The LM3900: A New Current-Differencing Quad of ± Input Amplifiers

PREFACE

With all the existing literature on "how to apply op amps" why should another application note be produced on this subject? There are two answers to this question; 1) the LM3900 operates in quite an unusual manner (compared to a conventional op amp) and therefore needs some explanation to familiarize a new user with this product, and 2) the standard op amp applications assume a split power supply (±15 V_{DC}) is available and our emphasis here is directed toward circuits for lower cost single power supply control systems. Some of these circuits are simply "re-biased" versions of conventional handbook circuits but many are new approaches which are made possible by some of the unique features of the LM3900.

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The LM3900: A New Current-Differencing Quad of ± Input Amplifiers

1.0 An Introduction to the New "Norton" Amplifier

The LM3900 represents a departure from conventional amplifier designs. Instead of using a standard transistor differential amplifier at the input, the non-inverting input function has been achieved by making use of a "current-mirror" to "mirror" the non-inverting input current about ground and then to extract this current from that which is entering the inverting input terminal. Whereas the conventional op amp differences input voltages, this amplifier differences input currents and therefore the name "Norton Amp" has been used to indicate this new type of operation. Many biasing advantages are realized when operating with only a single power supply voltage. The fact that currents can be passed between the input terminals allows some unusual applications. If external, large valued input resistors are used (to convert from input voltages to input currents) most of the standard op amp applications can be realized.

Many industrial electronic control systems are designed that operate off of only a single power supply voltage. The conventional integrated-circuit operational amplifier (IC op amp) is typically designed for split power supplies (± 15 V_{DC}) and suffers from a poor output voltage swing and a rather large minimum common-mode input voltage range (approximately $+ 2$ V_{DC}) when used in a single power supply application. In addition, some of the performance characteristics of these op amps could be sacrificed—especially in favor of reduced costs.

To meet the needs of the designers of low-cost, single-power-supply control systems, a new internally compensated amplifier has been designed that operates over a power supply voltage range of $+4$ V_{DC} to 36 V_{DC} with small changes in performance characteristics and provides an output peak-to-peak voltage swing that is only 1V less than the magnitude of the power supply voltage. Four of these amplifiers have been fabricated on a single chip and are provided in the standard 14-pin dual-in-line package.

The cost, application and performance advantages of this new quad amplifier will guarantee it a place in many single power supply electronic systems. Many of the "housekeeping" applications which are now handled by standard IC op amps can also be handled by this "Norton" amplifier operating off the existing ± 15 V_{DC} power supplies.

1.1 BASIC GAIN STAGE

The gain stage is basically a single common-emitter amplifier. By making use of current source loads, a large voltage gain has been achieved which is very constant over temperature changes. The output voltage has a large dynamic range, from essentially ground to one V_{BE} less than the power supply voltage. The output stage is biased class A for small signals but converts to class B to increase the load current which can be "absorbed" by the amplifier under large signal conditions. Power supply current drain is essentially independent of the power supply voltage and ripple on the supply line is also rejected. A very small input biasing current allows high impedance feedback elements to be used and even lower "effective" input biasing currents can be realized by using one of the amplifiers to supply essentially all of the bias currents for the other amplifiers by making use of the "matching" which exists between the 4 amplifiers which are on the same IC chip (see *Figure 84*).

The simplest inverting amplifier is the common-emitter stage. If a current source is used in place of a load resistor, a large open-loop gain can be obtained, even at low power-supply voltages. This basic stage (*Figure 1*) is used for the amplifier.
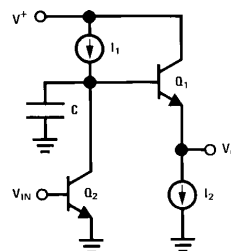
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FIGURE 1. Basic Gain Stage

All of the voltage gain is provided by the gain transistor, Q_2, and an output emitter-follower transistor, Q_1, serves to isolate the load impedance from the high impedance that exists at the collector of the gain transistor, Q_2. Closed-loop stability is guaranteed by an on-chip capacitor C = 3 pF, which provides the single dominant open-loop pole. The output emitter-follower is biased for class-A operation by the current source I_2.

