conversion loss also corresponds to large diode current.

Conclusion:

This investigation reveals that there does exist, as herein defined, a phase shift through a crystal mixer. In general, a mixer diode is operated at some particulat point on its i-v characteristic and the load impedance is matched to the mixer output impedance corresponding to that operating point. Under these conditions the phase shift through the mixer and load may be fully compensated, and thus be zero at the design operating point. The question then arises, what is the rate of change of phase shift with respect to diode current or local oscillator drive level (either may be used to describe the operating point)? This investigation reveals that the rate of change of phase shift with respect to diode current is at a minimum for large diode current. It is fortunate that this is also the condition for minimum mixer conversion loss as shown in Figure 12.

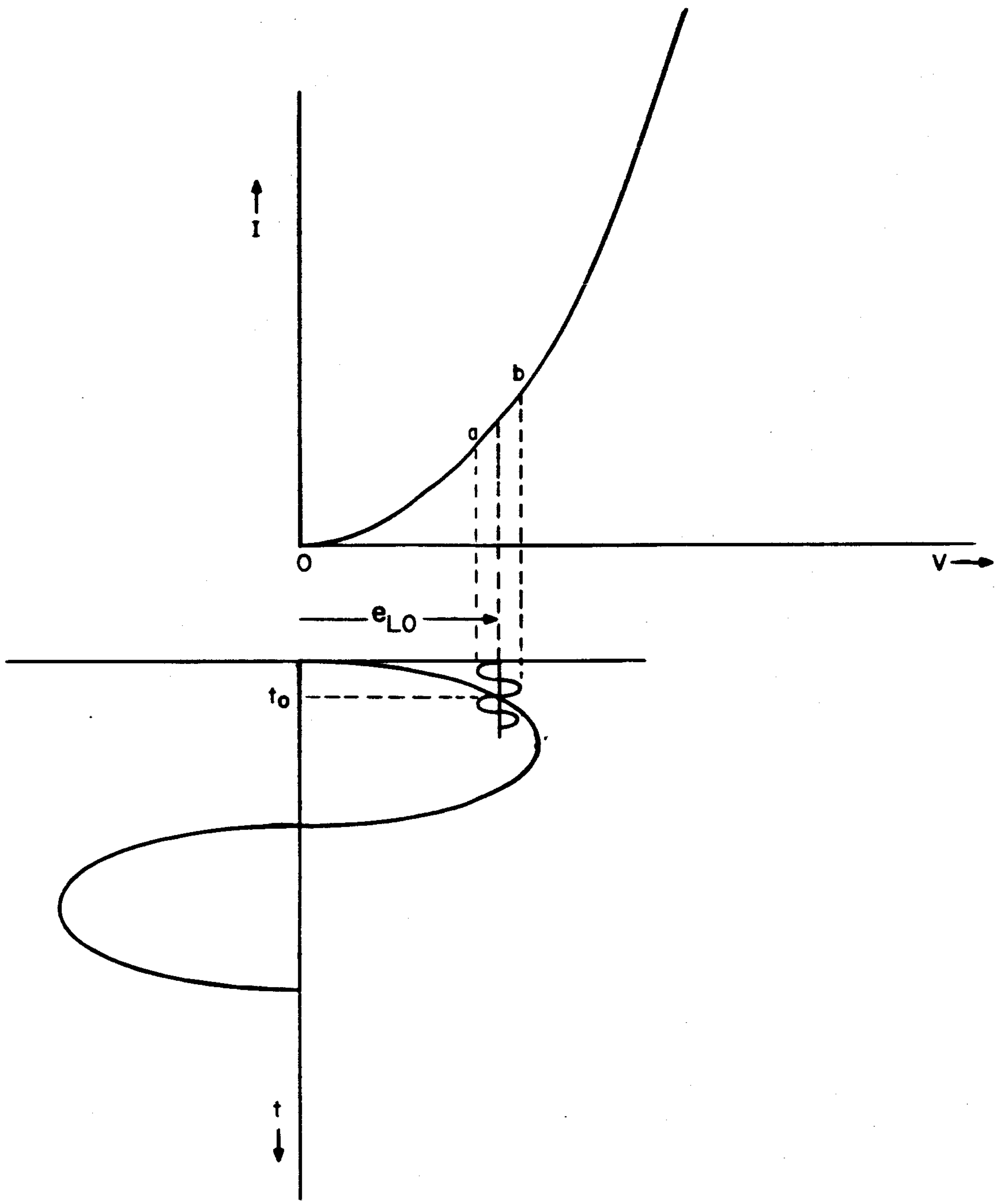

FIGURE 1 CHARACTERISTIC OF NONLINEAR RESISTANCE

