RC4200
Analog Multiplier

Description

The RC4200 analog multiplier has complete compensation for nonlinearity, the primary source of error and distortion. This multiplier also has three on-board operational amplifiers designed specifically for use in multiplier logging circuits. These amplifiers are frequency compensated for optimum AC response in a logging circuit, the heart of a multiplier, and can therefore provide superior AC response.

The RC4200 can be used in a wide variety of applications without sacrificing accuracy. Four-quadrant multiplication, two-quadrant division, square rooting, squaring and RMS conversion can all be easily implemented with predictable accuracy. The nonlinearity compensation is not just trimmed at a single temperature, it is designed to provide compensation over the full temperature range. This nonlinearity compensation combined with the low gain and offset drift inherent in a well-designed monolithic chip provides a very high accuracy and a low temperature coefficient.

The excellent linearity and versatility were achieved through circuit design rather than special grading or trimming, and therefore, unit cost is very low.

The RC4200 is ideal for use in low distortion audio modulation circuits, voltage-controlled active filters, and precision oscillators.

Features

♦ High accuracy
  Nonlinearity — 0.1%
  Temperature coefficient — 0.005%/°C
♦ Multiple functions
  Multiply, divide, square, square root, RMS-to-DC conversion, AGC and modulate/demodulate
♦ Wide bandwidth — 4 MHz
♦ Signal-to-noise ratio — 94 dB

RC4200

Absolute Maximum Ratings

Supply Voltage[1] .... -22V
Internal Power Dissipation[2] ....500 mV
Input Current .... -5 mA
Storage Temperature Range
    RM4200/4200A .... -65°C to +150°C
    RC4200/4200A .... -55°C to +125°C
Operating Temperature Range
    RM4200/4200A .... -55°C to +125°C
    RC4200/4200A ....0°C to +70°C

Notes:
1. For a supply voltage greater than -22V, the absolute maximum input voltage is equal to the supply voltage.
2. Observe package thermal characteristics.

Ordering Information

Part Number	Package	Operating Temperature Range
RC4200N	N	0°C to +70°C
RC4200AN	N	0°C to +70°C
RM4200D	D	-55°C to +125°C
RM4200AD	D	-55°C to +125°C
RM4200AD/883B	D	-55°C to +125°C

Notes:
/883B suffix denotes MIL-STD-883, Level B processing
N = 8-lead plastic DIP
D = 8-lead ceramic DIP

Connection Information

Pin	Function
1	I_2
2	V_{OS2}
3	$-V_S$
4	I_3 (Output)
5	I_4
6	GND
7	V_{OS1}
8	I_1

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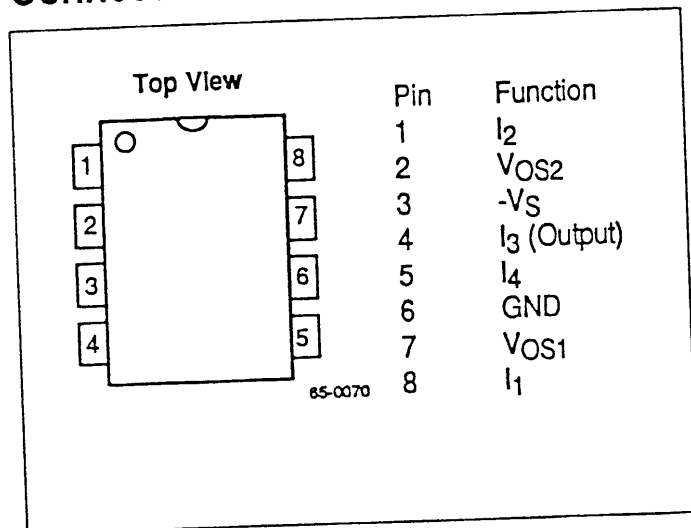
Thermal Characteristics
(Still air, soldered into PC board)

	8-Lead Plastic DIP	8-Lead Ceramic DIP
Max. Junction Temp.	+125°C	+175°C
Max. P_D T_A < 50°C	468 mW	833 mW
Therm. Res. θ_{JC}	—	45°C/W
Therm. Res. θ_{JA}	160°C/W	150°C/W
For T_A > 50°C Derate at	6.25 mW/°C	8.33 mW/°C

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For More Information, call 1-800-722

Electrical Characteristics

(Over operating temperature range, V_s = -15V unless otherwise noted)

Parameters	Test Conditions	4200A			4200			Units
		Min	Typ	Max	Min	Typ	Max	
Total Error as Multiplier	T_A = +25°C							
Untrimmed	Note 1			±2.0			±3.0	%
With External Trim			±0.2			±0.2		%
Versus Temperature			±0.005			±0.005		%/°C
Versus Supply (-9 to -18V)			±0.1			±0.1		%/V
Nonlinearity	50 µA $\leq I_{1,2,4} \leq$ 250 µA, T_A = +25°C (Note 2)			±0.1			±0.3	%
Input Current Range (I_1, I_2 and I_4)		1.0		1000	1.0		1000	µA
Input Offset Voltage	$I_1 = I_2 = I_4$ = 150 µA T_A = +25°C			±5.0			±10	mV
Input Bias Current	$I_1 = I_2 = I_4$ = 150 µA T_A = +25°C			300			500	nA
Average Input Offset Voltage Drift	$I_1 = I_2 = I_4$ = 150 µA			±50			±100	µV/°C
Output Current Range (I_3)	Note 3	1.0		1000	1.0		1000	µA
Frequency Response, -3dB point			4.0			4.0		MHz
Supply Voltage		-18	-15	-9.0	-18	-15	-9.0	V
Supply Current	$I_1 = I_2 = I_4$ = 150 µA T_A = +25°C			4.0			4.0	mA

Notes:
1. Refer to Figure 6 for example.
2. The input circuits tend to become unstable at I_1, I_2, I_4 < 50 µA and linearity decreases when I_1, I_2, I_4 > 250 µA (eq. @ $I_1 = I_2$ = 500 µA, nonlinearity error ≈ 0.5%).
3. These specifications apply with output (I_3) connected to an op amp summing junction. If desired, the output (I_3) at pin (4) can be used to drive a resistive load directly. The resistive load should be less than 700Ω and must be pulled up to a positive supply such that the voltage on pin (4) stays within a range of 0 to +5V.

RC4200

Functional Description

The RC4200 multiplier is designed to multiply two input currents (I_1 and I_2) and to divide by a third input current (I_4). The output is also in the form of a current (I_3). A simplified circuit diagram is shown in Figure 1. The nominal relationship between the three inputs and the output is:

$$I_3 = \frac{I_1 \, I_2}{I_4} \qquad (1)$$

The three input currents must be positive and restricted to a range of 1 μA to 1 mA. These currents go into the multiplier chip at op amp summing junctions which are nominally at zero volts. Therefore, an input voltage can be easily converted to an input current by a series resistor. Any number of currents may be summed at the inputs. Depending on the application, the output current can be converted to a voltage by an external op amp or used directly. This capability of combining input currents and voltages in various combinations provides great versatility in application.

Inside the multiplier chip, the three op amps make the collector currents of transistors Q1, Q2 and Q4 equal to their respective input currents (I_1, I_2, and I_4). These op amps are designed with current-source outputs and are phase-compensated for optimum frequency response

as a multiplier. Power drain of the op amps was minimized to prevent the introduction of undesired thermal gradients on the chip. The three op amps operate on a single supply voltage (nominally -15V) and total quiescent current drain is less than 4 mA. These special op amps provide significantly improved performance in comparison to 741-type op amps.

The actual multiplication is done within the log-antilog configuration of the Q1-Q4 transistor array. These four transistors, with associated proprietary circuitry, were specially designed to precisely implement the relationship

$$V_{BEN} = \frac{kT}{Q} \ln \frac{I_{CN}}{I_{SN}} \qquad (2)$$

Previous multiplier designs have suffered from an additional undesired linear term in the above equation; the collector current times the emitter resistance. The I_Cr_E term introduces a parabolic nonlinearity even with matched transistors. Raytheon has developed a unique and proprietary means of inherently compensating for this undesired I_Cr_E term. Furthermore, this Raytheon-developed circuit technique compensates linearity error over temperature changes. The nonlinearity versus temperature is significantly improved over earlier designs.

From equation (2) and by assuming equal transistor junction temperatures, summing base-to-emitter voltage drops around the transistor array yields:

$$\frac{KT}{q} \left[\ln \frac{I_1}{I_{S1}} = \ln \frac{I_2}{I_{S2}} - \ln \frac{I_3}{I_{S3}} - \ln \frac{I_4}{I_{S4}} \right] = 0 \qquad (3)$$

This equation reduces to:

$$\frac{I_1 I_2}{I_3 I_4} = \frac{I_{S1} I_{S2}}{I_{S3} I_{S4}} \qquad (4)$$

The ratio of reverse saturation currents, $I_{S1} I_{S2}/I_{S3} I_{S4}$, depends on the transistor matching. In a monolithic multiplier this matching is easily achieved and the ratio is very close to unity, typically 1.0 ±1%. The final result is the desired relationship:

$$I_3 = \frac{I_1 \, I_2}{I_4} \qquad (5)$$

The inherent linearity and gain stability combined with low cost and versatility makes this new circuit ideal for a wide range of nonlinear functions.
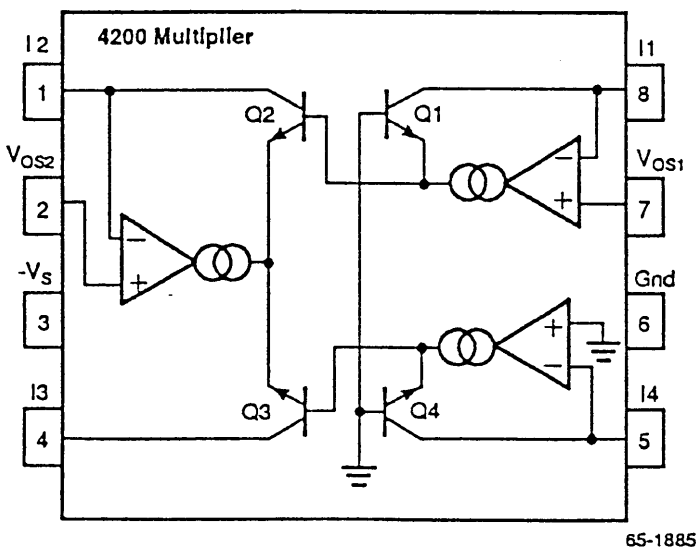

Figure 1. Functional Diagram

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RC4200

Extended Range

The input and output voltage ranges can be extended to include 0 and negative voltage signals by adding bias currents. The $R_S C_S$ filter circuits are eliminated when the input and biasing resistors are selected to limit the respective currents to 50 µA min. and 250 µA max.
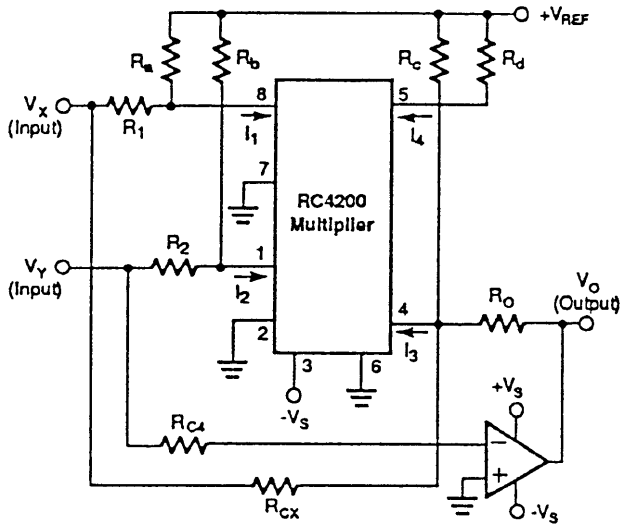
Extended Range Multiplier

Figure 5

65-1881

Resistors R_a and R_b extend the range of the V_X and V_Y inputs by picking values such that:

$$I_1(\text{min.}) = \frac{V_X(\text{min.})}{R_1} + \frac{V_{REF}}{R_a} = 50 \, \mu A,$$

and

$$I_1(\text{max.}) = \frac{V_X(\text{max.})}{R_1} + \frac{V_{REF}}{R_a} = 250 \, \mu A;$$

also

$$I_2(\text{min.}) = \frac{V_Y(\text{min.})}{R_2} + \frac{V_{REF}}{R_b} = 50 \, \mu A,$$

and

$$I_2(\text{max.}) = \frac{V_Y(\text{max.})}{R_2} + \frac{V_{REF}}{R_b} = 250 \, \mu A.$$

Resistor R_C supplies bias current for I_3 which allows the output to go negative.