This basic stage can provide an adequate open-loop voltage gain (70 dB) and has the desired large output voltage swing capability. A disadvantage of this circuit is that the DC input current, I_{IN}, is large; as it is essentially equal to the maximum output current, I_{OUT}, divided by β^2. For example, for an output current capability of 10 mA the input current would be at least 1 μA (assuming $\beta^2 = 10^4$). It would be desirable to further reduce this by adding an additional transistor to achieve an overall β^3 reduction. Unfortunately, if a transistor is added at the output (by making Q_1 a Darlington pair) the peak-to-peak output voltage swing would be somewhat reduced and if Q_2 were made a Darlington pair the DC input voltage level would be undesirably doubled.

To overcome these problems, a lateral PNP transistor has been added as shown in *Figure 2*. This connection neither reduces the output voltage swing nor raises the DC input voltage, but does provide the additional gain that was needed to reduce the input current.
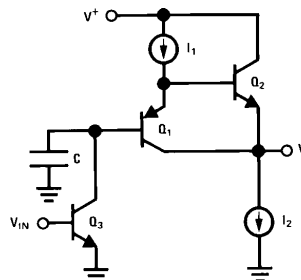
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FIGURE 2. Adding a PNP Transistor to the Basic Gain Stage

Notice that the collector of this PNP transistor, Q_1, is connected directly to the output terminal. This "bootstraps" the output impedance of Q_1 and therefore reduces the loading at the high-impedance collector of the gain transistor, Q_3.

In addition, the collector-base junction of the PNP transistor becomes forward biased under a large-signal negative output voltage swing condition. The design of this device has allowed Q_1 to convert to a vertical PNP transistor during this operating mode which causes the output to change from the class A bias to a class B output stage. This allows the amplifier to sink more current than that provided by the current source I_2, (1.3 mA) under large signal conditions.

1.2 OBTAINING A NON-INVERTING INPUT FUNCTION

The circuit of *Figure 2* has only the inverting input. A general purpose amplifier requires two input terminals to obtain both an inverting and a non-inverting input. In conventional op amp designs, an input differential amplifier provides these required inputs. The output voltage then depends upon the difference (or error) between the two input voltages. An input common-mode voltage range specification exists and, basically, input voltages are compared.

For circuit simplicity, and ease of application in single power supply systems, a non-inverting input can be provided by adding a standard IC "current-mirror" circuit directly across the inverting input terminal, as shown in *Figure 3*.

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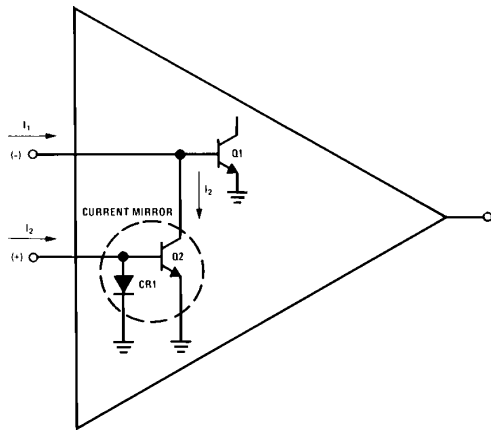
**FIGURE 3. Adding a Current Mirror to Achieve a
Non-inverting Input**

This operates in the current mode as now input currents are compared or differenced (this can be thought of as a Norton differential amplifier). There is essentially no input common-mode voltage range directly at the input terminals (as both inputs will bias at one diode drop above ground) but if the input voltages are converted to currents (by use of input resistors), there is then no limit to the common-mode input voltage range. This is especially useful in high-voltage comparator applications. By making use of the input resistors, to convert input voltages to input currents, all of the standard op amp applications can be realized. Many additional applications are easily achieved, especially when operating with only a single power supply voltage. This results from the built-in voltage biasing that exists at both inputs (each input biases at $+ V_{BE}$) and additional resistors are not required to

provide a suitable common-mode input DC biasing voltage level. Further, input summing can be performed at the relatively low impedance level of the input diode of the current-mirror circuit.