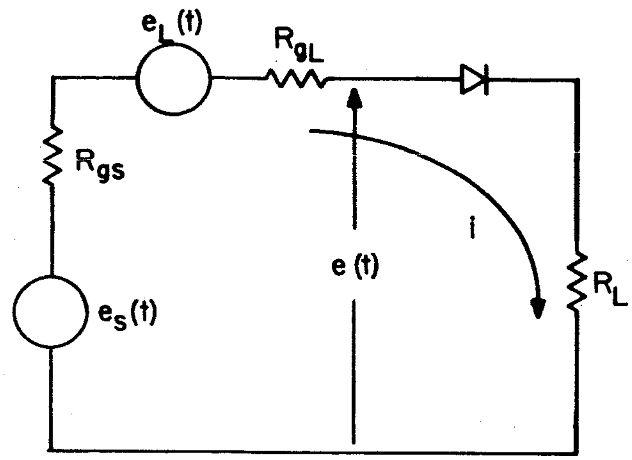

FIGURE 2a

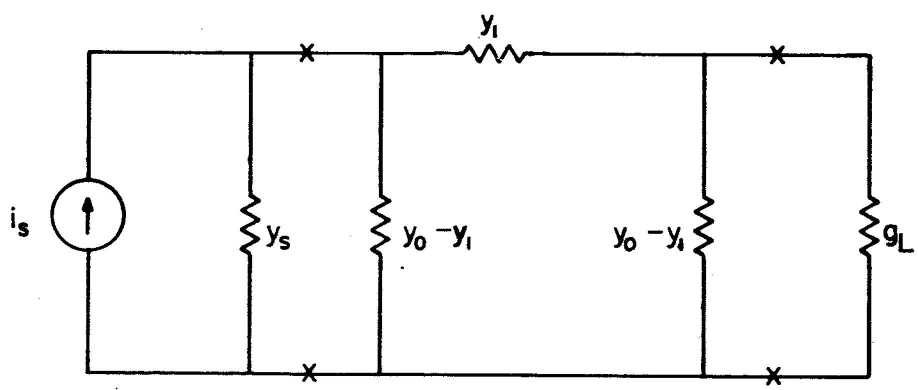

FIGURE 2b

FIGURE 2 CRYSTAL MIXER EQUIVALENT CIRCUITS

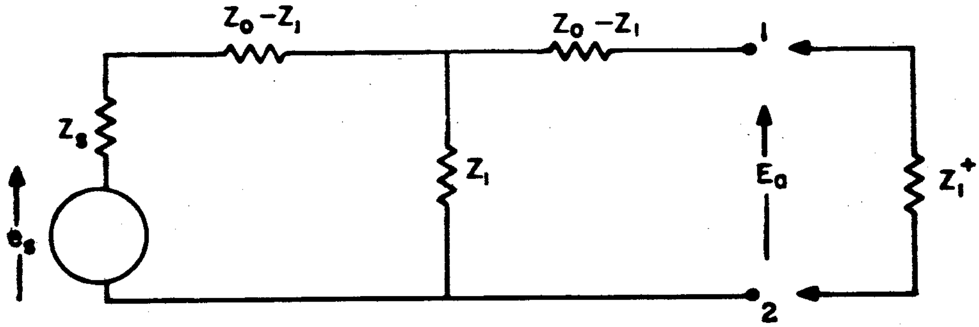

FIGURE 3 T—EQUIVALENT CIRCUIT OF A CRYSTAL MIXER

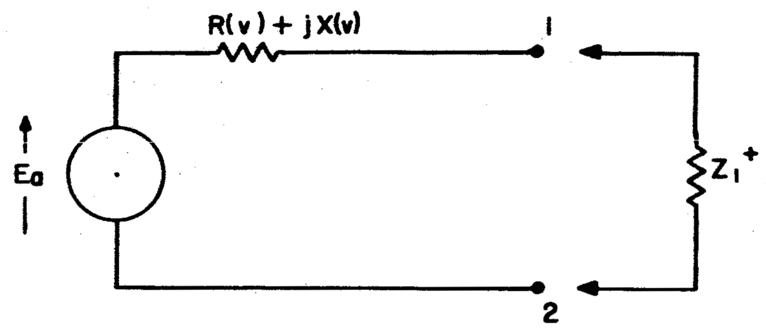

FIGURE 4 THEVENIN'S EQUIVALENT CIRCUIT OF A CRYSTAL MIXER

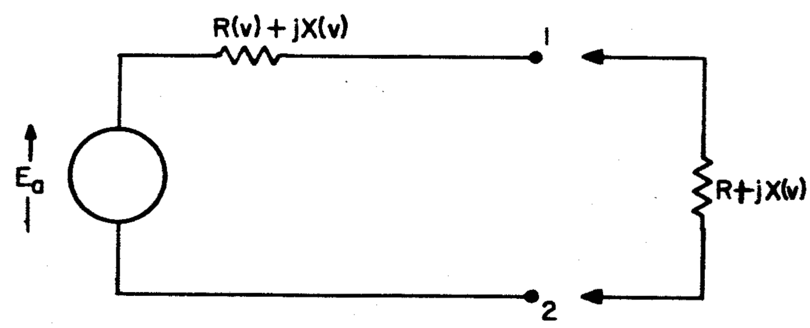

FIGURE 5 EQUIVALENT CIRCUIT AT THE OPTIMUM OPERATING POINT

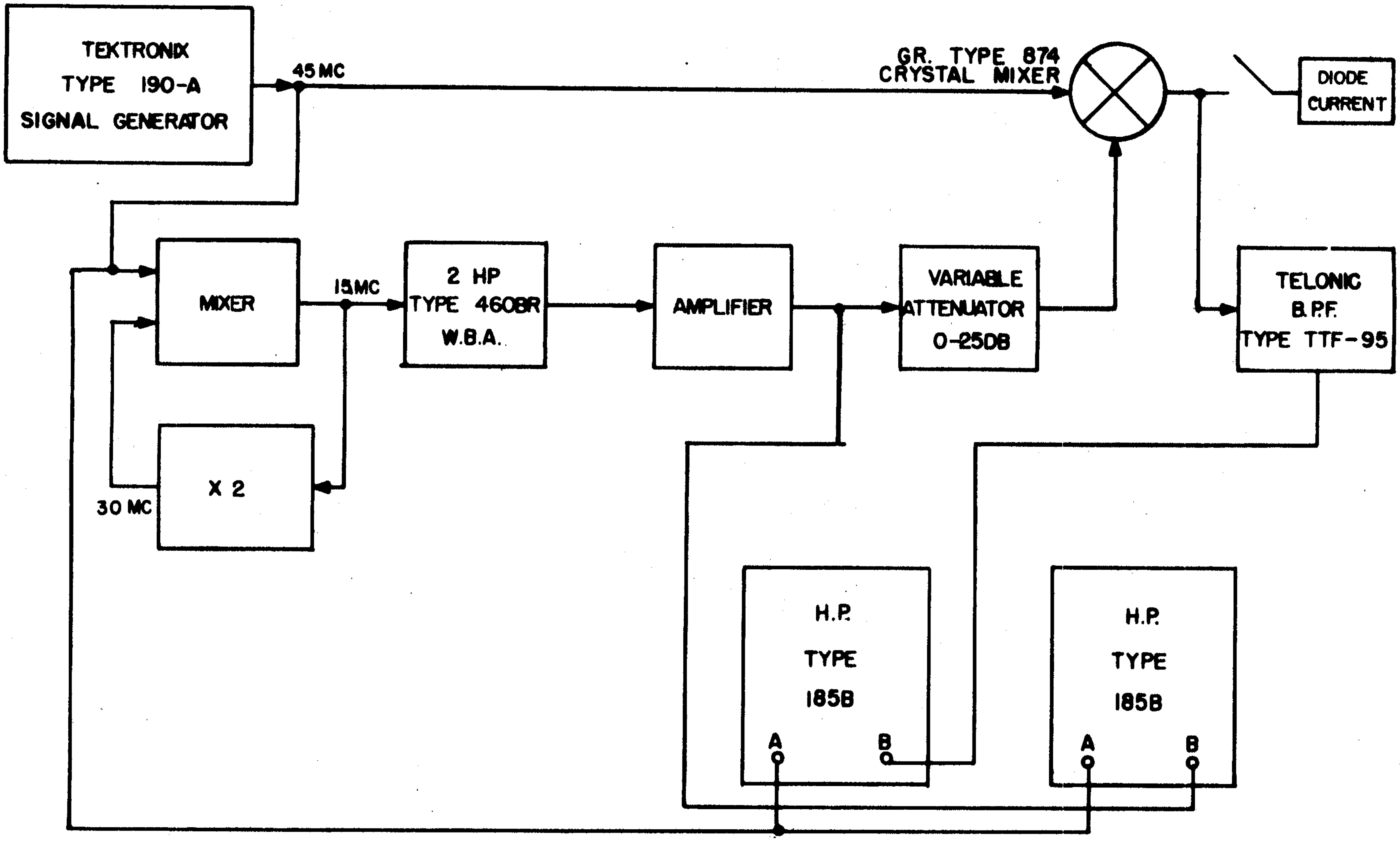