Resistors R_{CX} and R_{CY} permit equation (6) to balance, ie.:

$$\left(\frac{V_X}{R_1} + \frac{V_{REF}}{R_a}\right) + \left(\frac{V_Y}{R_2} + \frac{V_{REF}}{R_b}\right) = \left(\frac{V_0}{R_0} + \frac{V_{REF}}{R_c} + \frac{V_X}{R_{CX}} + \frac{V_Y}{R_{CY}}\right)\left(\frac{V_{REF}}{R_d}\right)$$

$$\frac{V_Y V_X}{R_1 R_2} + \frac{V_X V_{REF}}{R_1 R_b} + \frac{V_Y V_{REF}}{R_2 R_a} + \frac{V_{REF}}{R_a R_b} =$$

$$\frac{V_0 V_{REF}}{R_0 R_d} + \frac{V_X V_{REF}}{R_{CX} R_d} + \frac{V_Y V_{REF}}{R_{CY} R_d} + \frac{V_{REF}^2}{R_c R_d}$$

Cross-Product Cancellation

Cross-products are a result of the $V_X V_R$ and $V_Y V_R$ terms. To the extend that: $R_1 R_b = R_{CX} R_d$ and $R_2 R_a = R_{CY} R_d$, cross-product cancellation will occur.

Arithmetic Offset Cancellation

The offset caused by the V_{REF}^2 term will cancel to the extent that: $R_a R_b = R_c R_d$, and the result is:

$$\frac{V_Y V_X}{R_1 R_2} = \frac{V_0 V_{REF}}{R_0 R_d} \quad \text{or} \quad V_0 = V_X V_Y K$$

$$\text{Where } K = \frac{R_0 R_d}{V_{REF} R_1 R_2}$$

Resistor Values

Inputs:

$$V_X(\text{min.}) \leq V_X \leq V_X(\text{max.})$$

$$\Delta V_X = V_X(\text{max.}) - V_X(\text{min.})$$

$$V_Y(\text{min.}) \leq V_Y \leq V_Y(\text{max.})$$

$$\Delta V_Y = V_Y(\text{max.}) = V_Y(\text{min.})$$

$$V_{REF} = \text{Constant} \, (+7V \text{ to } +18V)$$

$$K = \frac{V_0}{V_X V_Y} \quad \text{(Design Requirement)}$$

$$R_1 = \frac{\Delta V_X}{200 \, \mu A}, R_2 = \frac{\Delta V_Y}{200 \, \mu A}, R_d = \frac{V_{REF}}{250 \, \mu A}$$

$$R_a = \frac{\Delta V_X V_{REF}}{250 \, \mu A \Delta V_X - 200 \, \mu A \, V_X(\text{max.})}$$

$$R_b = \frac{\Delta V_Y V_{REF}}{250 \, \mu A \Delta V_Y - 200 \, \mu A \, V_Y(\text{max.})}$$

$$R_c = \frac{R_a R_b}{R_d}, R_{cx} = \frac{R_1 R_b}{R_d}, R_{cy} = \frac{R_2 R_a}{R_d}$$

$$R_0 = \frac{\Delta V_X \Delta V_Y K}{160 \, \mu A}$$

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Basic Circuits

Current Multiplier/Divider

The basic design criteria for all circuit configurations using the 4200 multiplier is contained in equation (1):

i.e., $$I_3 = \frac{I_1 I_2}{I_4}$$

The current-product-balance equation restates this as:
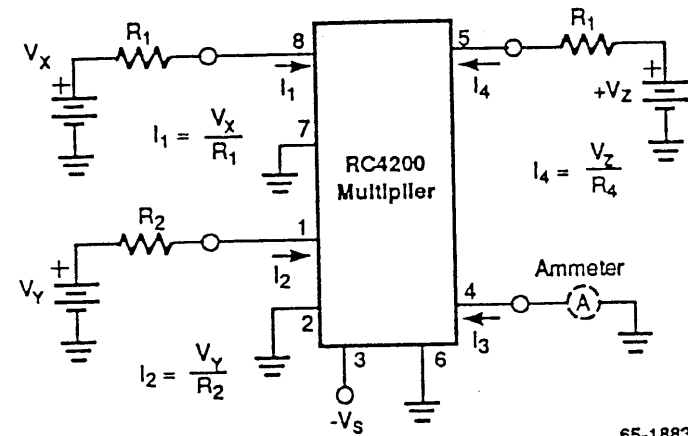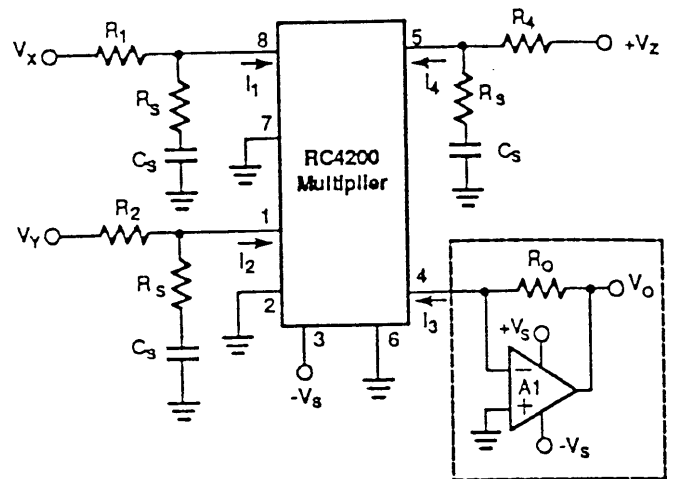
$$I_1 I_2 = I_3 I_4 \qquad (6)$$

Figure 2

Dynamic Range and Stability

The precision dynamic range for the 4200 is from +50 μA to +250 μA inputs for I_1, I_2 and I_4. Stability and accuracy degrade if this range is exceeded.

To improve the stability for input currents less than 50 μA, filter circuits ($R_S C_S$) are added to each input (see Figure 3).

R_S = 10k Ohms
C_S = 0.005 μF

Amplifier A1 is used to convert the I_3 current to an output voltage.

Multiplier: V_Z = constant $\neq 0$
Divider: V_Y = constant $\neq 0$

Figure 3

Voltage Multiplier/Divider
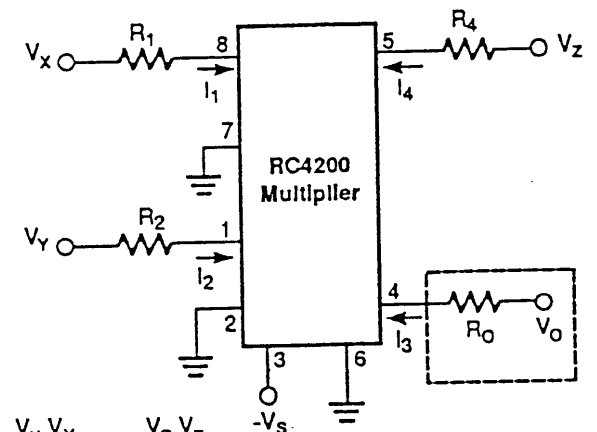
$$\frac{V_X V_Y}{R_1 R_2} = \frac{V_O V_Z}{R_O R_4}$$

Solving for $V_O = \dfrac{V_X V_Y}{V_Z} \cdot \dfrac{R_O R_4}{R_1 R_2}$

For a multiplier circuit $V_Z = V_R$ = constant

Therefore: $V_O = V_X V_Y K$ where $K = \dfrac{R_O R_4}{V_R R_1 R_2}$

For a divider circuit $V_Y = V_R$ = constant

Therefore: $V_O = \dfrac{V_X}{V_Z} K$ where $K = \dfrac{V_R R_O R_4}{R_1 R_2}$

Figure 4

Multiplying Circuit Offset Adjust

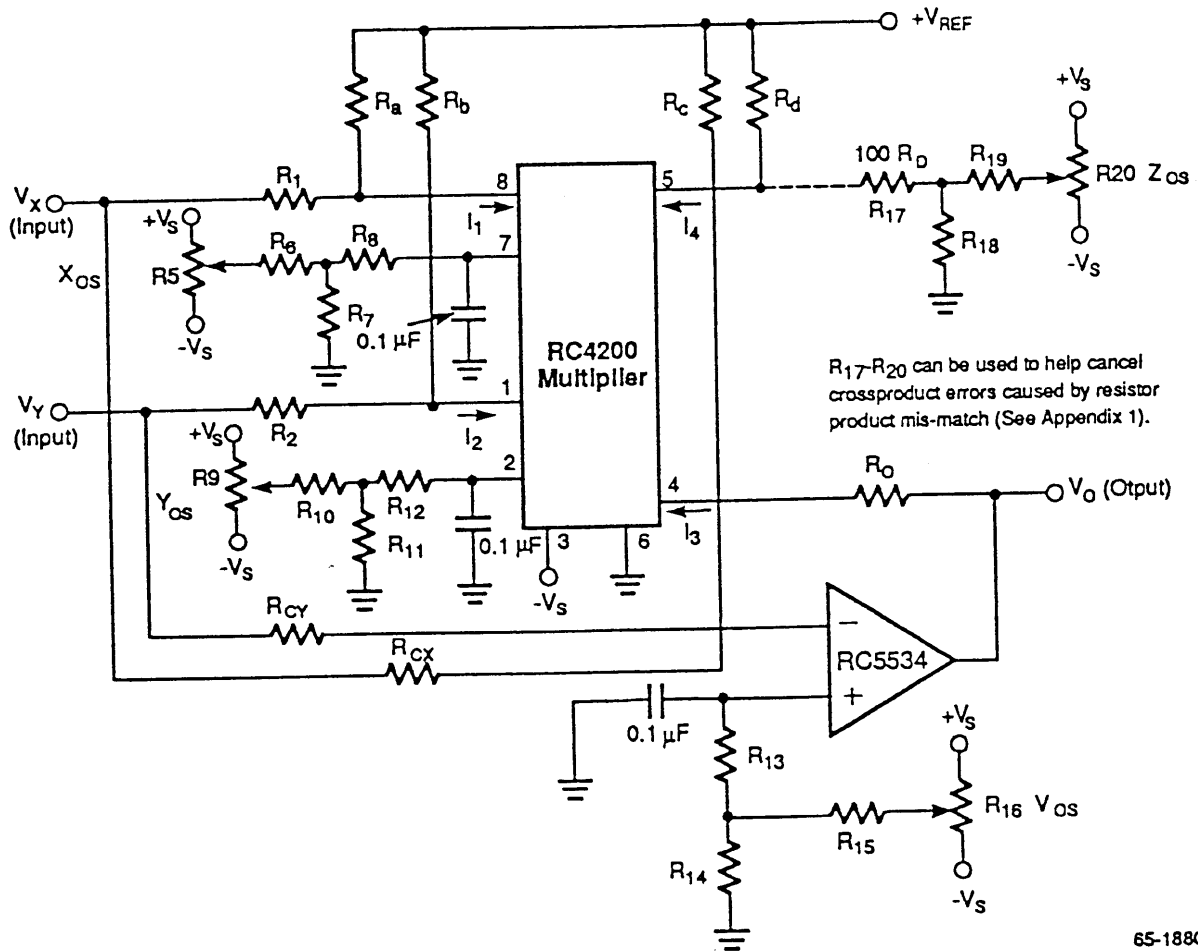$$10K \leq R_5 = R_9 = R_{16} \leq 50K$$
$$R_7 = R_{11} = R_{14} = 100\Omega$$
$$R_6 = R_{10} = 100\Omega\,(V_S/0.05)$$
$$R_{15} = 100\Omega\,(V_S/0.10)$$
$$R_8 = R_1 \,||\, R_a$$
$$R_{12} = R_2 \,||\, R_b$$
$$R_{13} = R_0 \,||\, R_C \,||\, R_{CX} \,||\, R_{CY}$$

R$_{17}$-R$_{20}$ can be used to help cancel crossproduct errors caused by resistor product mis-match (See Appendix 1).