1.3 THE COMPLETE SINGLE-SUPPLY AMPLIFIER

The circuit schematic for a single amplifier stage is shown in *Figure 4a*. Due to the circuit simplicity, four of these amplifiers can be fabricated on a single chip. One common biasing circuit is used for all of the individual amplifiers.

A new symbol for this "Norton" amplifier is shown in *Figure 4b*. This is recommended to avoid using the standard op amp symbol as the basic operation is different. The current source symbol between the inputs implies this new current-mode of operation. In addition, it signifies that current is

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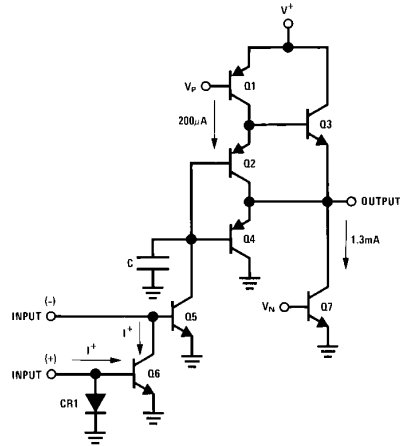
(a) Circuit Schematic

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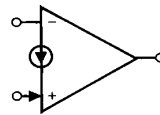
(b) New "NORTON" Amplifier Symbol

FIGURE 4. The Amplifier Stage

removed from the $(-)$ input terminal. Also, the current arrow on the $(+)$ input lead is used to indicate that this functions as a current input. The use of this symbol is helpful in understanding the operation of the application circuits and also in doing additional design work with the LM3900.

The bias reference for the PNP current source, V_p which biases Q_1, is designed to cause the upper current source (200 µA) to change with temperature to give first order compensation for the β variations of the NPN output transistor, Q_3. The bias reference for the NPN "pull-down" current sink, V_n, (which biases Q_7) is designed to stabilize this current (1.3 mA) to reduce the variation when the temperature is changed. This provides a more constant pull-down capability for the amplifier over the temperature range. The transistor, Q_4, provides the class B action which exists under large signal operating conditions.

The performance characteristics of each amplifier stage are summarized below:

Power-supply voltage range ....4 to 36 V_{DC} or
± 2 to ± 18 V_{DC}

Bias current drain per amplifier
stage ....1.3 mA_{DC}

Open loop:

Voltage gain (R_L = 10k) ....70 dB

Unity-gain frequency ....2.5 MHz

Phase margin......................................40°

Input resistance.................................1 $M\Omega$

Output resistance................................8 $k\Omega$

Output voltage swing ....$(V_{CC} - 1)$ V_{pp}

Input bias current..............................30 nA_{DC}

Slew rate 0.5V/μs

As the bias currents are all derived from diode forward voltage drops, there is only a small change in bias current magnitude as the power-supply voltage is varied. The open-loop gain changes only slightly over the complete power supply voltage range and is essentially independent of temperature changes. The open-loop frequency response is compared with the "741" op amp in *Figure 5*. The higher unity-gain crossover frequency is seen to provide an additional 10 dB of gain for all frequencies greater than 1 kHz.
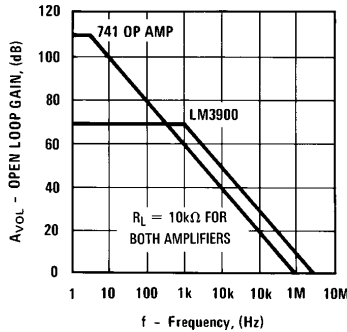
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FIGURE 5. Open-loop Gain Characteristics

The complete schematic diagram of the LM3900 is shown in *Figure 6*. The one resistor, R_5, establishes the power consumption of the circuit as it controls the conduction of transistor Q_{28}. The emitter current of Q_{28} is used to bias the NPN output class-A biasing current sources and the collector current of Q_{28} is the reference for the PNP current source of each amplifier.

