AN INTEGRATED LINEAR-TRANSCONDUCTANCE ANALOG MULTIPLIER

by

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A Thesis Submitted to the Faculty of the

DEPARTMENT OF ELECTRICAL ENGINEERING

In Partial Fulfillment of the Requirements
For the Degree of

MASTER OF SCIENCE

In the Graduate College

THE UNIVERSITY OF ARIZONA

1970

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ACKNOWLEDGEMENTS

The project described in this report was funded by and accomplished at Motorola Semiconductor Products Division, Phoenix, Arizona. This effort was accomplished in cooperation with a continuing hybrid digital-analog computer study directed by Professor G. A. Korn.

The Hybrid-Digital-Analog Computer Laboratory is supported by the Electrical Engineering Department and Engineering Experiment Station funds, and partially supported by a National Science Foundation Institutional grant to The University of Arizona and a National Aeronautics and Space Administration grant No. G-646.

The writer is, in particular, grateful to Motorola Semiconductor Products Division for the contribution of company facilities and to Dr. R. Mattson for his contribution of University facilities.

The author wishes to thank Mr. J. Thompson and Mr. J. Solomon, both of Motorola's Semiconductor Products Division, whose superb insight and guidance made this project possible. The author also wishes to thank Mr. R. Nelson who helped to characterize this project and Mrs. S. Mayberry who typed this report.

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ABSTRACT

A new solid-state, monolithic integrated analog multiplier based on the controlled-transconductance principle is designed for high speed, high accuracy and excellent linearity. A novel differential amplifier circuit permits cancellation of undesired nonlinearities and carefully matched device geometries account for the good accuracy of the new multiplier. With an inherent voltage gain of $A_V = 1/10$, the multiplier will provide a ± 10 v. output at ± 10 ma. with linearity approaching 0.1% at 100 hz. The multiplier package may be used in conjunction with an external operational amplifier and external resistors, and offers the advantages of monolithic construction (low cost, excellent temperature tracking) with the precision of adjustable-parameter, discrete construction.

INTRODUCTION

The precision linear-analog multiplier is an extremely useful and powerful building block in the construction of complex analog computers and many types of automatic control mechanisms. It allows the user to perform multiplication, division, square, square root, mean-square and trigonometric computations directly and at high speeds with good accuracy. The analog multiplier can also be used as a balanced modulator, a voltage-controlled function generator and in automatic-gain-control applications.

Linearity on the order of 0.1% error of full scale output at 100 hz is possible with the analog approach. Of the two most important commercial techniques available today, that is, diode quarter-square and non-linear transconductance, the former is more accurate, but operates at lower speeds. The non-linear transconductance (N.L.T.) multiplier has the advantages of integrated, solid-state construction, low cost and can approach 1.0% linearity at 100 hz. The N.L.T. Multiplier can operate at frequencies above 1 Mhz with a linearity of about 5% of full scale. Generally, the N.L.T. Multiplier suffers from poor temperature tracking.

An Integrated Linear Transconductance Multiplier (I.L.T.) is described herein. This I.L.T. multiplier has excellent tracking characteristics, linearity approaching 0.1% at 100 hz, and a bandwidth exceeding 1 Mhz. In addition, it is a true four quadrant linear multiplier. A first-order analysis and a second-order circuit analysis with emitter-base-junction offsets is developed in the text.

Present State-of-the-Art

Table 1 lists several commercial multipliers available today and their specifications. Table 2 lists the specifications of the new Linear-Transconductance, Four Quadrant Multiplier.

Table 1

Some Commercially Available Multipliers and Their Specifications

Manufacturer	Model	Unit Cost	-3db Response	Full-Output Freq.	Max. Output Voltage	Max. Output Current	Transfer Function	Input Impedance	Total Output Error	Error Temp. Drift	Working Temp. Range
Analog Devices	421A	$ 95	600 kHz	50 kHz	10V	5 mA	XY/10	10k ±10%	± 1%	±0.2%/°C	-25 to +85°C
Burr-Brown Research	4031/25	$145	-	3 Hz	10V	5 mA	XY/10	X=25kΩ Y=100kΩ	±0.5%	±0.11%/°C	-25 to +60°C
Intronics	M401	$95	125 kHz	20 kHz	10V	5 mA	XY/10	10kΩ	± 2%	±0.03%/°C	-25 to +85°C
Optical Electronics	5485	$91	100 kHz	10 kHz	10V	10 mA	XY/10	100kΩ	± 3%	±0.3%/°C	-55 to +70°C

Table 2

Specifications For The Linear-Transconductance,

Four Quadrant Multiplier

Output Impedance (without op. amp)	7.5 K ohms
-3db Response (without op. amp)	4 Mhz
Full-Output Frequency (without op. amp)	4 Mhz
1% Error Frequency	2 KC
Maximum Output Voltage Excursion	$\pm$ 10V
Maximum Output Current (with op. amp)	10 ma
Transfer Function	XY/10
Input Impedance	1 Meg. ohm
Total Output Error at 100 Hz	0.2%
Scale Factor Temperature Drift Percent of Full Scale	0.05%/^{o}C
Output Offset Temperature Drift	1 mV/^{o}C
Working Temperature Range	-55^{o}C to +85^{o}C
Power Supplies	$\pm$ 15V DC

OVERALL DESIGN

The Linear Four Quadrant Multiplier

The following is an approximate analysis of a new type of analog multiplier circuit which can linearly multiply two voltages to produce a product with linearity approaching 0.1%. The operation of the circuit is unusual in the sense that it responds linearly to either of two inputs x or y, yet it provides a product xy as a result of nonlinear processing of both x and y. It achieves this operation by preconditioning one of the inputs with a nonlinearity[1] which is the complement of the nonlinearity used for multiplication. In the analysis which follows, we will derive equations which show that the signal distortion arises primarily from the effects of voltage offset between the various devices used in the multiplier. A second source of nonlinearity arises from the presence of ohmic drops in the transistor base regions, but it can be shown that at low frequencies, this effect is negligible when compared with that due to voltage offsets.