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Procedure:

1. Set all trimmer pots to 0V on the wiper.

2. Connect V_X input to ground. Put in a full scale square wave on V_Y input. Adjust $X_{OS}(R_5)$ for no square wave on V_0 output (adjust for 0 feedthrough).

3. Connect V_Y input to ground. Put in a full scale square wave on V_X input. Adjust $Y_{OS}(R_9)$ for no square wave on V_0 output (adjust for 0 feedthrough).

4. Connect V_X and V_Y to ground. Adjust $V_{OS}(R_{16})$ for 0V on V_0 output.

Figure 6

Extended Range Divider
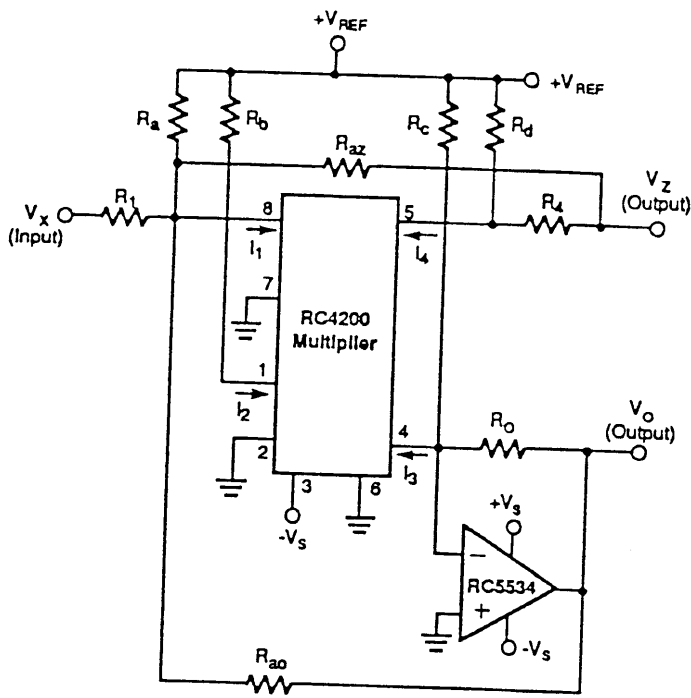
Figure 7

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Note: it is necessary to match the above resistor cross products to within the amount of error tolerable in the output offset, i.e., with a 10V F.S. output, 0.1% resistor cross-product match will give 0.1% x 10V = 10 mV untrimmable output offset voltage.

Resistor Values

Inputs:

$$V_X(min.) \leq V_X \leq V_X(max.)$$
$$\Delta V_X = V_X(max.) = V_X(min.)$$
$$V_Z(min.) \leq V_Z \leq V_Z(max.)$$
$$\Delta V_Z = V_Z(max.) = V_Z(min.)$$
$$V_{REF} = Constant \ (+7V \ to \ +18V)$$

Outputs:

$$V_0(min.) \leq V_0 \leq V_0(max.)$$
$$\Delta V_0 = V_0(max.) = V_0(min.)$$

$$K = \frac{V_0 V_Z}{V_X} \ (Design \ Requirement)$$

$$R_0 = \frac{\Delta V_0}{750 \ \mu A} \ , \ R_b = \frac{V_{REF}}{250 \ \mu A} \ , \ R_4 = \frac{\Delta V_Z}{200 \ \mu A}$$

$$R_c = \frac{\Delta V_0 V_{REF}}{750 \ \mu A \Delta V_0 - 700 \ \mu A \ V_0(max.)}$$

$$R_d = \frac{\Delta V_Z V_{REF}}{250 \ \mu A \Delta V_Z - 200 \ \mu A \ V_Z(max.)}$$

$$R_a = \frac{R_c R_d}{R_b} \ , \ R_{az} = \frac{R_c R_4}{R_b} \ , \ R_{ao} = \frac{R_0 R_d}{R_b}$$

$$R_1 = \frac{\Delta V_0 \Delta V_Z}{600 \ \mu A K}$$

As with the extended range multiplier, resistors R_{az} and R_{ao} are added to cancel the cross-product error caused by the biasing resistors, i.e.,

$$\left(\frac{V_X}{R_1} + \frac{V_0}{R_{ao}} + \frac{V_Z}{R_{az}} + \frac{V_{REF}}{R_a} \right) \left(\frac{V_{REF}}{R_b} \right) = \left(\frac{V_0}{R_0} + \frac{V_{REF}}{R_c} \right) \left(\frac{V_Z}{R_4} + \frac{V_{REF}}{R_d} \right)$$

$$\frac{V_X V_{REF}}{R_1 R_b} + \frac{V_0 V_{REF}}{R_{ao} R_b} + \frac{V_Z V_{REF}}{R_{az} R_b} + \frac{V_{REF}^2}{R_a R_b} =$$

$$\frac{V_0 V_Z}{R_0 R_4} + \frac{V_0 V_{REF}}{R_0 R_d} + \frac{V_Z V_{REF}}{R_4 R_c} + \frac{V_{REF}^2}{R_c R_d}$$

To cancel cross-product and arithmetic offset:

$$R_{ao} R_b = R_0 R_d, \ R_{az} R_b = R_4 R_c \ and \ R_a R_b = R_c R_d$$

and the result is:

$$\frac{V_X V_{REF}}{R_1 R_b} = \frac{V_0 V_Z}{R_0 R_4} \quad or \quad V_0 = \frac{V_X}{V_Z K}$$

$$where \ K = \frac{V_{REF} R_0 R_4}{R_1 R_b}$$

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Extended Range Circuit Errors

The extended range configurations have a disadvantage in that additional accuracy errors may be introduced by resistor product mismatching.

Multiplier (Figure 6)

An error in resistor product matching will cause an equivalent feedthrough or output offset error:

1. $R_1R_b = R_{CX}R_d \pm \alpha$, V_X feedthrough $(V_Y = 0) = \pm \alpha V_X$

2. $R_2R_a = R_{CY}R_d \pm \beta$, V_Y feedthrough $(V_X = 0) = \pm \beta V_Y$

3. $R_aR_b = R_CR_d \pm \gamma$, V_0 offset $(V_X = V_Y = 0) = \pm \gamma V_{REF}^*$

*Output offset errors can always be trimmed out with the output op amp offset adjust, V_{OS} (R_{16}).

Reducing Mismatch Errors (Figure 6)

You need not use .01% resistors to reduce resistor product mismatch errors. Here are a couple of ways to squeeze maximum accuracy out of the extended range multiplier (see Figure 6) using 1% resistors.

Method #1

V_X feedthrough, for example, occurs when $V_Y = 0$ and $V_{OSY} \neq 0$. This V_X feedthrough will equal $\pm V_X V_{OSY}$. Also, if $V_{OSZ} \neq 0$, there is a V_X feedthrough equal to $\pm V_X V_{OSZ}$. A resistor-product error of α will cause a V_X feedthrough of $\pm \alpha V_X$. Likewise, V_Y feedthrough errors are: $\pm V_Y V_{OSX}$, $\pm V_Y V_{OSZ}$ and $\pm \beta V_Y$.

Total feedthrough:
$\pm V_X V_{OSY} \pm V_Y V_{OSX} \pm \alpha V_X \pm \beta V_Y \pm (V_X + V_Y) V_{OSZ}$

By carefully adjusting $X_{OS}(R_5)$, $Y_{OS}(R_9)$ and $Z_{OS}(R_{20})$ this equation can be made to very nearly equal zero and the feedthrough error will practically disappear.

A residual offset will probably remain which can be trimmed out with $V_{OS}(R_{16})$ at the output of amp.

Method #2

Notice that the ratios of $R_1R_b : R_{CX}R_d$ and $R_2R_a : R_{CY}R_d$ are both dependent of R_d, also that R_1, R_2, R_a and R_b are all functions of the maximum input requirements. By designing a multiplier for the same input ranges on both V_X and V_Y then $R_1 = R_2$, $R_{CX} = R_{CY}$ and $R_a = R_b$. (Note: It is acceptable to design a four quadrant multiplier and use only two quadrants of it.)

Select R_d to be 1% or 2% below (or above) the calculated value. This will cause α and β to both be positive (or negative) by nearly the same amount. Now the effective value of R_d can be trimmed with an offset adjustment $Z_{OS}(R_{20})$ on pin 5.

This technique will cause: 1) a slight gain error which can be compensated for with the R_0 value, and 2) an output offset error that can be trimmed out with $V_{OS}(R_{16})$ on the output op amp.

Extended Range Divider (Figure 8)

The only cross-product error of interest is the V_Z feedthrough $(V_X = 0$ and $V_{OSX} \neq 0)$ which is easily adjusted with $X_{OS}(R_5)$.

Resistor product mismatch will cause scaling errors (gain) that could be a problem for very low values of V_Z. Adjustments to $Y_{OS}(R_{18})$ can be made to improve the high gain accuracy.

Square Root and Squaring (Figures 10 and 12)

These circuits are functions of single variables so feedthrough, as such, is not a consideration. Cross product errors will effect incremental accuracy that can be corrected with $Y_{OS}(R_{14})$ or $Z_{OS}(R_{10})$.

RC4200

Appendix 2 — Applications

Design Considerations for RMS-to-DC Circuits

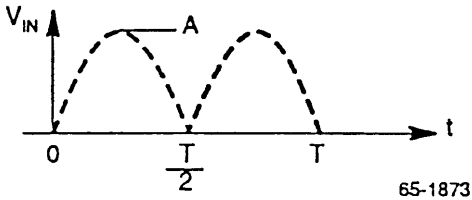#### Average Value

Consider $V_{in} = A\sin\omega t$. By definition,

$$V_{AG} = \frac{2}{T}\int_0^{\frac{T}{2}} V_{IN}dt$$

Where T = Period

$$\omega = 2\pi f$$
$$= \frac{2\pi}{T}$$

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$$V_{AG} = \frac{2}{T}\int_0^{\frac{T}{2}} A\sin\omega t\, dt$$

$$= \frac{2A}{T}\left[-\frac{1}{\omega}\cos\omega t\right]_0^{\frac{T}{2}}$$

$$= \frac{2A}{2\pi}\left[-\cos(\pi) + \cos(0)\right]$$

Average Value of $A\sin\omega t$ is $\frac{2}{\pi}A$

RMS Value

Again, consider $V_{IN} = A\sin\omega t$

$$V_{rms} = \sqrt{V_{AVG}} = \sqrt{\frac{1}{T}\int_0^T [V_{IN}]^2 dt}$$

V_{rms} for $A\sin\omega t\, dt$:

$$V_{rms} = \sqrt{\frac{1}{T}\int_0^T A^2\sin^2\omega t\, dt}$$

$$V_{rms} = \sqrt{\frac{A^2}{T}\int_0^T \left[\frac{1}{2} - \frac{1}{2}\cos 2\cos 2\omega t\right]dt}$$

$$V_{rms} = \sqrt{\frac{A^2}{2}\left[\frac{T}{2} - \frac{1}{4\omega}\sin 2\omega t\right]_0^T}$$