The biasing circuit is initially "started" by Q_{20}, Q_{30} and CR_6. After start-up is achieved, Q_{30} goes OFF and the current flow through the reference diodes: CR_5, CR_7 and CR_8, is dependent only on $V_{BE}/(R_6 + R_7)$. This guarantees that the power supply current drain is essentially independent of the magnitude of the power supply voltage.

The input clamp for negative voltages is provided by the multi-emitter NPN transistor Q_{21}. One of the emitters of this transistor goes to each of the input terminals. The reference voltage for the base of Q_{21} is provided by R_6 and R_7 and is approximately $V_{BE}/2$.

2.0 Introduction to Applications of the LM3900

Like the standard IC op amp, the LM3900 has a wide range of applications. A new approach must be taken to design circuits with this "Norton" amplifier and the object of this note is to present a variety of useful circuits to indicate how conventional and unique new applications can be designed—especially when operating with only a single power supply voltage.

To understand the operation of the LM3900 we will compare it with the more familiar standard IC op amp. When operating on a single power supply voltage, the minimum input common-mode voltage range of a standard op amp limits the smallest value of voltage which can be applied to both inputs and still have the amplifier respond to a differential input signal. In addition, the output voltage will not swing completely from ground to the power supply voltage. The output voltage depends upon the difference between the input voltages and a bias current must be supplied to both inputs. A simplified diagram of a standard IC op amp operating from a single power supply is shown in *Figure 7*. The ($+$) and ($-$) inputs go only to current sources and therefore are free to be biased or operated at any voltage values which are within the input common-mode voltage range. The current sources at the input terminals, I_B^+ and I_B^-, represent the bias currents which must be supplied to both of the input transistors of the op amp (base currents). The output circuit is modeled as an active voltage source which depends upon the open-loop gain of the amplifier, A_v, and the difference which exists between the input voltages, $(V^+ - V^-)$.
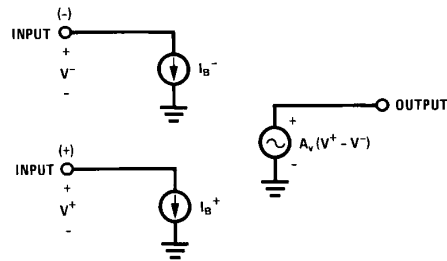
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FIGURE 7. An Equivalent Circuit of a Standard IC Op Amp