FIGURE 6 BLOCK DIAGRAM OF EQUIPMENT SET-UP USED TO OBTAIN PHASE SHIFT DATA

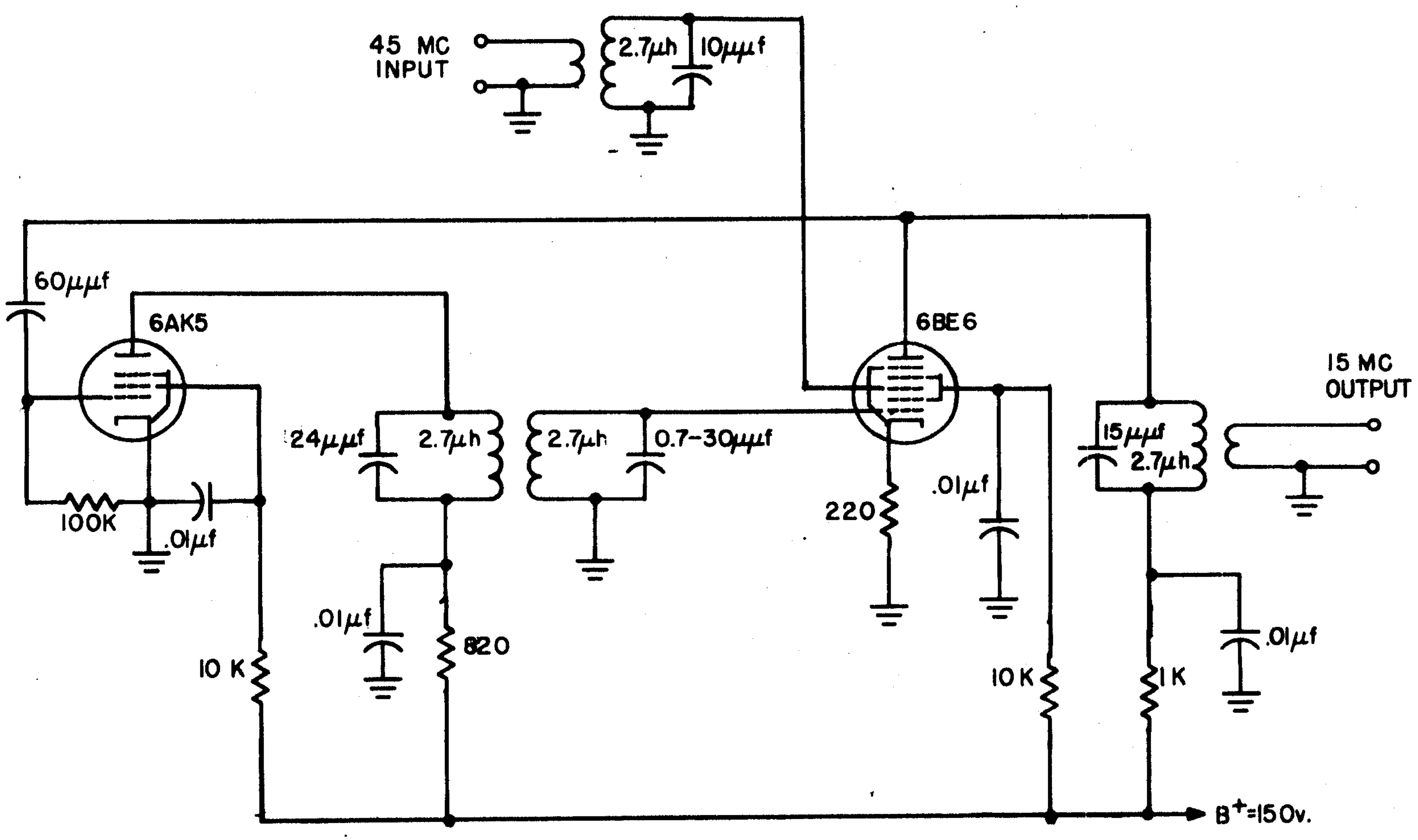

FIGURE 7 REGENERATIVE FREQUENCY DIVIDER

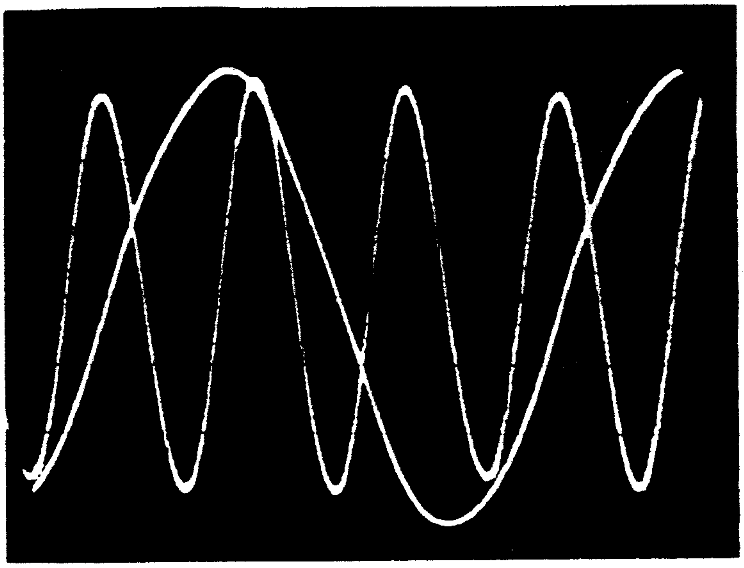

FIGURE 8a PHASE COHERENT L.O. AND SIGNAL VOLTAGE OF 15 MC/S AND 45 MC/S RESPECTIVELY FOR LARGE SIGNAL LEVELS

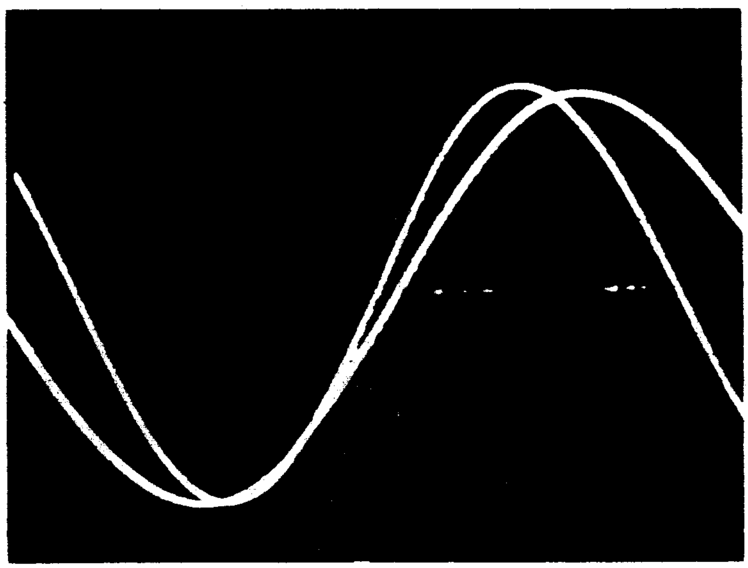

FIGURE 8b DIODE CURRENT 5.2 ma

FIGURE 8. TIME SHIFT IN ZERO CROSSINGS BETWEEN PHASE COHERENT 45 MC/S INPUT FREQUENCY AND 60 MC/S i-f FOR LARGE SIGNAL LEVELS