$$V_{rms} = \sqrt{\frac{A^2}{2}\left[\frac{T}{2}\right]}$$
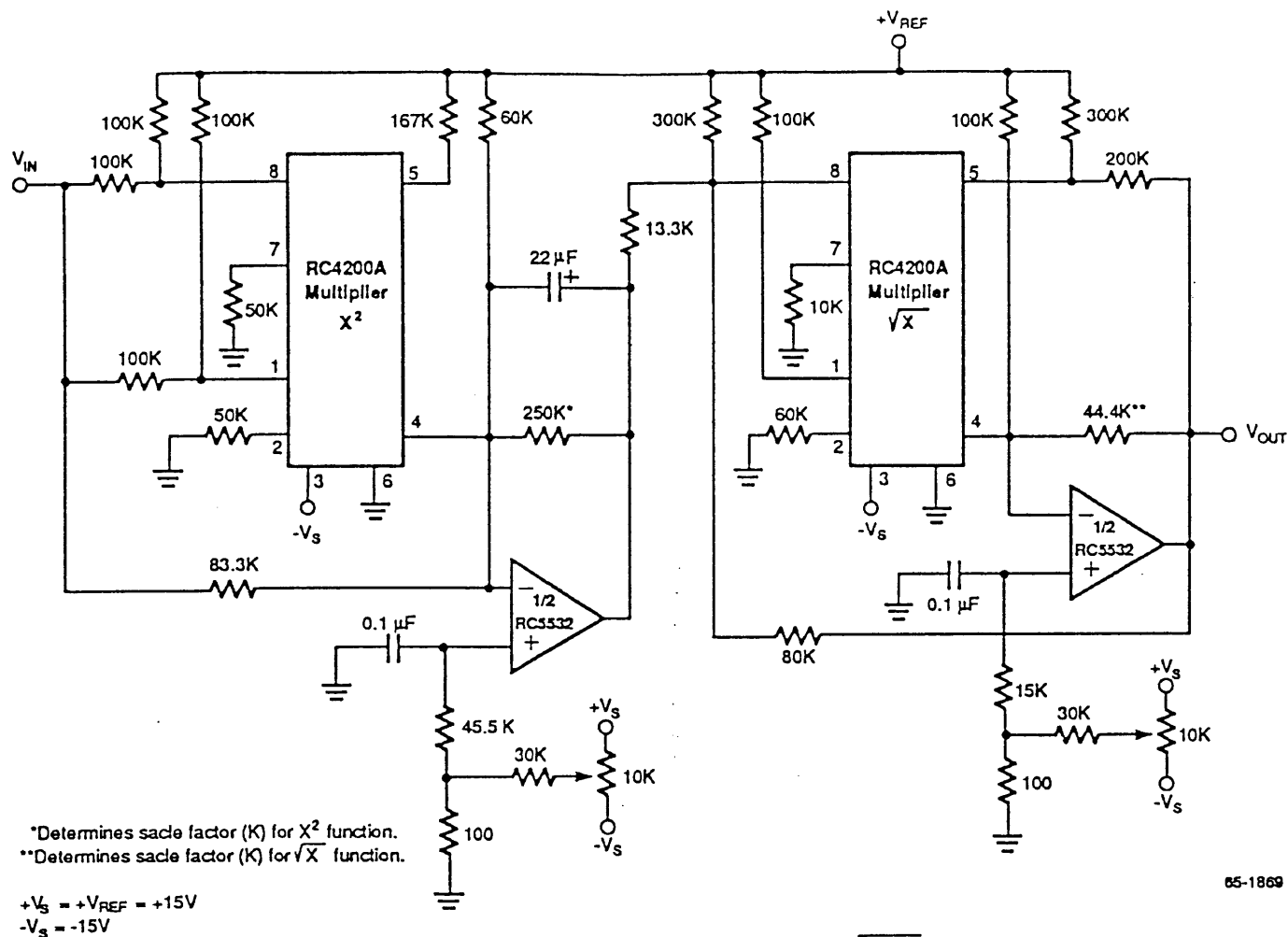
$$V_{rms} = \sqrt{\frac{A^2}{2}}$$

therefore, the rms value of $A\sin\omega t$ becomes:

$$V_{rms} = \frac{A}{\sqrt{2}}$$

RMS Value for Rectified Sine Wave

Consider $V_{in} = |A\sin\omega t|$, a rectified wave. To solve, integrate of each half cycle.

i.e. $\frac{1}{T}\int_0^T V_{in}^2\, dt =$

$$\frac{1}{T}\left[\int_0^{\frac{T}{2}} A^2\sin^2\omega t\, dt + \int_{\frac{T}{2}}^T (-A\sin\omega t)^2 dt\right]$$

This is the same as $\frac{1}{1}\int_0^T A^2\sin^2\omega t\, dt$

so, $|A\sin\omega t|_{rms} = A\sin\omega t_{rms}$

Practical Consideration: $|A\sin\omega t|$ has high-order harmonics; $A\sin\omega t$ does not. Therefore, non-ideal integrators may cause different errors for two approaches.

(a)

(b)   65-1872

$$\text{Avg}\left[\frac{V_{IN}^2}{V_0}\right] = V_0$$

$$\text{Implies } V_0 = \sqrt{\text{Avg}(|V_{IN}^2|)}$$
$$V_0 = \sqrt{\text{Avg } V_{IN}^2}$$

Figure 15

Figure 16. RMS to DC Converter $V_{OUT} = \sqrt{\overline{\int V_{in}^2}}$

*Determines sacle factor (K) for X^2 function.
**Determines sacle factor (K) for $\sqrt{X}$ function.

$+V_s = +V_{REF} = +15V$
$-V_s = -15V$

65-1869

Amplitude Modulator with A.G.C. (Figure 17)

In many AC modulator applications, unwanted output modulation is caused by variations in carrier input amplitude. The versatility of the RC4200 multiplier can be utilized to eliminate this undesired fluctuation. The extended range multiplier circuit (Figure 5) shows an output amplitude inversely proportional to the reference voltage V_{REF}.

$$i.e., V_0 = \frac{V_X V_Y}{V_{REF}} \frac{R_0 R_d}{R_1 R_2}$$

By making V_{REF} proportional to V_Y (where V_Y is the carrier input) such that:

$$V_{REF} = V_H = \int(|V_Y|),$$

Then the denominator becomes a variable value that automatically provides constant gain, such that the modulating input (V_X) modulates the carrier (V_Y) with a fixed scale factor even though the carrier varies in amplitude.

If V_H is made proportional to the average value of $A\sin\omega t$ (i.e., $2A/\pi$) and scaled by a value of $\pi/2$ then:

$$V_H = A$$

and if: V_X = Modulating input (V_M)

and: V_Y = Carrier input ($A\sin\omega t$)

Then: $V_0 = K V_M \sin\omega t$ where $K = \frac{R_0 R_d}{R_1 R_2}$

The resistor scaling is determined by the dynamic range of the carrier variation and modulating input.

The resistor values are solved, as with the other extended range circuits, in terms of the input voltages.

Input voltages:
   Modulation voltage (V_M): $0 \le V_M \le V_X(max.)$
   Carrier (V_Y): $V_Y = A\sin\omega t$
   Carrier amplitude fluctuation (ΔA):
   $A(min.) \sin t \le V_Y \le A(max.) \sin\Omega\omega t$
Dynamic Range (N): $A(max.)/A(min.)$,
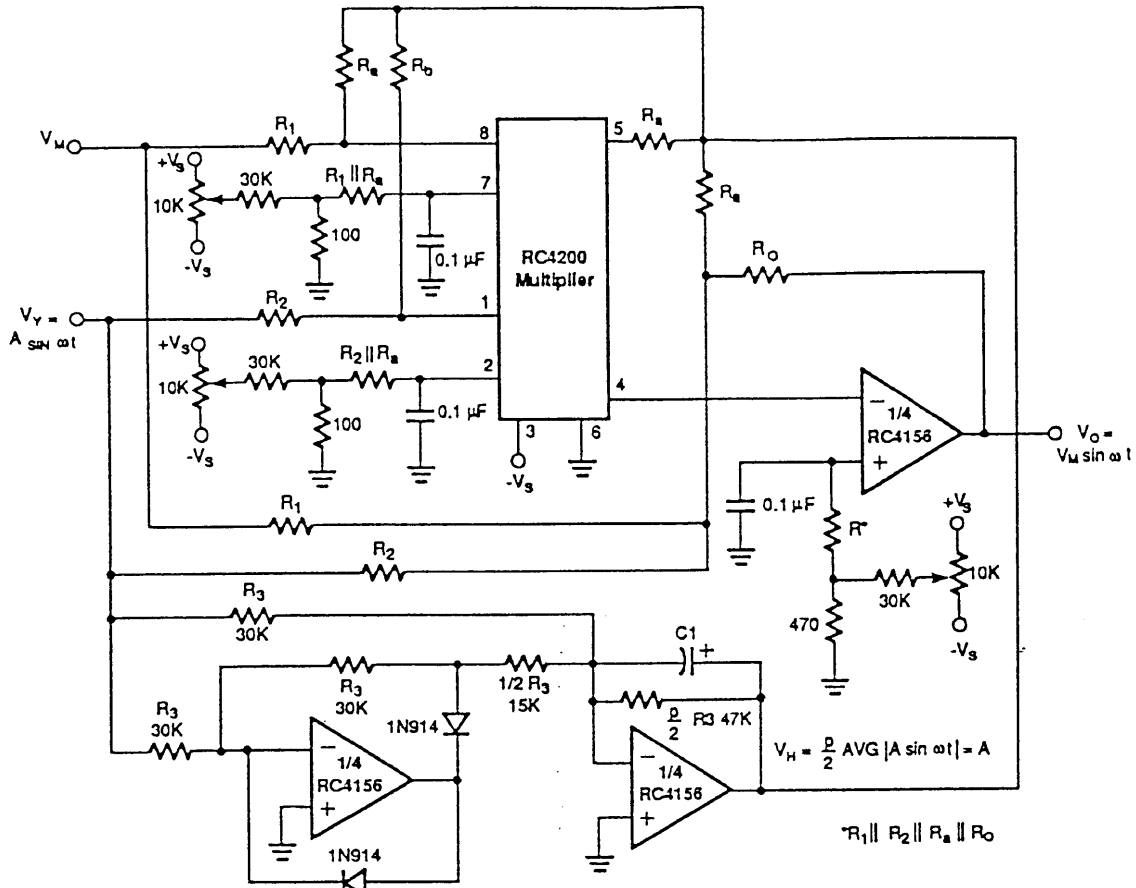$A(max.) = V_H(max.)$ and $(A(min.) = VH(min.)$

Figure 17. Amplitude Modulator with A.G.C.

65-1866

The maximum and minimum values for I_1 and I_2 lead to:

$$I_1(\text{max.}) = \frac{V_X(\text{max.})}{R_1} + \frac{V_H(\text{max.})}{R_a} = 250 \ \mu A$$

$$I_1(\text{min}) = \frac{V_H(\text{min.})}{R_a} = 50 \ \mu A \quad V_M(\text{min.}) = 0$$

$$I_2(\text{max}) = \frac{A(\text{max.})}{R_2} + \frac{V_H(\text{max.})}{R_a} = 250 \ \mu A$$

$$I_2(\text{min.}) = \frac{V_H(\text{min.})}{R_a} = 50 \ \mu A$$

For a dynamic range of N, where

$$N = \frac{A(\text{max.})}{A(\text{min.})} < 5,$$

These equations combine to yield:

$$R_1 = \frac{V_X(\text{max.})}{(5-N)50\mu A}, R_2 \frac{A(\text{max.})}{(5-N)50\mu A},$$

$$R_a = \frac{A(\text{min.})}{50\mu A} \text{ and } R_O = K \frac{R_1 R_2}{R_a}$$