The circuit model configuration for the multiplier is shown in Figure 1, where it has been assumed that all of the transistors are identical and have small voltage offset generators in series with each of their emitters. Other than these voltage offset generators, the devices are assumed to be perfectly matched and behave according to the ideal Ebers-Moll model. We assume that the lower differential amplifiers have been sufficiently linearized by the emitter resistors

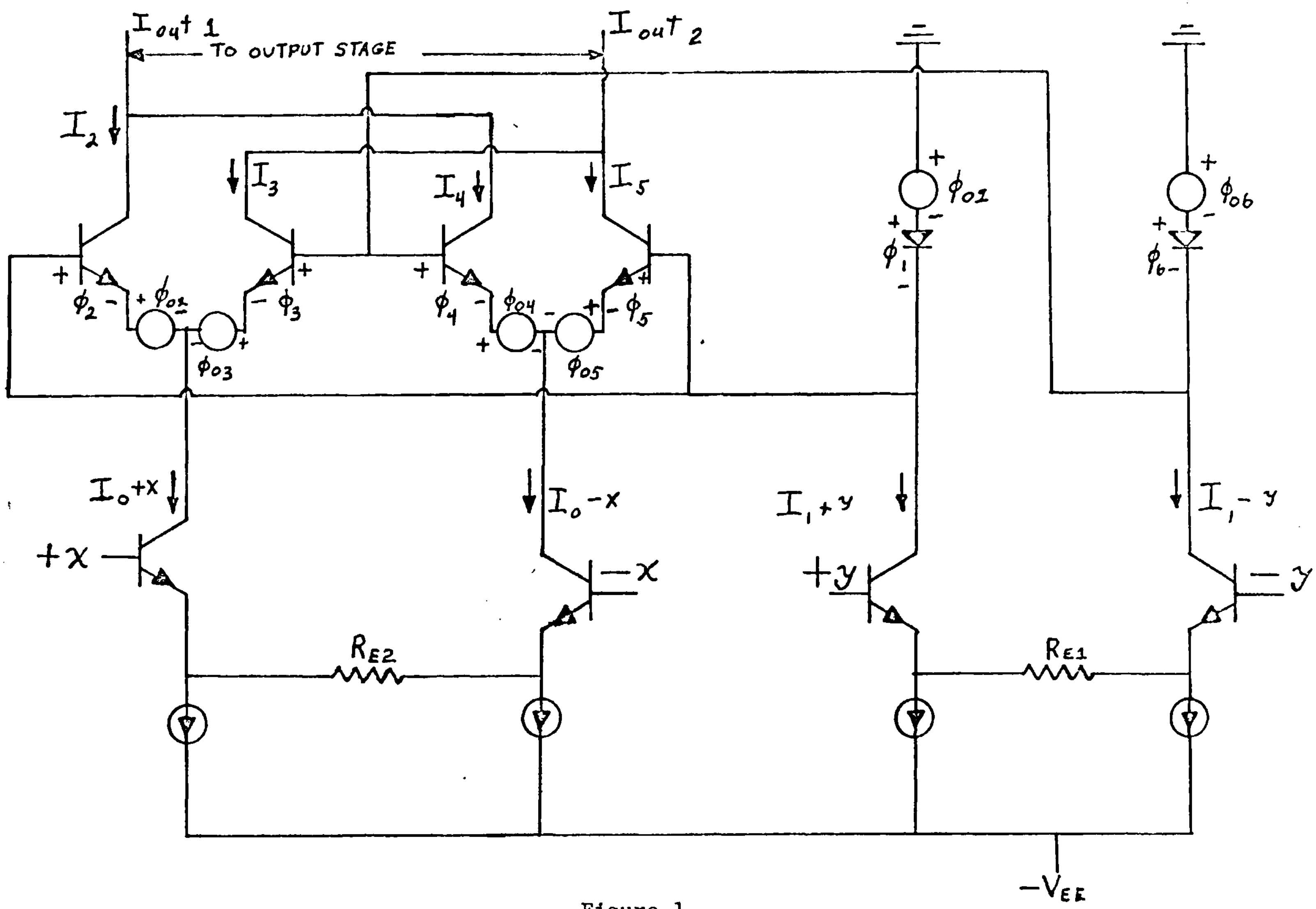

Figure 1

Multiplier Circuit Model Diagram

R_{e1} and R_{e2} that the currents $I_o \pm x$ and $I_1 \pm y$ are linearly dependent on the input voltages V_1 and V_2. Thus,

$$x = \frac{V_2}{R_{e2} + 2r_{e2}}$$

$$y = \frac{V_1}{R_{e1} + 2r_{e1}}$$

where V_2 and V_1 are the differential voltages at the x and y inputs respectively.

We will consider, therefore, only the current transfer from the collectors of the multiplier outputs. That is, we want to calculate the ratio I_{out}/xy. We begin by writing the Ebers-Moll expressions[2] relating the collector currents and the base emitter potentials as shown below. (It is assumed that leakage currents can be ignored and that α's are absorbed into the constant a_{11}.)

$$\begin{aligned}
I_1 + y &= a_{11}e^{q\phi_1/kT} \\
I_1 - y &= a_{11}e^{q\phi_6/kT} \\
I_2 &= a_{11}e^{q\phi_2/kT} \\
I_3 &= a_{11}e^{q\phi_3/kT} \qquad (1) \\
I_4 &= a_{11}e^{q\phi_4/kT} \\
I_5 &= a_{11}e^{q\phi_5/kT}
\end{aligned}$$