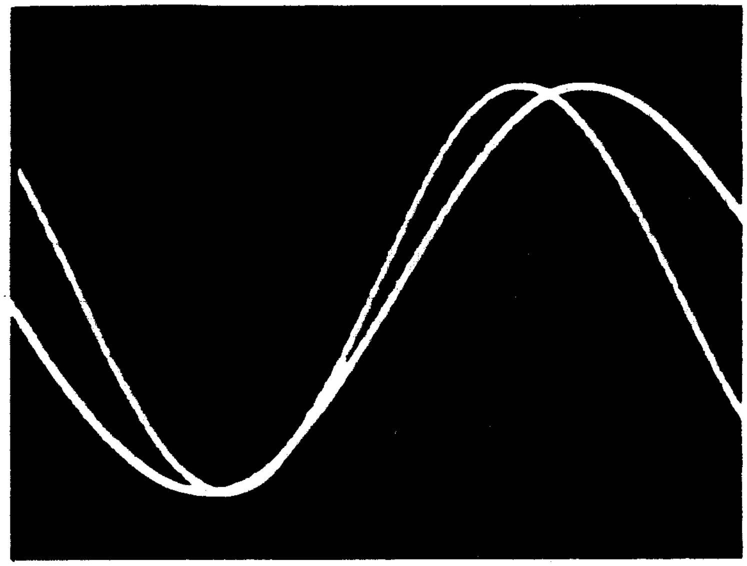

FIGURE 8c DIODE CURRENT 2.7ma

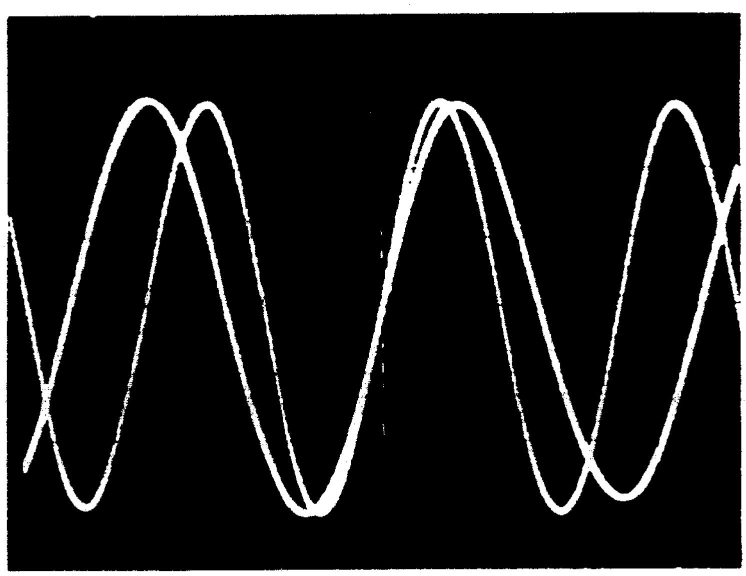

FIGURE 8d DIODE CURRENT 1.4 ma

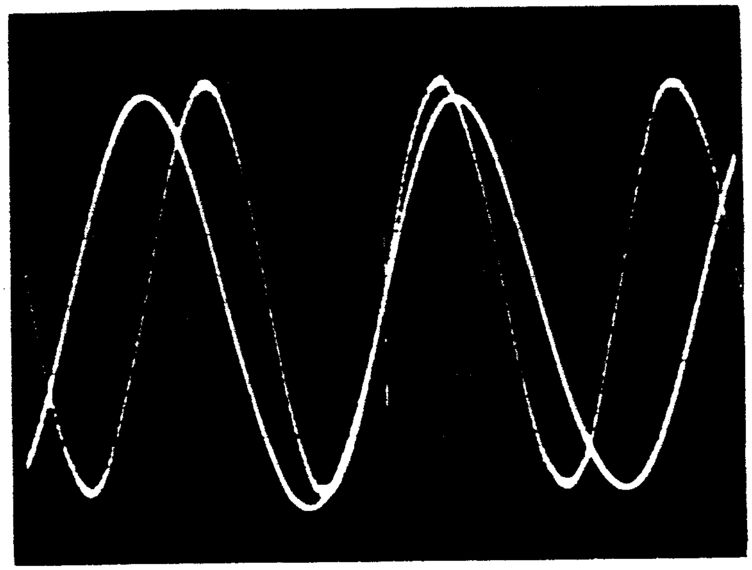

FIGURE 8e DIODE CURRENT 0.6 ma

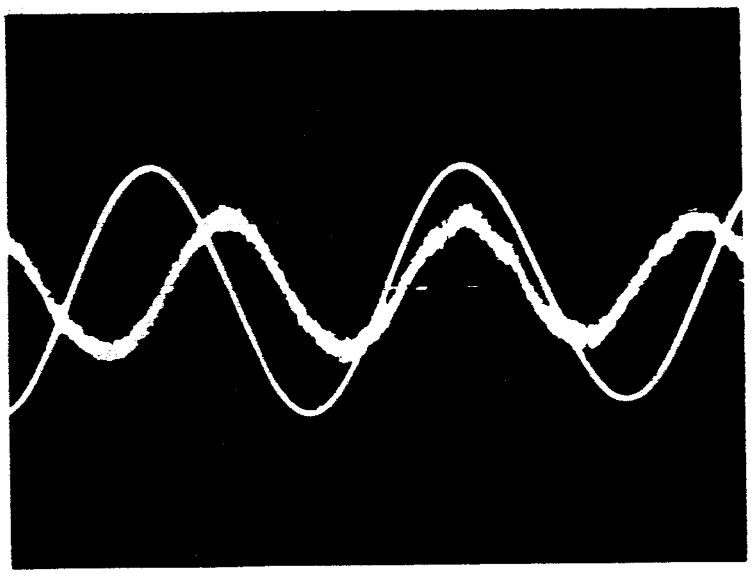

FIGURE 8f DIODE CURRENT 0.1 ma

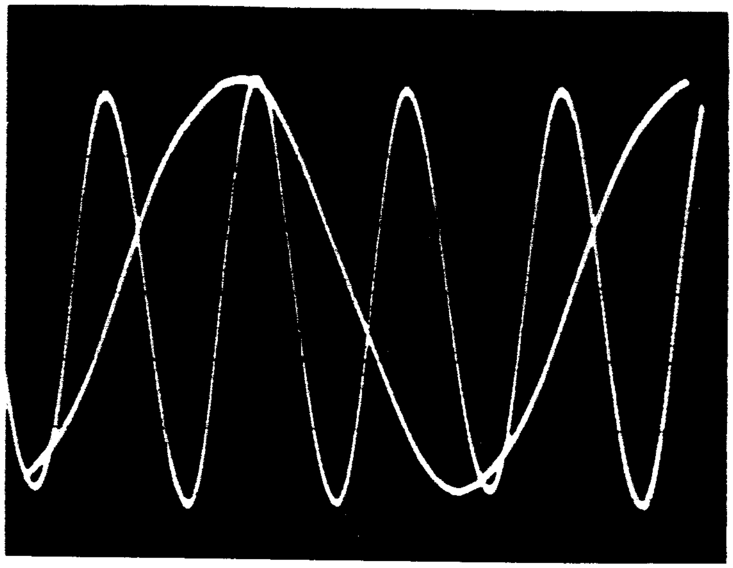