Example #1

$V_Y = A\sin\omega t \ 2.5V \le A \le 10V$, therefore N = 4
$0V \le V_M \le 10V$, therefore $V_X(\text{max.}) = 10V$
K = 1, therefore $V_0 = V_M \sin\omega t$

$$R_1 = \frac{V_X(\text{max.})}{50\mu A} = \frac{10V}{50\mu A} = 200K$$

$$R_1 = \frac{A(\text{max.})}{50\mu A} = \frac{10V}{50\mu A} = 200K$$

$$R_a = \frac{A(\text{min.})}{50\mu A} = \frac{2.5V}{50\mu A} = 50K$$

$$R_O = K \frac{R_1 R_2}{R_a} = 1 \frac{200K \times 200K}{50K} = 800K$$

Example #2

$V_Y = A\sin\omega t \ 3 \le A \le 6$, therefore N = 2
$0V \le V_M \le 8V$, therefore $V_X(\text{max.}) = 8V$
K = .2, therefore $V_0 = .2 \ V_M \sin\omega t$
so:
$R_1 = 53.3K$, $R_2 = 40K$
$R_a = 60K$ and $R_O = 7.11K$

$$R_s = 10K, C_s = 0.005\,\mu F$$

$$V_s = \frac{V_X V_Y}{V_Z} K$$
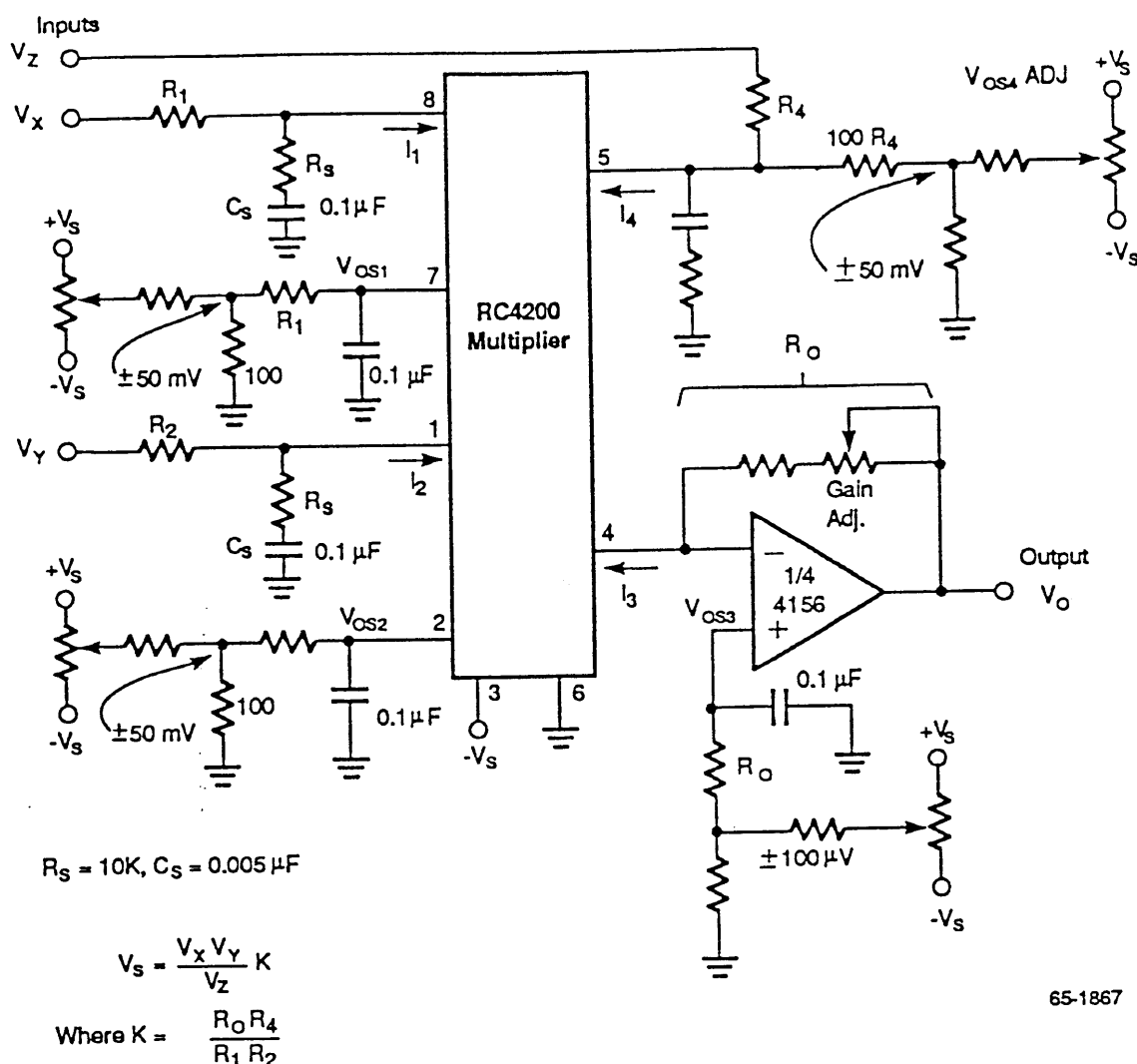
$$\text{Where } K = \frac{R_O R_4}{R_1 R_2}$$

Figure 18. First Quadrant Multiplier/Divider

Limited Range, First Quadrant Applications

The following circuit has the advantage that cross-product errors are due only to input offsets and nonlinearity error is slightly less for lower input currents.

The circuit also has no standby current to add to the noise content, although the signal-to-noise ratio worsens at very low input currents (1-5 µA) due to the noise current of the input stages.

The $R_S C_S$ filter circuits are added to each input to improve the stability for input currents below 50 µA.

Caution

The bandpass drops off significantly for lower currents (<50 µA) and non-symmetrical rise and fall times can cause second harmonic distortion.

Thermal Symmetry
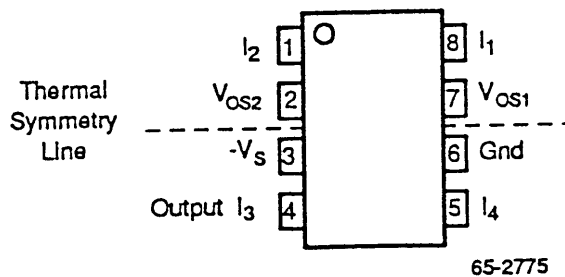
The scale factor is sensitive to temperature gradients across the chip in the lateral direction. Where possible, the package should be oriented such that sources generating temperature gradients are located physically on the line of thermal symmetry. This will minimize scale-factor error due to thermal gradients.
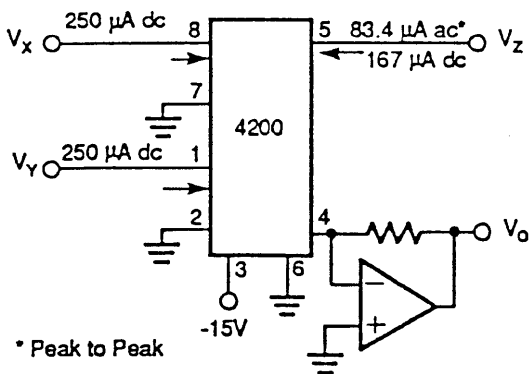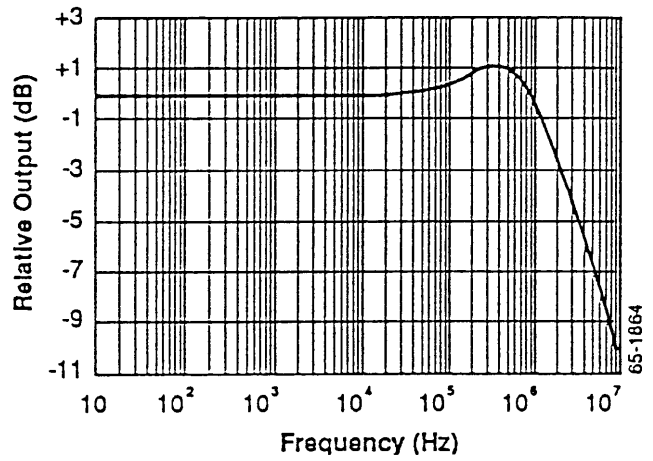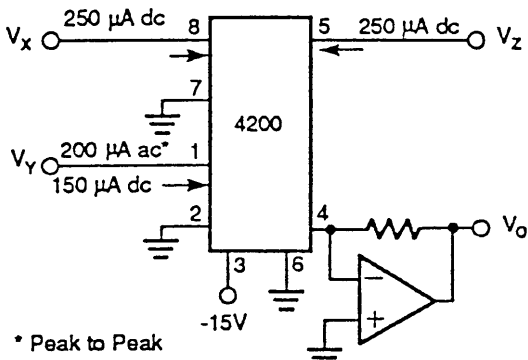
RC4200

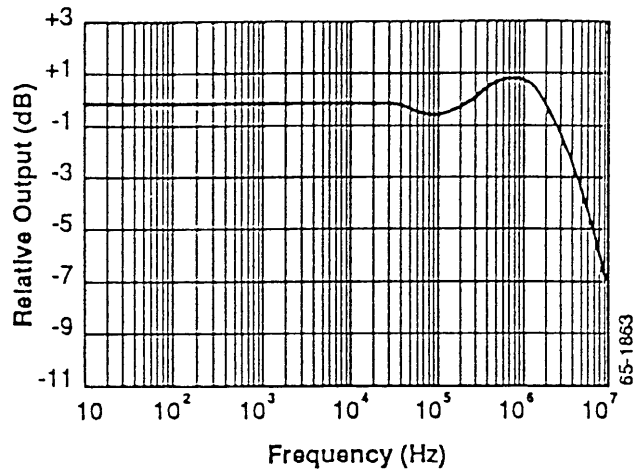**Raytheon Semiconductor**

For More Information, call 1-800-722-7074.
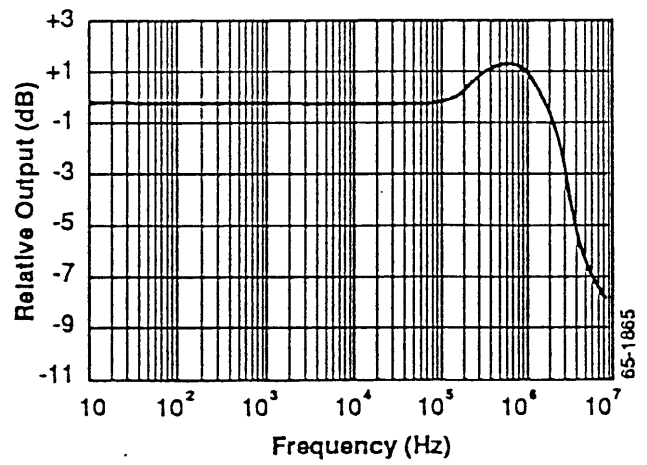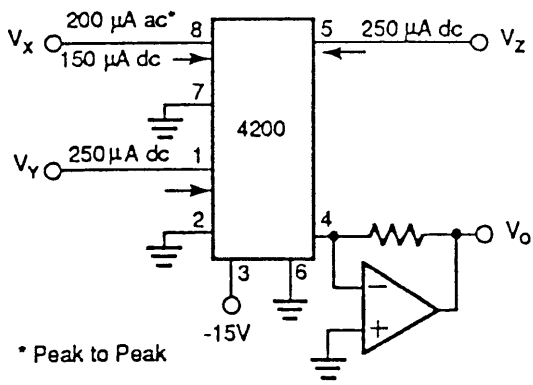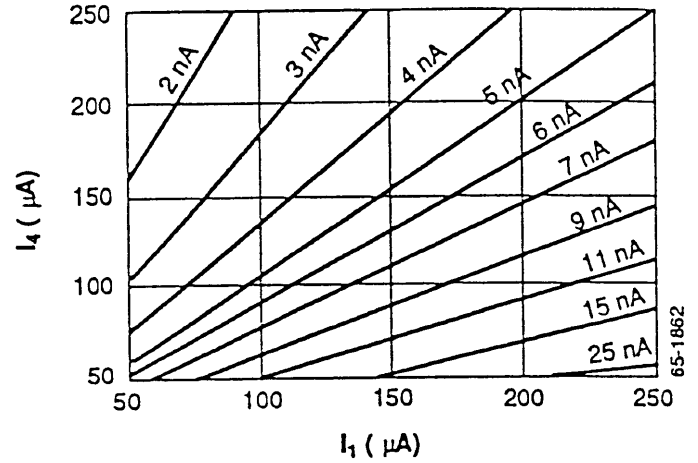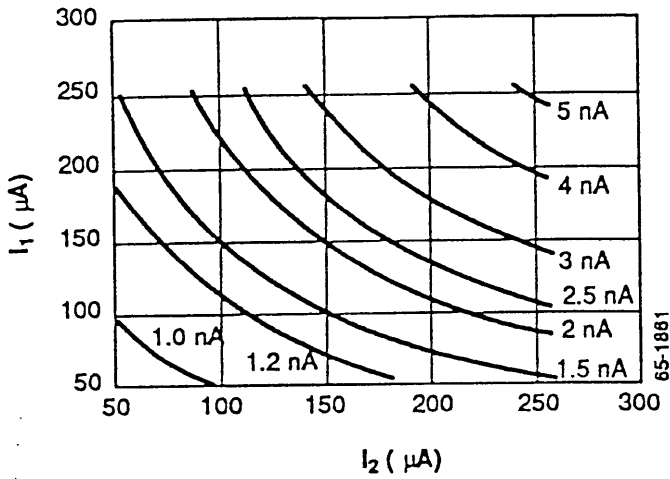
Figure 19