These expressions can be simplified by dividing some currents by others as follows:

$$\frac{I_1 + y}{I_1 - y} = e^{\frac{q}{kT}(\phi_1 - \phi_6)}$$

$$\frac{I_2}{I_3} = e^{\frac{q}{kT}(\phi_2 - \phi_3)} \qquad (2)$$

$$\frac{I_4}{I_5} = e^{\frac{q}{kT}(\phi_4 - \phi_5)}$$

We also note that:

$$I_2 + I_3 = I_o + x$$
$$I_4 + I_5 = I_o - x \qquad (3)$$

$$\phi_1 + \phi_{01} + \phi_2 + \phi_{02} + \phi_{03} - \phi_3 - \phi_6 - \phi_{06} = 0$$
$$\phi_1 + \phi_{01} + \phi_5 + \phi_{05} - \phi_{04} - \phi_4 - \phi_6 - \phi_{06} = 0 \qquad (4)$$

Simplifying Equation (3) and combining with (2) we get an expression for I_2:

$$1 + \frac{I_3}{I_2} = \frac{I_o + x}{I_2} \qquad (5)$$

$$I_2 = \frac{I_o + x}{1 + \frac{I_3}{I_2}} = \frac{I_o + x}{1 + e^{\frac{q}{kT}(\phi_3 - \phi_2)}} \qquad (6)$$

Manipulating Equations (4) and (2), Equation (6) can be simplified:

$$\phi_3 - \phi_2 = (\phi_1 - \phi_6) + (\phi_{01} - \phi_{06}) + (\phi_{02} - \phi_{03}) \quad (7)$$

$$e^{\frac{q}{kT}(\phi_3 - \phi_2)} = e^{\frac{q}{kT}(\phi_1 - \phi_6)} \, e^{\frac{q}{kT}\left[(\phi_{01} - \phi_{06}) + (\phi_{02} - \phi_{03})\right]}$$

$$e^{\frac{q}{kT}(\phi_3 - \phi_2)} = e^{\frac{q}{kT}(\phi_1 - \phi_6)} \, e^{\frac{q}{kT}(\phi_a + \phi_b)} = \frac{I_1 + y}{I_1 - y} e^{\frac{q}{kT}(\phi_a + \phi_b)}$$

where: $\phi_a = (\phi_{01} - \phi_{06})$

$$\phi_b = (\phi_{02} - \phi_{03}) \quad (8)$$

$\phi_c = (\phi_{04} - \phi_{05})$

Then:

$$I_2 = \frac{I_o + x}{1 + \frac{I_1 - y}{I_1 + y} e^{\frac{q}{kT}(\phi_a + \phi_b)}} \quad (9)$$

Similarly, we can derive expressions for the currents I_3, I_4 and I_5 as shown below:

$$I_3 = \frac{I_o + x}{1 + \frac{I_1 - y}{I_1 + y} e^{\frac{q}{kT}(\phi_a + \phi_b)}} \quad (10)$$

$$I_4 = \frac{I_o - x}{1 + \frac{I_1 - y}{I_1 + y} e^{\frac{q}{kT}(\phi_c - \phi_a)}} \quad (11)$$

$$I_5 = \frac{I_o - x}{1 + \frac{I_1 + y}{I_1 - y} e^{\frac{q}{kT}(\phi_a - \phi_c)}} \tag{12}$$

The multiplier output currents can now be obtained by noting:

$$\begin{aligned} I_{out\ 1} &= I_2 + I_4 \\ I_{out\ 2} &= I_3 + I_5 \end{aligned} \tag{13}$$

Which produces:

$$I_{out\ 1} = \frac{I_o + x}{1 + \frac{I_1+y}{I_1-y} e^{\frac{q}{kT}(\phi_a+\phi_b)}} + \frac{I_o - x}{1 + \frac{I_1-y}{I_1+y} e^{\frac{q}{kT}(\phi_c-\phi_a)}} \tag{14}$$

and

$$I_{out\ 2} = \frac{I_o + x}{1 + \frac{I_1-y}{I_1+y} e^{\frac{q}{kT}(\phi_a+\phi_b)}} + \frac{I_o - x}{1 + \frac{I_1+y}{I_1-y} e^{\frac{q}{kT}(\phi_a-\phi_c)}} \tag{15}$$

Ideal Multiplier Operation Ignoring Voltage Offsets

To get a feel for the multiplier operation, we will first simplify Equations (14) and (15) for the case where all offsets are zero. ($\phi_a = \phi_b = \phi_c = 0$):

$$I_{out\ 1} = \frac{I_o + x}{1 + \frac{I_1 + y}{I_1 - y}} + \frac{I_o - x}{1 + \frac{I_1 - y}{I_1 + y}} \qquad (16)$$

Reducing this algebraically, $I_{out\ 1}$ simplifies to:

$$I_{out\ 1} = I_o - \frac{xy}{I_1} \qquad (17)$$

Similarly:

$$I_{out\ 2} = I_o + \frac{xy}{I_1} \qquad (18)$$

Combining the Equations (17) and (18), the expression for the differential output current is:

$$\Delta I_{out} = (I_{out\ 2} - I_{out\ 1}) = \frac{-2xy}{I_1} \qquad (19)$$

From (19) we see that the differential output current is simply equal to twice the product of the two input signal currents x and y divided by the constant I_1. Thus, if offsets are ignored and if the transistors are designed to follow the ideal Ebers-Moll expression, we can expect nearly ideal multiplication to result.