FIGURE 9a PHASE COHERENT L.O. AND SIGNAL VOLTAGES OF 15mc/s AND 45mc/s RESPECTIVELY FOR SMALL SIGNAL LEVELS

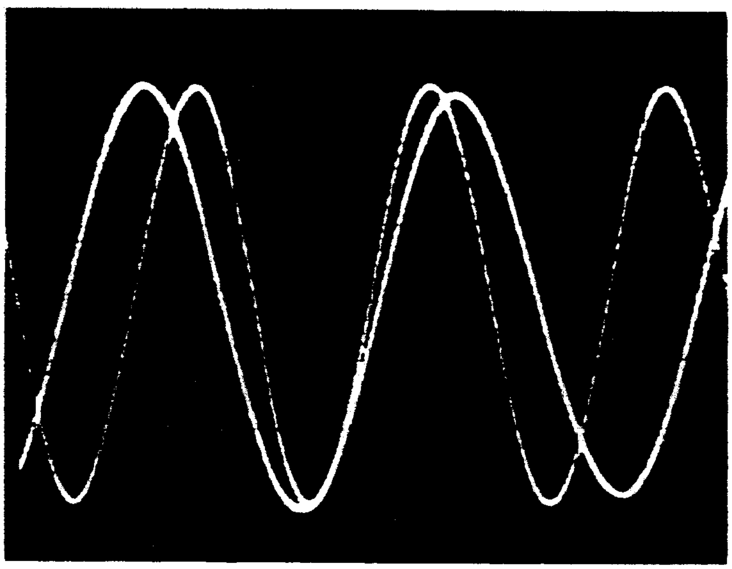

FIGURE 9b DIODE CURRENT 5.2ma

FIGURE 9 TIME SHIFT IN ZERO CROSSINGS BETWEEN 45mc/s INPUT FREQUENCY AND 60mc/s i-f FREQUENCY FOR SMALL SIGNAL LEVELS

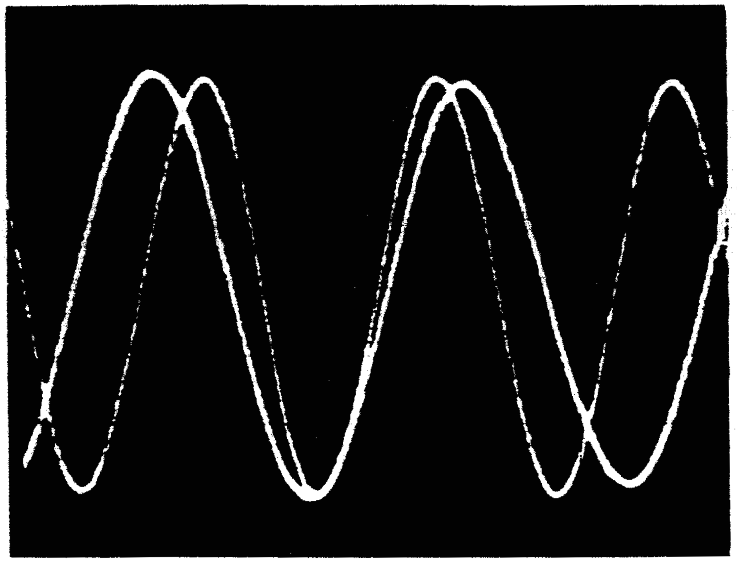

FIGURE 9 c DIODE CURRENT 2.6ma