Figure 20a. Output Noise Current (I₃) vs. Input Currents (I₁, I₂)
for I₄ = 250 µA

Figure 20a. Output Noise Current (I₃) vs. Input Currents (I₄, I₁)
for I₂ = 250 µA
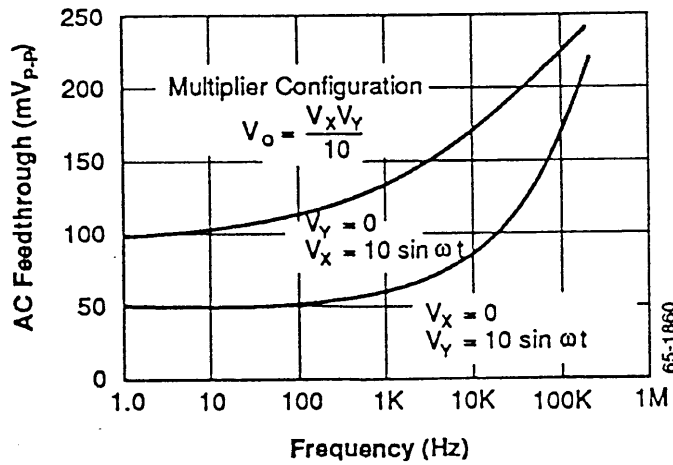
Figure 21. AC Feedthrough vs. Frequency

RC4200

Schematic Diagram
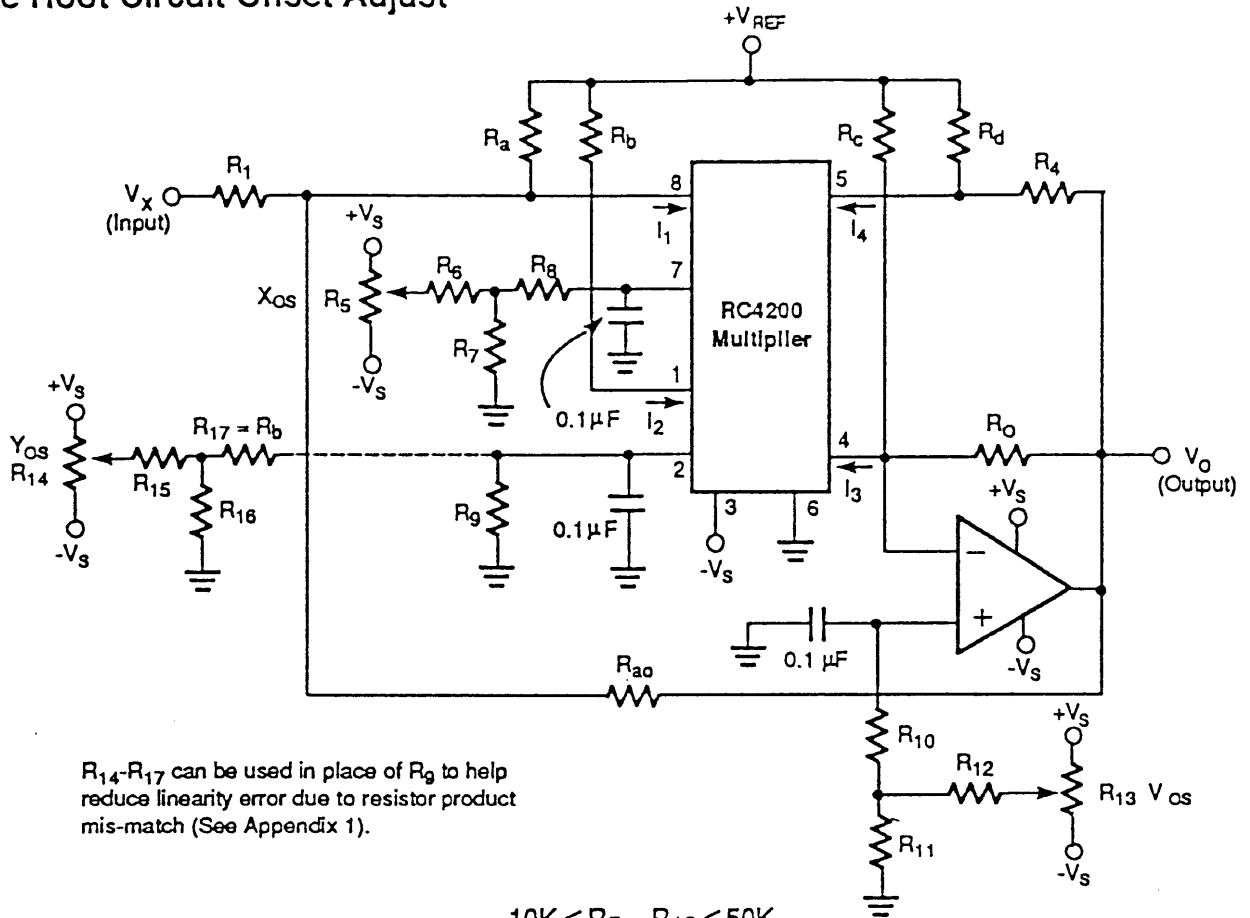
Raytheon Semiconductor
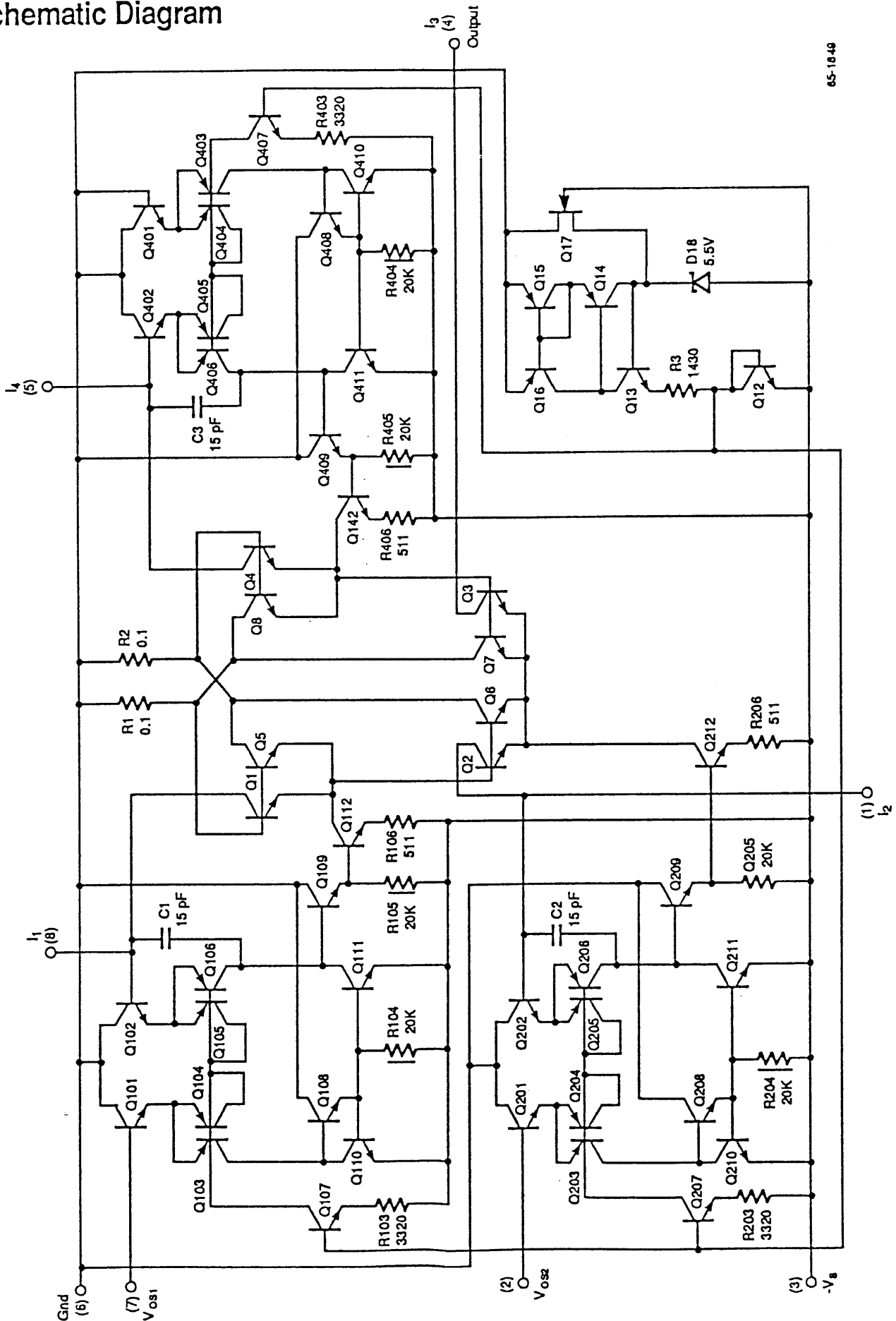
Square Root Circuit Offset Adjust

$$10K \leq R_5 = R_{13} \leq 50K$$

$$R_7 = 100\Omega$$

$$R_6 = R_7 \; \frac{V_S}{0.05}$$

$$R_8 = R_1| \; |R_a| \; |R_{ao}$$

$$R_9 = R_b$$

$$R_{10} = R_o| \; |R_c$$

$$R_{11} = 100\Omega$$

$$R_{12} = R_{11} \; \frac{V_S}{0.1}$$

R₁₄-R₁₇ can be used in place of R₉ to help reduce linearity error due to resistor product mis-match (See Appendix 1).
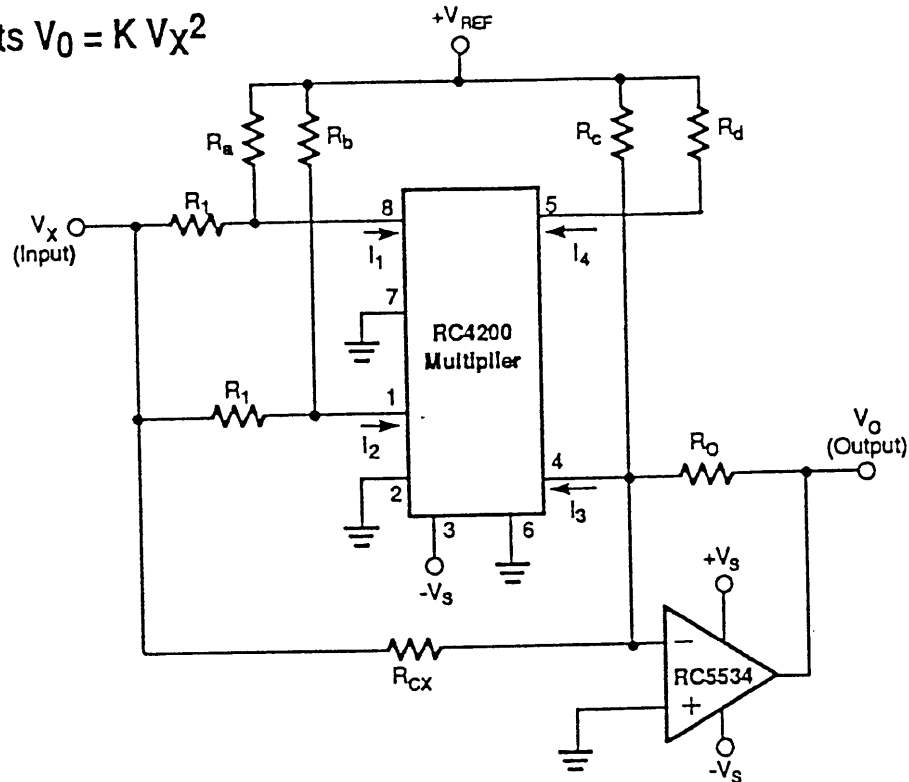
65-1876