Multiplier Responses Including Voltage Offsets

Starting from the expressions for the output currents including offsets Equations (14) and (15) and assuming the offsets are small compared with kT/q, we can simplify the offset terms as follows:

$$\begin{aligned} e^{\frac{q}{kT}(\phi_a+\phi_b)} &= 1 + \frac{q}{kT}(\phi_a+\phi_b)(1+\varepsilon_1) \\ e^{\frac{q}{kT}(\phi_c-\phi_a)} &= 1 + \frac{q}{kT}(\phi_c-\phi_a)(1+\varepsilon_2) \\ e^{\frac{q}{kT}(\phi_a-\phi_c)} &= 1 - \frac{q}{kT}(\phi_c-\phi_a)(1-\varepsilon_2) \\ e^{\frac{-q}{kT}(\phi_a+\phi_b)} &= 1 - \frac{q}{kT}(\phi_a+\phi_b)(1-\varepsilon_1) \end{aligned} \tag{20}$$

Then, (14) and (15) becomes:

$$\begin{aligned} I_{out\ 1} &= \frac{I_o + x}{1 + \frac{I_1+y}{I_1-y}(1+\varepsilon_1)} + \frac{I_o - x}{1 + \frac{I_1-y}{I_1+y}(1+\varepsilon_2)} \\ I_{out\ 2} &= \frac{I_o + x}{1 + \frac{I_1-y}{I_1+y}(1-\varepsilon_1)} + \frac{I_o - x}{1 + \frac{I_1+y}{I_1-y}(1-\varepsilon_2)} \end{aligned} \tag{21}$$

Again, combining the output current terms to produce the differential output current:

$$\Delta I_{out} = (I_{out\,2} - I_{out\,1}) \tag{22}$$

and assuming that ε_1 and ε_2 are much smaller than one. Equation (22) can be written:

$$\begin{aligned}\Delta I_{out} = {} & \frac{(I_o + x)\ (I_1 + y)}{2I_1}\left[1 + \frac{\varepsilon_1\ (I_1 - y)}{2I_1}\right] + \\ & \frac{(I_o - x)\ (I_1 - y)}{2I_1}\left[1 + \frac{\varepsilon_2\ (I_1 + y)}{2I_1}\right] - \\ & \frac{(I_o + x)\ (I_1 - y)}{2I_1}\left[1 - \frac{\varepsilon_2\ (I_1 + y)}{2I_1}\right] - \\ & \frac{(I_o - x)\ (I_1 + y)}{2I_1}\left[1 - \frac{\varepsilon_2\ (I_1 - y)}{2I_1}\right]\end{aligned} \tag{23}$$

which reduces to:

$$\begin{aligned}\Delta I_{out} = {} & \frac{2xy}{I_1} + \frac{\varepsilon_1}{2I_1^{\,2}}\left[(I_1 + x)(I_1 + y)(I_1 - y)\right] \\ & + \frac{\varepsilon_2}{2I_1^{\,2}}\left[(I_1 - x)(I_1 + y)(I_1 - y)\right]\end{aligned} \tag{24}$$

This is the general expression for the differential output current including the error terms which arise from the presence of voltage offsets. The error terms may be simplified further by substituting expressions for ε_1 and ε_2 from Equations (20) and (24). The result is:

$$\Delta I_{out} = \frac{2xy}{I_1} + \frac{(I_1^{\,2} - y^2)}{2I_1^{\,2}}\ \frac{q}{kT}\left[2x\ \phi_a + \phi_b(I_o + x) + \phi_c(I_o - x)\right] \tag{25}$$

where: $\phi_a = \phi_{01} - \phi_{06}$

$\phi_b = \phi_{02} - \phi_{03}$

$\phi_c = \phi_{04} - \phi_{05}$

Here we see that the differential output current may be written as the sum of a product term, a constant offset, a linear error term in x and error terms containing y^2 and y^2x as below:

$$\Delta I_{out} = \frac{2xy}{I_1} + K_o + K_1 x + K_2 y^2 + K_3 y^2 x \qquad (26)$$

where: $K_o = I_o \, (\phi_b + \phi_c) \, \frac{q}{2kT}$

$K_1 = (2\phi_a + \phi_b - \phi_c) \, \frac{q}{2kT}$

$K_2 = \frac{1}{{I_1}^2} \, K_o$

$K_3 = \frac{-K_1}{{I_1}^2}$

As an example of the effects of these error terms, consider first the case where y is equal to a constant, c_1. Then the output current contains terms only in x or constants as below:

$$\Delta I_o = (K_o + K_2 \, c_1) + x \left(\frac{2c_1}{I_1} + K_1 + K_3 {c_1}^2\right) \qquad (27)$$

As a second example, let x be constant, c_2 then we get terms in y and y^2:

$$\Delta I_o = y \, \frac{(2c_2)}{I_1} + (K_o + K_1 c_2) + y^2 \, (K_2 + K_3 c_2) \qquad (28)$$

Thus, we generally expect to see second harmonic error terms when y is applied and x is held constant. We do not expect to see any significant harmonics in the case where y is constant and x is applied alone.

As a final calculation, it can be shown using Equation (28) that the percent total harmonic distortion expected in y when x is held constant and y equals $A_y \cos \omega t$ is:

$$\%THD = \frac{I_o \phi_{os} \left(1 + \frac{x\phi_{os}}{I_o \phi_{os}}\right) A_y}{8 \frac{kT}{q} I_1 x}$$

where: $\phi_{os} = f(\phi_a, \phi_b)$

$\phi_{os} = f'(\phi_a, \phi_b)$

In general, the accuracy of the multiplier is measured by applying signals simultaneously to both the x and y inputs and measuring the deviation from the ideal product xy. Such a calculation can be made by using Equation (26) but will not be attempted here.

CIRCUIT DESCRIPTION

Figure 2 is the complete integrated circuit. R_x and R_y are the emitter resistors to the input differential amplifier pairs and perform the linearizing function on Q_7 and Q_8 for the X-input and Q_{14} and Q_{13} for the Y-input. R_x and R_y also set the overall gain of the multiplier. Q_{11}, Q_{12}, Q_{19}, Q_{20}, Q_{17} and Q_{18} form the lower current sources shown in Figure 1. Q_9, Q_{10} and Q_{15}, Q_{16} form constant-current Darlington input-pairs and increase the input impedance to about 1 Meg. ohm.

Figure 3 is the complete multiplier with the external components shown. Transistors Q_{21} and Q_{22} form a composite differential amplifier whose output is single-ended with respect to ground and responds only to the difference of currents from pins 2 and 14.[3] Q_{23} is a current source that sets the bias current in the collector of Q_{22}. Thus, the collector of Q_{22} is shifted negative in voltage to ground and is single-ended about ground.