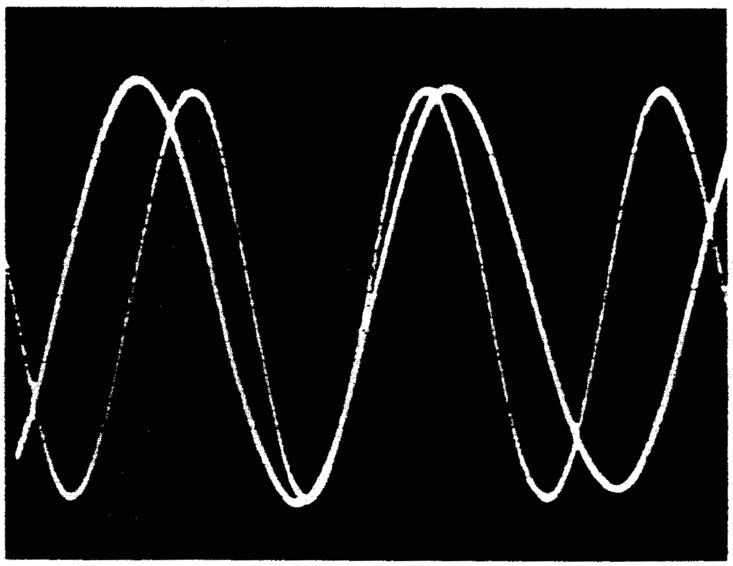

FIGURE 9 d DIODE CURRENT 1.4ma

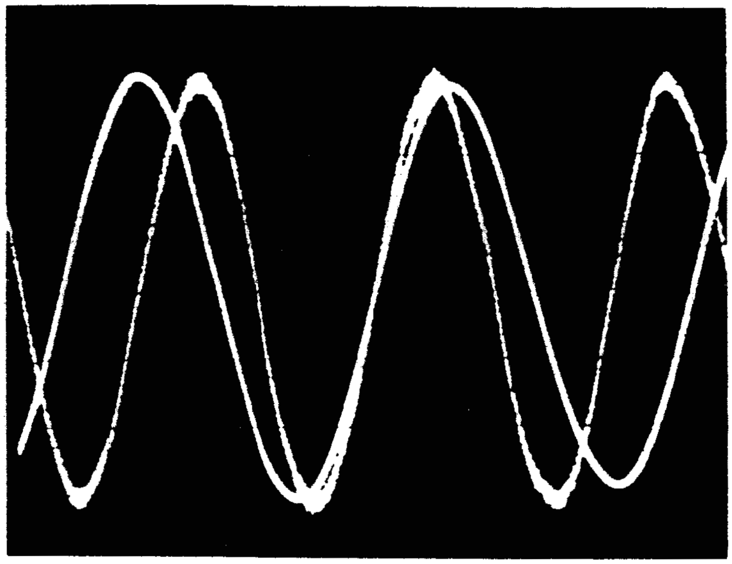

FIGURE 9e DIODE CURRENT 0.6ma

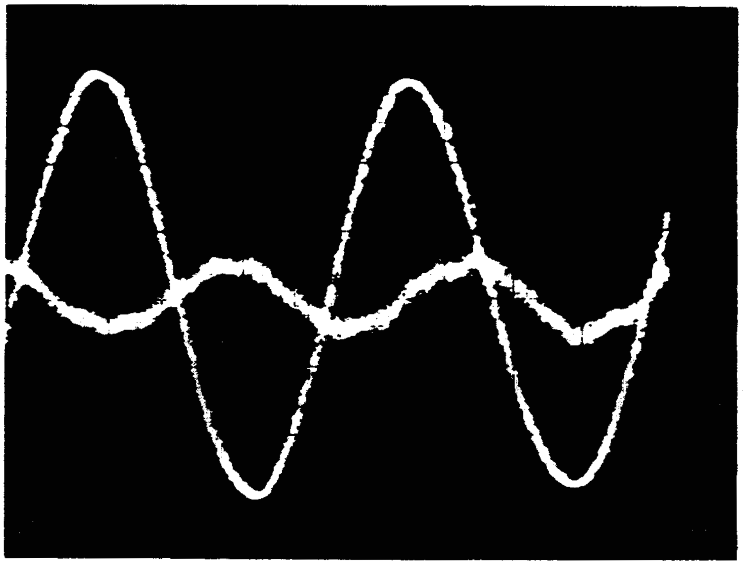

FIGURE 9 f DIODE CURRENT 0.1ma

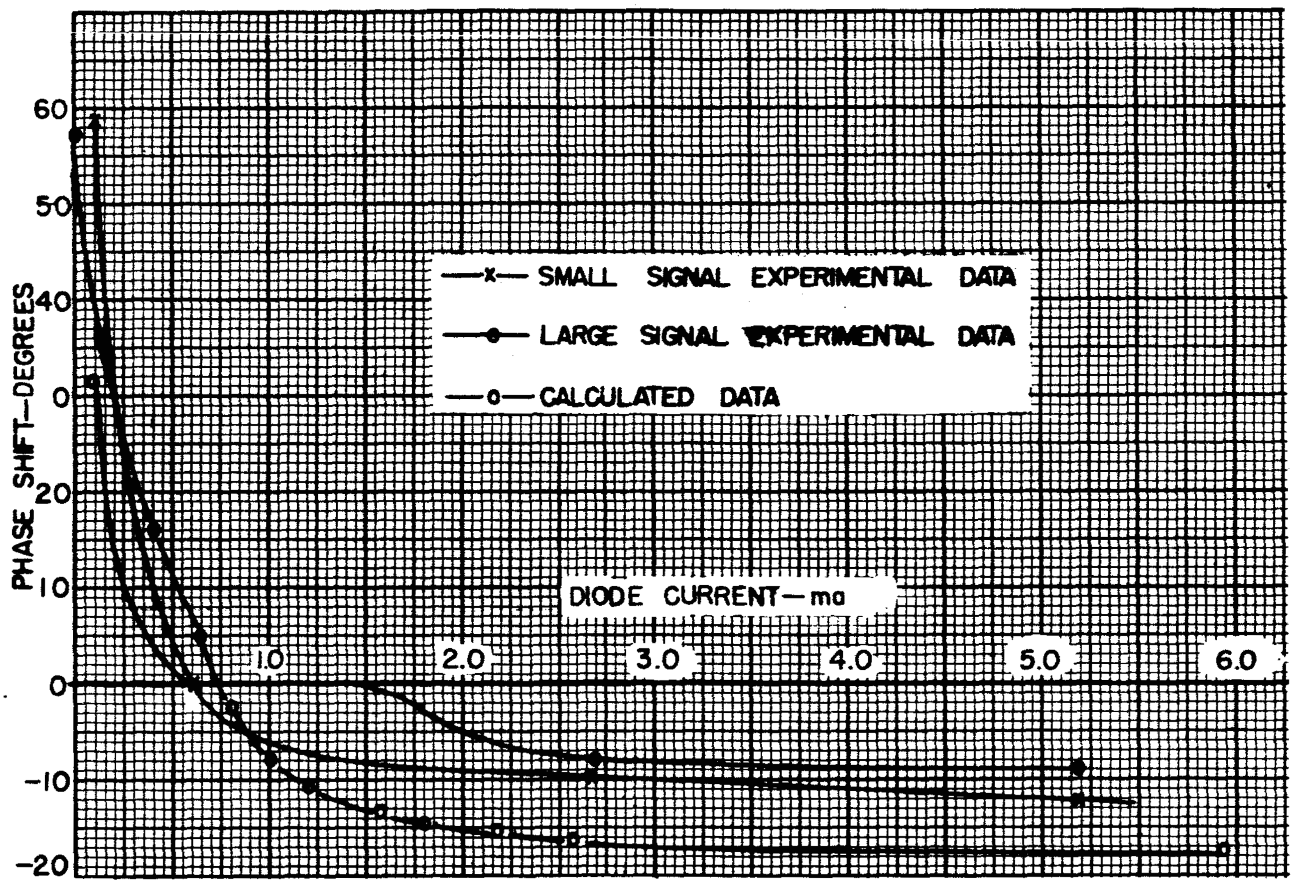

FIGURE 10 PHASE SHIFT THROUGH A MIXER AS A FUNCTION OF L.O. VOLTAGE (DIODE CURRENT)