Procedure

1. Set both trimmer pots to 0V on the wiper.

2. Put in a full scale (0 to V_X(max.) squarewave on V_X input. Adjust $X_{OS}(R_5)$ for proper peak-to-peak amplitude on V_0 output. (Scaling adjust)

3. Connect V_X input to ground. Adjust $V_{OS}(R_{13})$ for 0V on V_0 output.

Figure 10

Squaring Circuits $V_0 = K V_X^2$

65-1875

$$\frac{V_X^2}{R_1^2} + \frac{2V_X V_{REF}}{R_1 R_a} + \frac{V_{REF}^2}{R_a^2} = \frac{V_0 V_{REF}}{R_0 R_d} + \frac{V_{REF}^2}{R_c R_d} + \frac{V_X V_{REF}}{R_c R_d}$$

If $R_a^2 = R_c R_d$ and $R_1 R_a = 2R_{cx} R_d$

then $\dfrac{V_0 V_{REF}}{R_0 R_d} = \dfrac{V_X^2}{R_1^2}$ or $V_0 = KV_X^2$ where $K = \dfrac{R_0 R_d}{V_{REF} R_1^2}$

$V_X(\text{min.}) \leq V_X \leq V_X(\text{max.})$ $\Delta V_X = V_X(\text{max.}) - V_X(\text{min.})$

$K = \dfrac{V_0}{V_X^2}$ (Design Requirement)

$R_1 = \dfrac{\Delta V_X}{200\mu A}$

$R_a = \dfrac{\Delta V_X V_{REF}}{250\mu A \, \Delta V_X - 200\mu A \, V_X(\text{max.})}$

$R_d = \dfrac{V_{REF}}{250\mu A}$

$R_c = \dfrac{R_a^2}{R_d}$

$R_{cx} = \dfrac{R_1 R_a}{2R_d}$

$R_0 = \dfrac{\Delta V_X^2 K}{160\mu A}$

Figure 11

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For More Information, call 1-800-7

Appendix 1 — System Errors

There are four types of accuracy errors which affect overall system performance. They are:

1. Nonlinearity — Incremental deviation from absolute accuracy. [1]
2. Scaling Error — Linear deviation from absolute accuracy.
3. Output Offset — Constant deviation from absolute accuracy.
4. Feedthrough[2] — Cross-product errors caused by input offsets and external circuit limitations.

The nonlinearity error in the transfer function of the 4200 is ±0.1% max. (±0.03% max. for 4200A).

$$i.e., I_3 = \frac{I_1 I_2}{I_4} \pm 0.1\% \text{ F.S.} [4]$$

The other system errors are caused by voltage offsets on the inputs of the 4200 and can be as high as ±3.0% (±2.0% for 4200A).

$$i.e., V_0 = \frac{V_X V_Y}{V_Z} \frac{R_0 R_4}{R_1 R_2} \pm 3.0\% \text{ F.S.} [3][4]$$
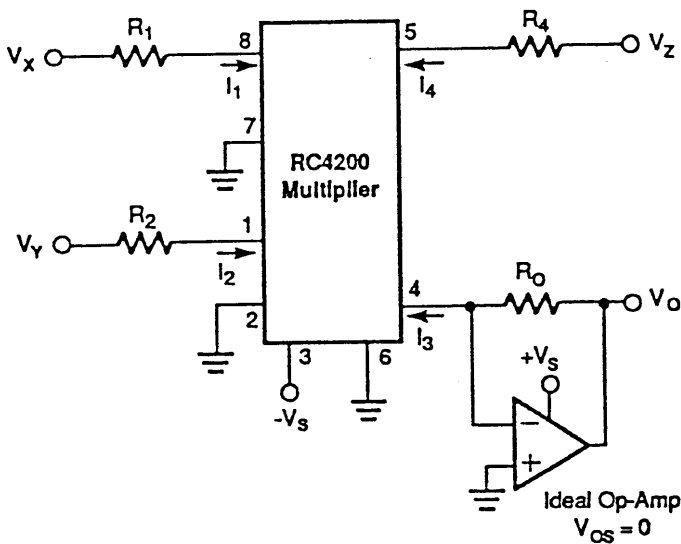

Figure 13

65-1871

Notes:
1. The input circuits tend to become unstable at $I_1, I_2, I_4 < 50\ \mu A$ and linearity decreases when $I_1, I_2, I_4 > 250\ \mu A$ (e.g., @ $I_1 = I_2 = 500\ \mu A$ nonlinearity error ≈ 0.5%).
2. This section will not deal with feedthrough which is proportional to frequency of operation and caused by stray capacitance and/or bandwidth limitations. (refer to Figure 21.)
3. Not including resistor tolerance or output offset on the op amp.
4. For $50\ \mu A \leq I_1, I_2, I_4 \leq 250\ \mu A$.

Errors Caused by Input Offsets

$$V_0 = \frac{R_0 R_4}{R_1 R_2} \left[\frac{V_X V_Y}{V_Z} \pm \frac{1}{V_Z} \underbrace{V_Y V_{OSX}}_{} \pm \underbrace{V_X V_{OSY}}_{} \pm \underbrace{V_0 V_{OSZ}}_{} \pm \underbrace{V_{OSX} V_{OSY}}_{} \right]$$

V_Y Feedthrough
V_X Feedthrough
Scaling Error
Output Offset Error

System errors can be greatly reduced by externally trimming the input offset voltages of the 4200. (±3.0% F.S. for 4200 and ±0.1% for 4200A.)

65-1870
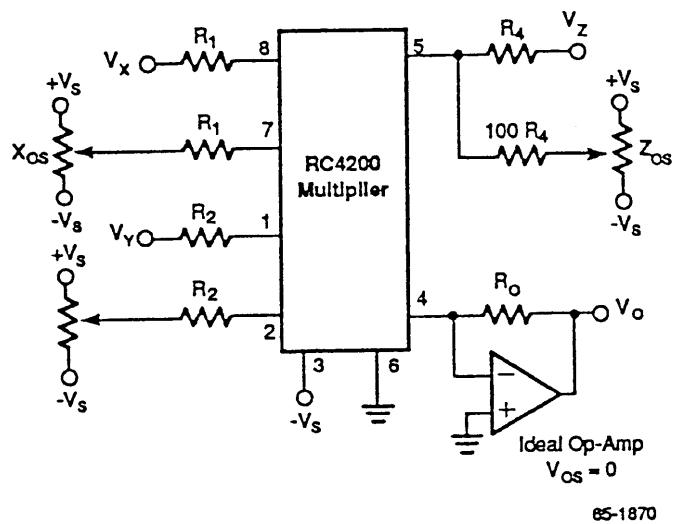
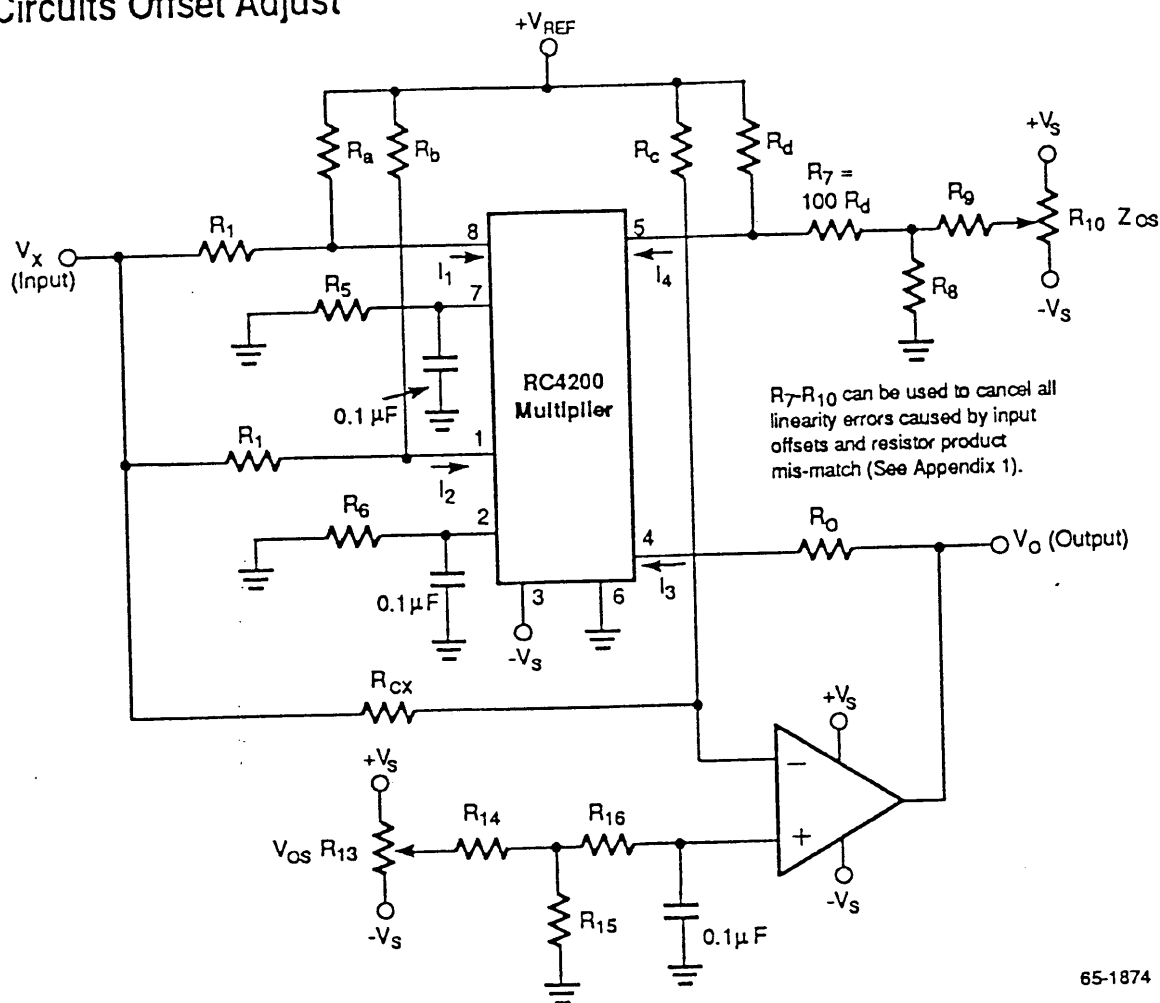if; $X_{OS} = X_{OSX}$, $Y_{OS} = Y_{OSY}$, $Z_{OS} = -V_{OSZ}$,

then $V_0 \frac{V_X V_Y}{V_Z} \frac{R_0 R_4}{R_1 R_2}$ ±0.3% F.S.[3]

Figure 14. 4200 with Input Offset Adjustment

Raytheon Semiconductor For More Information, call 1-800-722-707⁴

Squaring Circuits Offset Adjust

R_7-R_{10} can be used to cancel all linearity errors caused by input offsets and resistor product mis-match (See Appendix 1).