The MC-1539 is an operational amplifier and provides impedance isolation and current drive capability for the output load. The Operational Amplifier is fed back for non-inverting, unity-gain multiplication and provides a very low output impedance with low distortion at the lower frequencies.

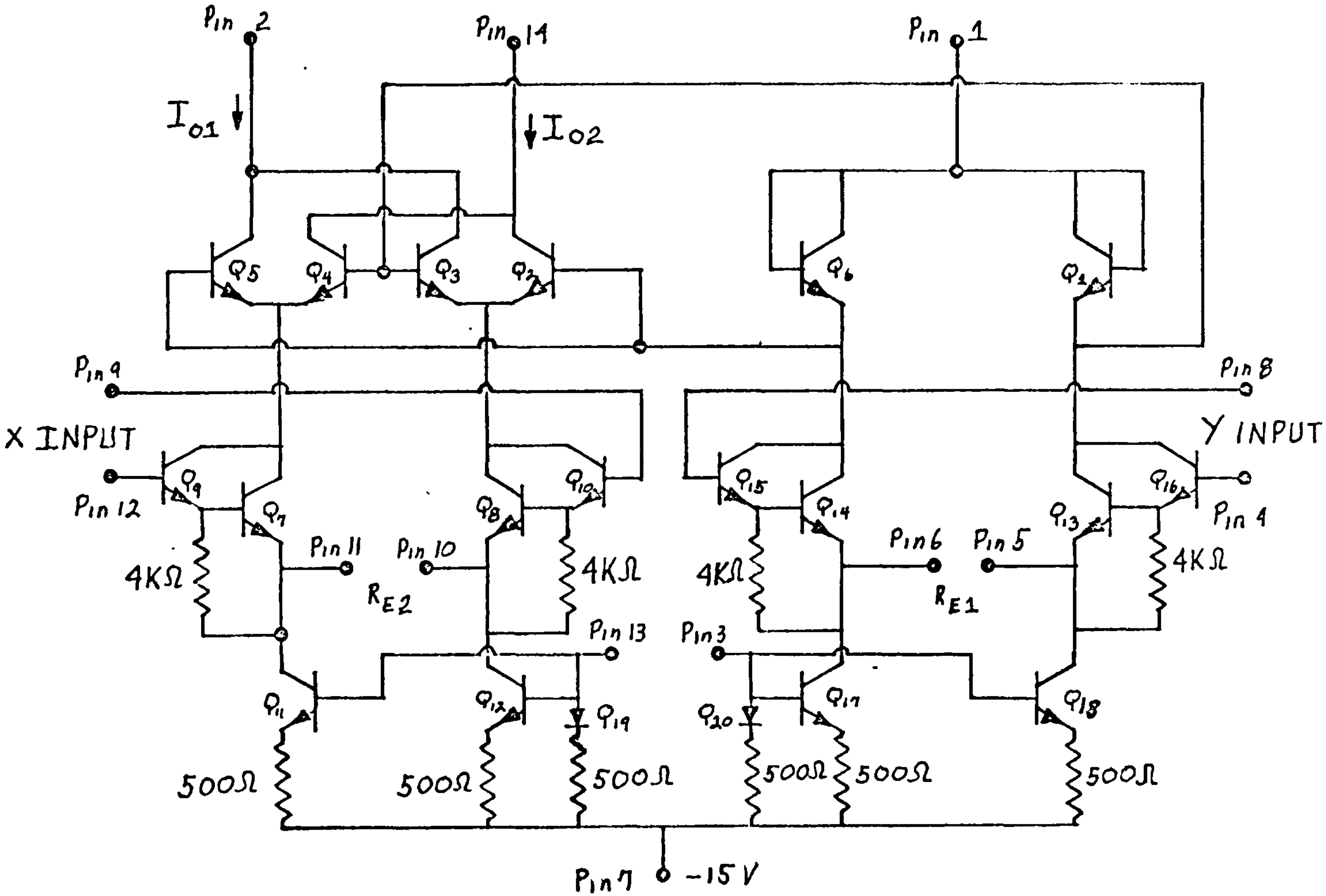

Figure 2

Monolithic Realization of Linear Multiplier

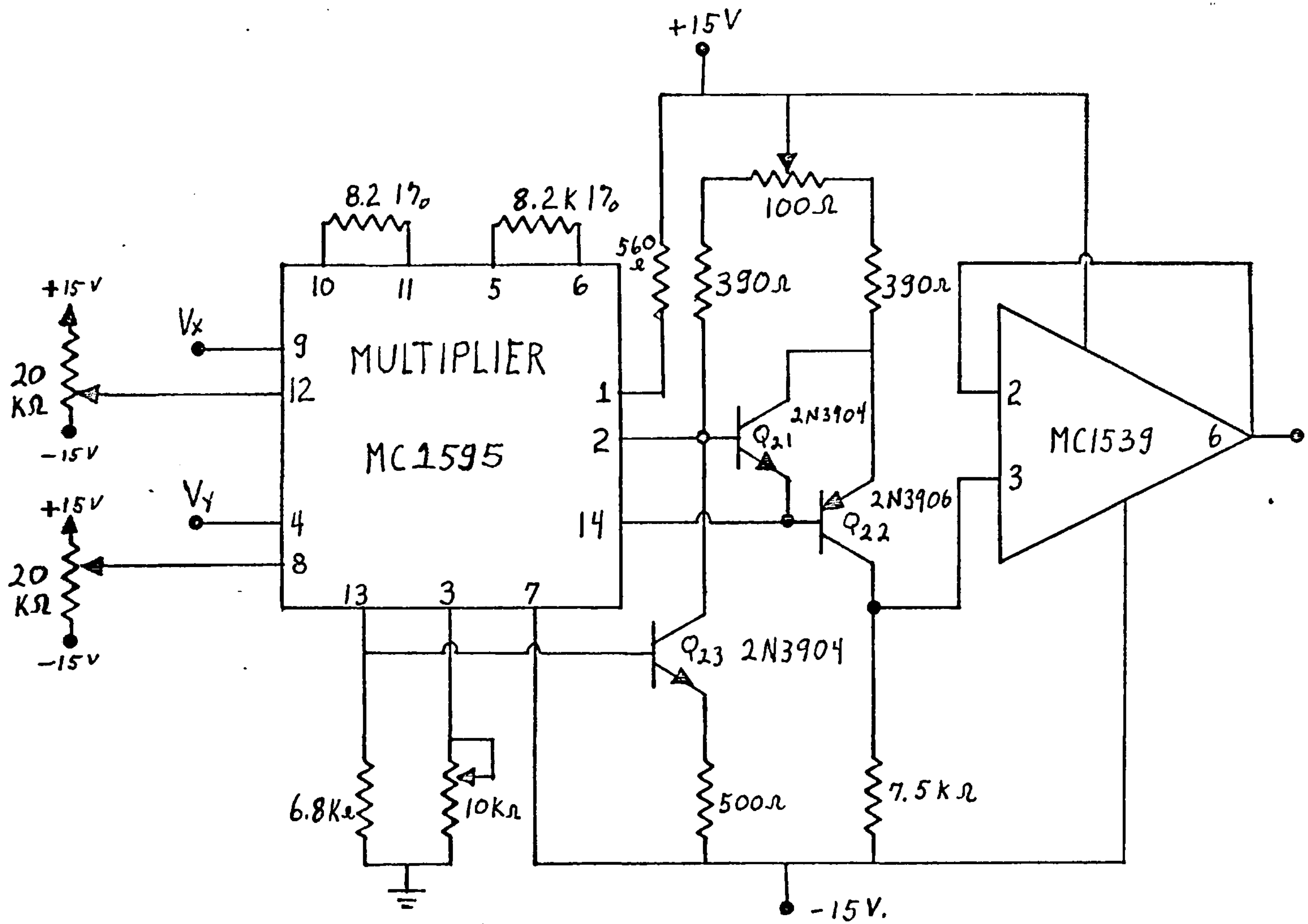

Figure 3. Schematic Diagram of Entire Multiplier with External Components

Circuit Analysis of Output Stage

The output stage,[3] see Figure 4, performs the function of differential input to single-ended output conversion and also shifts the output signal negative in voltage so that the output signal is referenced to ground. The output stage presents a very low impedance to the multiplier collectors at pins 2 and 14 and a current gain to the output load of unity. Thus, the single pole rolloff is caused by the stray capacitance, C_s, at the load resistor, R_L. The mismatch of the base-emitter voltages, ϕ_n and ϕ_p, is nulled by potentiometer R_4 which also adjusts the DC offset term to zero volts.

As one moves along the path from point "a" to point "b" through the base-emitter junctions of Q_{21} and Q_{22}, the voltage from point "a" to point "c" drops by ϕ_n volts and then increases from point "c" to point "b" by ϕ_p volts. If $\phi_n = \phi_p$, points "a" and "b" are at the same potential.

Assuming β_n and β_p are large (above $\beta = 100$), then:

$$V_a = (I_b + I_{02})R \quad \text{and}$$

$$V_b = (I_7 + I_8)R \quad \text{and, also}$$

$$I_{01} = I_7, \quad I_{out} = I_8$$

So,

$$(I_b + I_{02})R = (I_7 + I_8)R$$

$$I_b + I_{02} = I_7 + I_8$$

$$I_b + I_{02} - I_{01} = I_8 \; .$$

Finally, $I_{out} = I_b + I_{02} - I_{01}$, and the output current is equal to the difference of the input currents plus the constant term, I_b. If ϕ_n and ϕ_p track well with temperature, no temperature drift is introduced by the output stage.