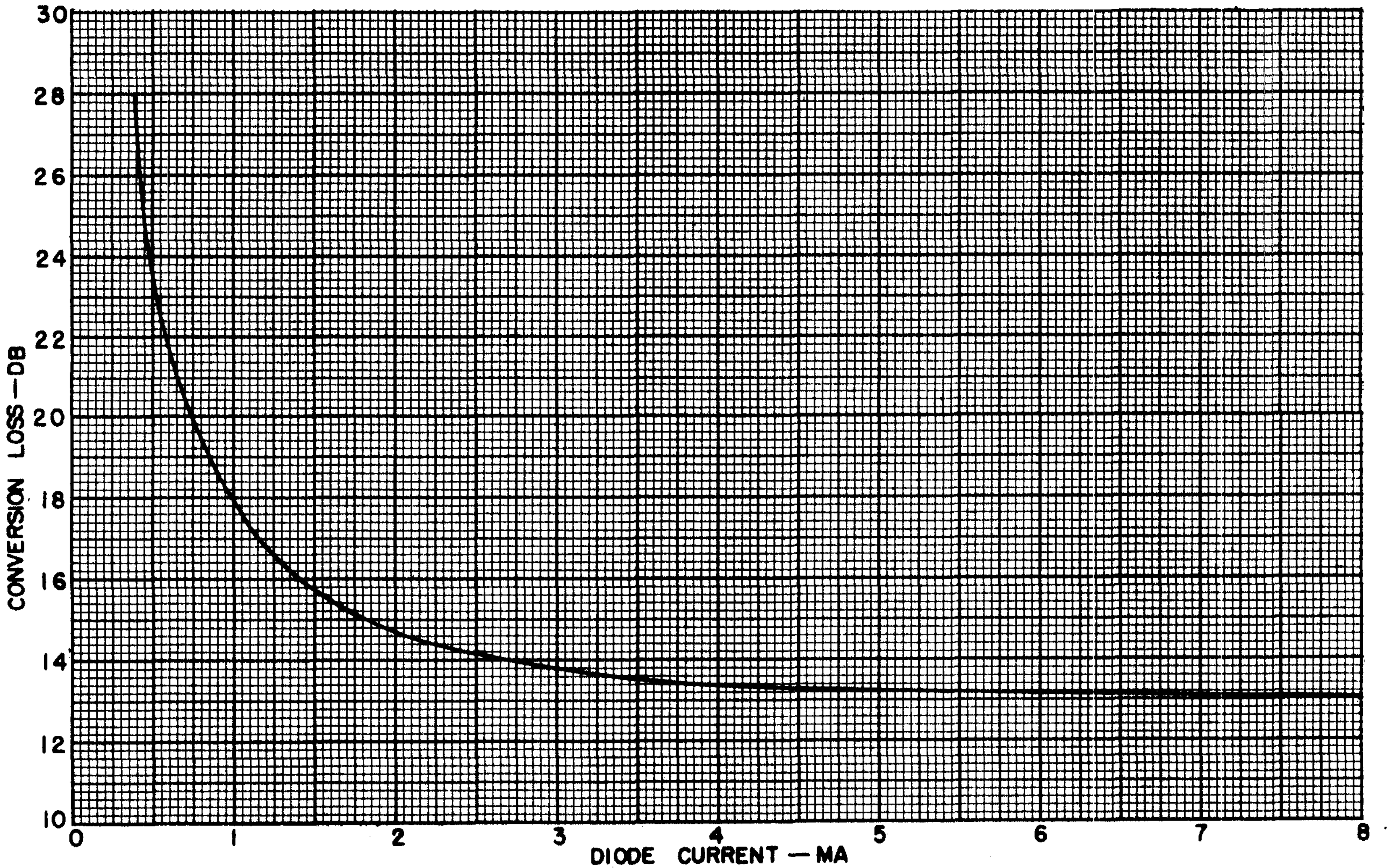

FIGURE 11 CONVERSION LOSS OF IN23 CRYSTAL DIODE
USED IN GR. TYPE 874 MIXER

TABLE I EXPERIMENTAL DATA FOR OUTPUT IMPEDANCE AS A FUNCTION OF DIODE CURRENT

DIODE CURRENT	OUTPUT IMPEDANCE
(milliamperes)	(ohms)
0	17.5 - j337.5
.4 MA	185 - j180.0
.6 MA	187 - j150
.65MA	190 - j120
.80MA	150 - j70
1.0	120 - j42.5
1.2	97.5 - j30.0
1.6	87.5 - j22.0
1.8	82.5 - j18.75
2.2	77.5 - j16.25
2.6	75.0 - j12.50
2.95	70.0 - j12.5
3.2	68.0 - j12.5
4.1	67.5 - j12.0
5.8	66.3 - j11.3

TABLE II EXPERIMENTAL DATA FOR CONVERSION LOSS AS A FUNCTION OF DIODE CURRENT

DIODE CURRENT (milliamperes)	CONVERSION LOSS (decibels)
4.2 MA	13.2
3.8 MA	13.64
3.0 MA	13.82
2.5 MA	14.23
2.1 MA	14.61
1.8 MA	15.10
1.3 MA	16.42
1.65MA	15.42
1.0 MA	17.88
1.15MA	17.25
0.9 MA	18.82
0.65MA	20.90
0.5 MA	22.72
0.45MA	24.15
0.4 MA	27.45