65-1874

$$10K \leq R_{10}, R_{11} \leq 50K$$

$$R_8, R_{15} = 100\Omega$$

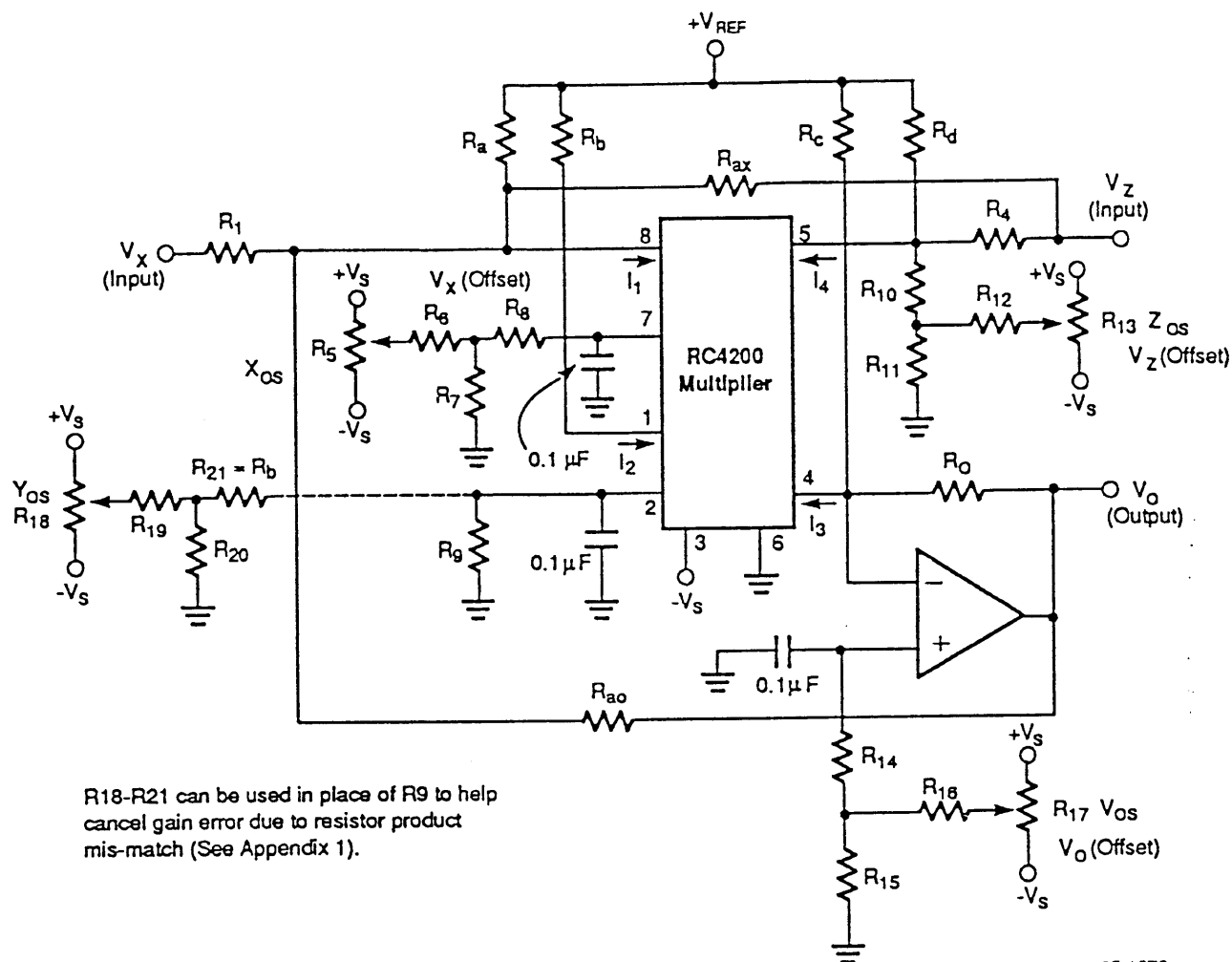$$R_9, R_{14} = 100\Omega \frac{V_S}{0.1}$$

$$R_5, R_6 = R_1 | |R_a$$

$$R_{16} = R_0 | |R_c| |R_{cx}$$

Procedure
1. Set both trimmer pots to 0V on the wiper.
2. Put in a full scale ($\pm V_X$) squarewave on V_X input. Adjust $Z_{os}(R_{10})$ for uniform output.
3. Connect V_X input to ground. Adjust $V_{OS}(R_{11})$ for 0V on V_O output.

Figure 12

Divider Circuit with Offset Adjustment

65-1878

General

$10K \leq R_5 = R_{13} = R_{17} \leq 50K$

$R_7 + R_8 \approx R_1 | \; |R_a| \; |R_{az}| \; |R_{ao}$

$R_6 \approx R_7 (V_S/0.05)$

$R_9 = R_b$

$R_{10} \approx 100 \times R_4$

$R_{11} = 20K$

$R_{12} = 100K$

$R_{14} + R_{15} \approx R_o | R_c$

$R_{16} \approx R_{15} (V_S/0.10)$

Example: Two-Quad Divider

$V_0 = K (V_X/V_Z)$, $K = k$, $V_{REF} = +V_S = +15V$

$-10 \leq V_X \leq +10$, therefore $\Delta V_X = 20$

$0 \leq V_Z \leq +10$, therefore $\Delta V_Z = 10$

$-10 \leq V_0 \leq +10$, therefore $\Delta V_0 = 20$

$R_0 = 26.7K$	$R_1 = 333K$
$R_b = 60K$	$R_5, R_{13}, R_{17} = 10K$
$R_4 = 50K$	$R_7, R_{15} = 1K$
$R_c = 37.5K$	$R_8, R_{11} = 20K$
$R_d = 300K$	$R_6, R_9, R_{16} = 300K$
$R_a = 187.5K$	$R_{10} = 4.7M$
$R_{az} = 31.25K$	$R_{12} = 100K$
$R_{ao} = 133K$	

R18-R21 can be used in place of R9 to help cancel gain error due to resistor product mis-match (See Appendix 1).
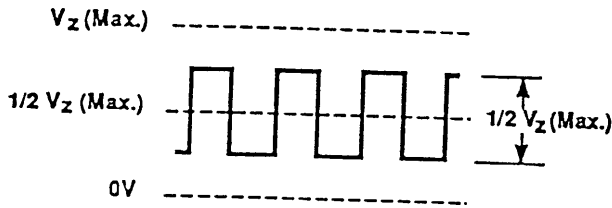
Figure 8

RC4200

Divider Circuit Offset Adjustment Procedure

1. Set each trimmer pot to 0V on the wiper.

2. Connect V_X (input) to ground. Put a DC voltage of approximately $1/2$ V_Z(max.) DC on the V_Z (input) with an AC (squarewave is easiest) voltage of $1/2$ V_Z (max.) peak-to-peak superimposed on it. Adjust X_{OS} (R_5) for zero feedthrough. (No AC at V_0)

3. Connect V_X (input) to V_Z (input) and put in the $1/2 V_Z$(max.) DC with an AC of approximately 20 mV less than V_Z(max.).

   Adjust Z_{OS}(R_{13}) for zero feedthrough.

$\approx$ 10 mV

65-1868
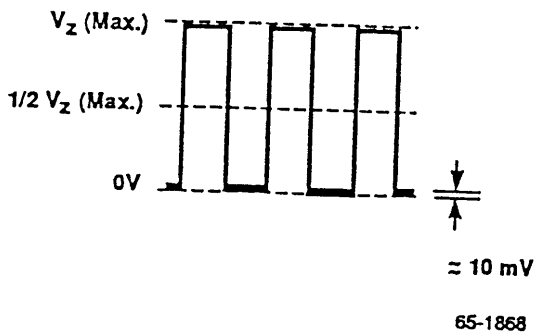
4. Return V_X (input) to ground and connect V_Z(max.) DC on V_Z(input). Adjust output V_{OS}(R_{17}) for $V_0 = 0V_0$

5. Connect V_X (input) to V_Z (input) and put in V_Z (max.) DC. (The output will equal K.) Decrease the input slowly until the output ($V_0 = K$) deviates beyond the desired accuracy. Adjust Z_{OS} to bring it back into tolerance and return to Step 4. Continue Steps 4 and 5 until V_Z reduces to the lowest value desired.

Note: As the input to V_X and V_Z gets closer to zero (an illegal state) the system noise will predominate so much that an integrating voltmeter will be very helpful.

Square Root Circuit $V_0 = N \sqrt{V_X}$
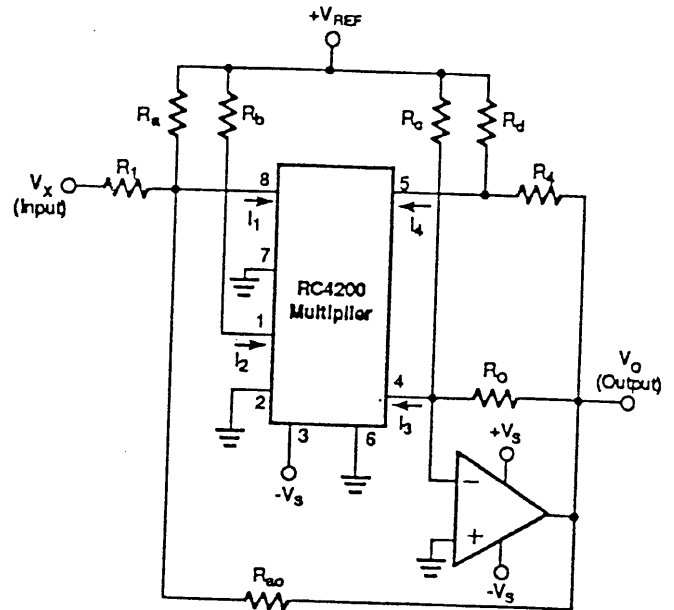
65-1877

$$\frac{V_X V_{REF}}{R_1 R_b} + \frac{V_{REF}^2}{R_a R_b} + \frac{V_0 V_{REF}}{R_{ao} R_b} - \frac{V_0^2}{R_0 R_4} + \frac{V_0 V_{REF}}{R_c R_4} + \frac{V_0 V_{REF}}{R_0 R_d} + \frac{V_{REF}^2}{R_c R_d}$$

If $R_a R_b = R_c R_d$ and $R_{ao} R_b R_0 R_d + R_{ao} R_b R_c R_4 = R_c R_d R_0 R_4$

Then $\dfrac{V_0^2}{R_0 R_4} = \dfrac{V_X V_{REF}}{R_1 R_b}$ or $V_0^2 = V_X K$ where $K = \dfrac{V_{REF} R_0 R_4}{R_1 R_b}$

and $V_0 = N\sqrt{V_X}$ where $N = \sqrt{K}$

$0 \leq V_X \leq V_X$(max.) and V_0(max.) $= N\sqrt{V_X}$(max)

$$N = \frac{V_0}{\sqrt{V_X}} \text{(Design Requirement)}$$

$$R_1 = \frac{V_0(\text{max})^2}{74\mu A \, N^2}$$

$$R_a = R_d = \frac{V_{REF}}{50\mu A}$$

$$R_b = R_c = \frac{V_{REF}}{150\mu A}$$

$$R_4 = \frac{V_0(\text{max})}{50\mu A}$$

$$R_{ao} = \frac{V_0(\text{max})}{125\mu A}$$

$$R_0 = \frac{V_0(\text{max})}{225\mu A}$$

Figure 9

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For More Information, call 1-800-722-7074.

8 Lead Plastic Dual Inline Package (PDIP) – .300" Body Width

Symbol	Inches Min.	Inches Max.	Millimeters Min.	Millimeters Max.	Notes
A	—	.210	—	5.33	
A1	.015	—	.38	—	
A2	.115	.195	2.93	4.95	
B	.014	.022	.36	.56	
B1	.045	.070	1.14	1.78	
C	.008	.015	.20	.38	4
D	.348	.430	8.84	10.92	2
D1	.005	—	.13	—	
E	.300	.325	7.62	8.26	
E1	.240	.280	6.10	7.11	2
e	.100 BSC		2.54 BSC		
eB	—	.430	—	10.92	
L	.115	.160	2.92	4.06	
N	8°		8°		5

Notes:

1. Dimensioning and tolerancing per ANSI Y14.5M-1982.
2. "D" and "E1" do not include mold flashing. Mold flash or protrusions shall not exceed .010 inch (0.25mm).
3. Terminal numbers are for reference only.
4. "C" dimension does not include solder finish thickness.
5. Symbol "N" is the maximum number of terminals.