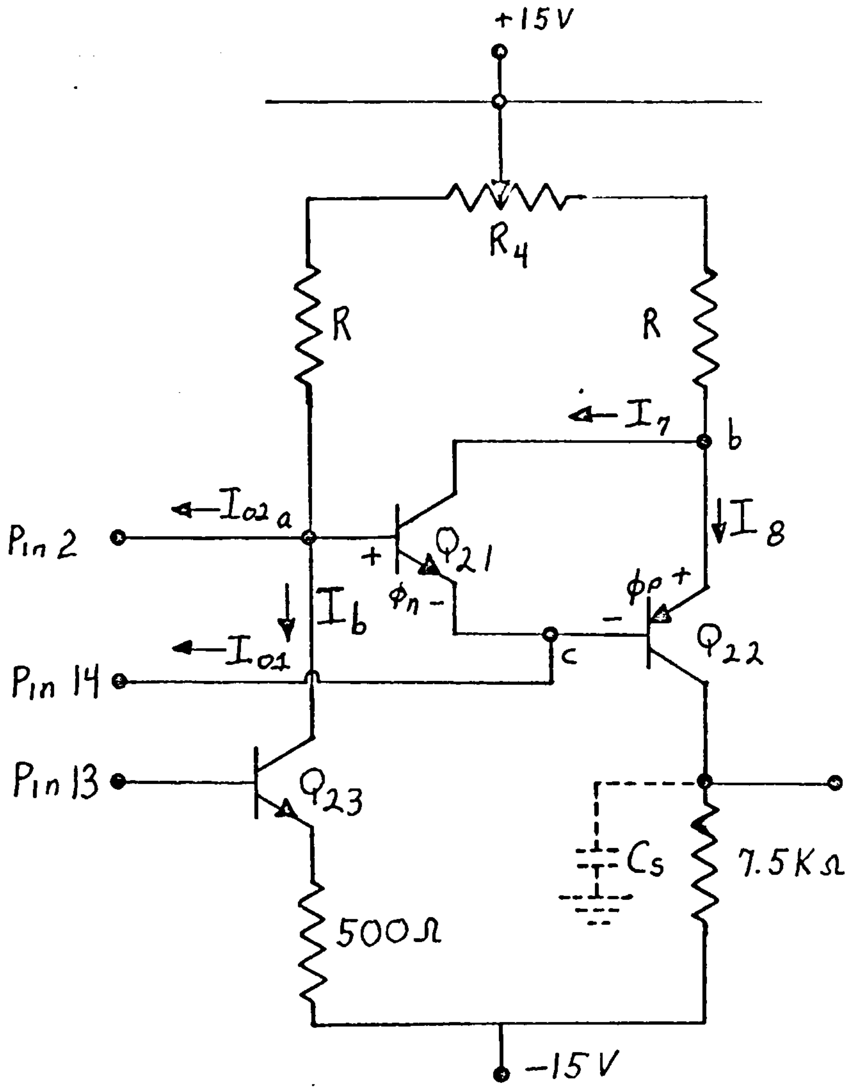

Figure 4. Circuit Diagram of Output Stage

ADJUSTMENT PROCEDURE

With Figure 3 implemented, the following steps should be followed:[4]

Step 1. Short the x and y inputs to ground first and adjust the offset adjustment potentiometer for zero volts to ground at the output of the amplifier.

Step 2. With the y input grounded, place + 10 V DC on the x input and adjust the y offset potentiometer until zero volts appear at the multiplier output.

Step 3. Short the x input to ground and place + 10 V DC on the y input. Then adjust the x offset potentiometer until the multiplier output is set at zero volts.

Step 4. Place + 10 V DC on both the x and y inputs and adjust the gain-adjust poteniometer until the output reads |10| V DC.

The above procedure should be repeated until no further changes are noticed at the output of the multiplier.

APPENDIX: RESULTS AND DEFINITIONS

Test Results

Figure 6 is the error curve presentation from the test set-up of Figure 5. The output of the test jig is the difference of the true input and the output of the multiplier. With the y-input set at +10 V DC, the multiplier applies the transfer junction 10X/10 + error to the input signal.

Figure 7 displays the output of the multiplier with cos ωt applied to both x and y inputs. The trigonometric identity $\cos^2 X = \frac{1}{2}\cos 2X + \frac{1}{2}$ is implemented directly.

Table 3 lists the results of DC measurements taken with various inputs.

The particular integrated circuit used for the collection of data herein was chosen as the most typical of the units available. Linearity error ranged from 0.1% to as high as 0.8% for the sample tested.

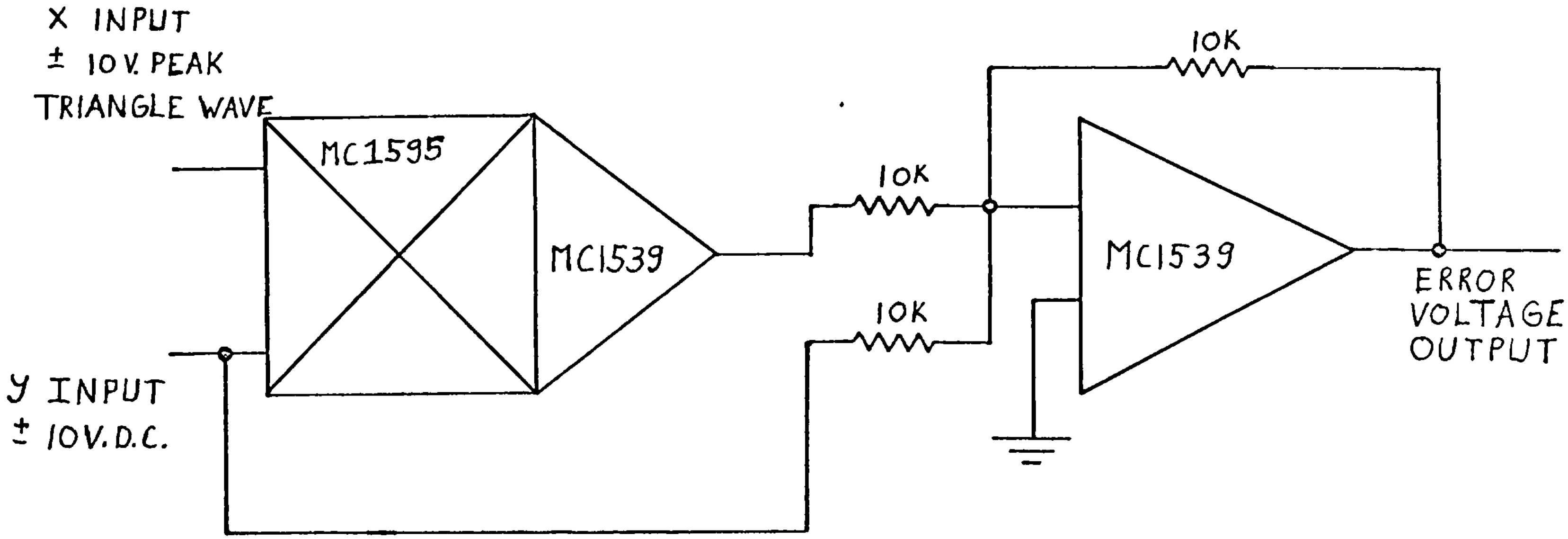

Figure 5

Test Circuit Block Diagram

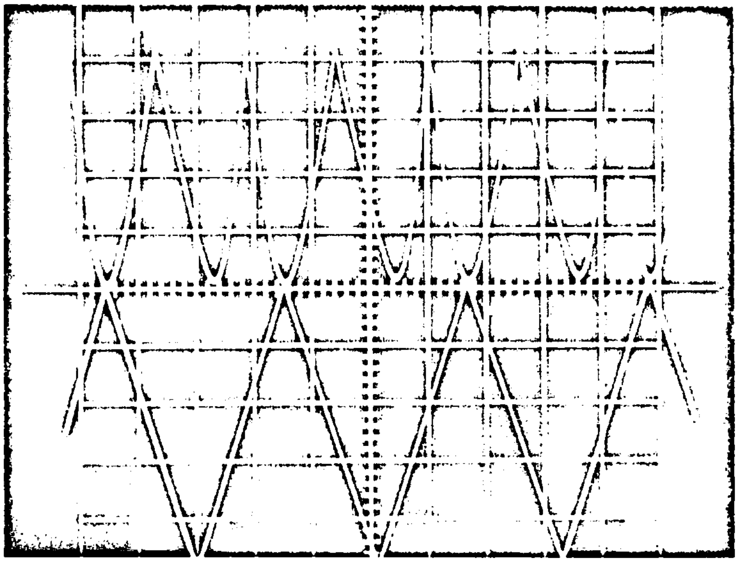

Figure 6

Typical Error Curve Presentation

Top: Output

Vertical Scale 10 mV/cm
Horizontal Scale 2.5 mSec/cm

Bottom: Input

Vertical Scale 4.0V/cm
Horizontal Scale 2.5 mSec/cm

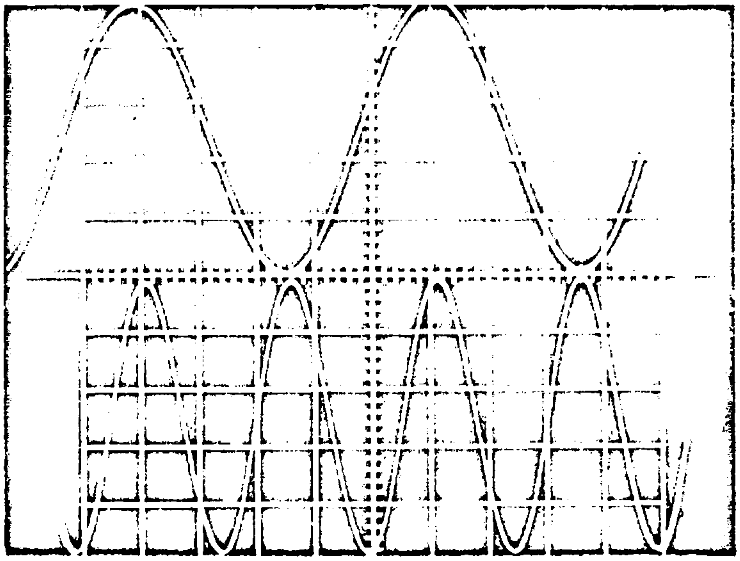

Figure 7

Output with cos ωt on Both Inputs

Top: Input

Vertical Scale 4.0 V/cm
Horizontal Scale 2.5 mSec/cm

Bottom: Output

Vertical Scale 2.0 V/cm
Horizontal Scale 2.5 mSec/cm

Table 3

Static Errors With Various Inputs

X Input (volts)	Y Input (volts)	Output (volts)
+10	+10	10.001
	-10	- 9.970
+8	+10	7.997
	-10	- 7.970
+6	+10	5.975
	-10	- 5.984
+4	+10	3.977
	-10	- 3.980
+2	+10	1.987
	-10	- 1.995
0	+10	0.001
	-10	- 0.001
-2	+10	- 2.005
	-10	2.004
-4	+10	- 3.993
	-10	3.998
-6	+10	- 5.989
	-10	5.995
-8	+10	- 7.987
	-10	7.980
-10	+10	-10.008
	-10	9.998

Definitions of Specifications

The accuracy of a multiplier in any given application can be determined if the specifications are understood.[5] Due to lack of standardized terminology, great care must be taken when comparing specifications of multipliers produced by different manufacturers.

Accuracy

This is the accuracy of the multiplier with dc on both inputs in the multiply mode at room temperature (25°C) and with rated supply voltage (±15 Vdc). It includes both offset error and linearity error. The accuracy is expressed as a percentage of full-scale output- not percentage of signal. For example, an accuracy of 0.5% with a ± 10V multiplier would mean:

$$E_o = \frac{XY}{10} \pm (0.5\%)\ (10\ V)$$

$$E_o = \frac{XY}{10} \pm \underset{\text{(error term)}}{0.05\ V}$$

Note that:

(a) When comparing specifications of units with ± 10V outputs, be sure that "full-scale" means 10V rather than 20V.

(b) "Linearity" and "accuracy" do not mean the same thing. In some cases, both terms are used. For example, a ± 10V Unit with ± 0.1% linearity and ± 2% accuracy means that:

$$E_o = (1 \pm 0.02)\,\frac{XY}{10} \pm (0.1\%)(10\ V)$$

$$E_o = \frac{XY}{10} \pm 0.02\,\frac{XY}{10} \pm 0.01\ V$$

The multiplier of this example would have an accuracy of ± 2.1%

Drift

Temperature changes cause two effects - an output offset and a variation in scale factor (multiplication constant). The temperature drift at the output is measured with both inputs at zero volts and is given in $mV/^{o}C$.

Scale-factor drift is measured by applying a 10V sine wave to one input and a constant 10 volts to the other input and observing the gain error. Scale-factor drift is given in $\%/^{o}C$ and is also averaged over the total temperature range. For example, a multiplier with an output offset drift of 0.5 $mV/^{o}C$ and a scale-factor drift of $0.2\%/^{o}C$ would have an output of:

$$E_o = (1 \pm 0.002\ \Delta T)\ \frac{XY}{10} \pm (0.5\ mV)\ \Delta T$$

where ΔT is the temperature deviation in degrees centigrade.

Frequency Response, 1% Absolute Error ($X = 10_{cos}\ \omega t$, $y = \pm\ 10$ VDC)

The useful bandwidth of a multiplier is the frequency band for which the multiplying error is not significantly larger than the static error. Thus, the frequency at which 1% absolute error occurs is usually the frequency of interest. The frequency response is the frequency at which an absolute error of 1% occurs with a full-scale signal applied and with the unit in the multiply mode.

The term "absolute error" is defined as the peak error due to both amplitude error and phase-shift error with a full-scale signal. This is also referred to as "dynamic error".

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