An equivalent circuit for the "Norton" amplifier is shown in *Figure 8*. The ($+$) and ($-$) inputs are both clamped by diodes to force them to be one-diode drop above ground—always! They are not free to move and the "input common-mode voltage range" directly at these input terminals is very small—a few hundred mV centered about 0.5 V_{DC}. This is
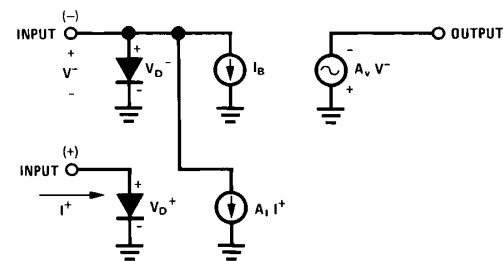
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FIGURE 8. An Equivalent Circuit of the "Norton" Amplifier
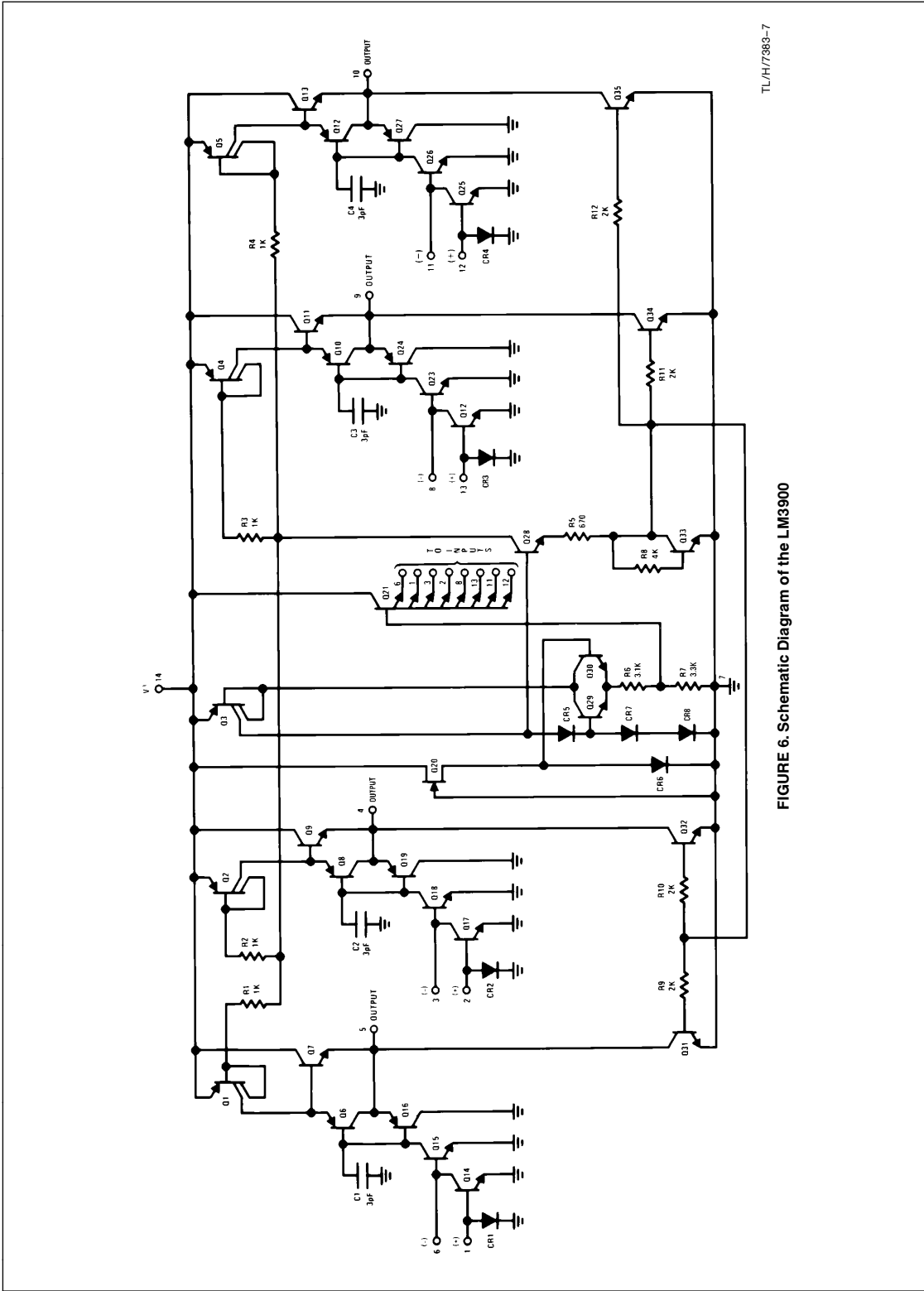
FIGURE 6. Schematic Diagram of the LM3900

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why external voltages must be first converted to currents (using resistors) before being applied to the inputs—and is the basis for the current-mode (or Norton) type of operation. With external input resistors—there is *no limit* to the "input common-mode voltage range". The diode shown across the (+) input actually exists as a diode in the circuit and the diode across the (−) input is used to model the base-emitter junction of the transistor which exists at this input.

Only the (−) input must be supplied with a DC biasing current, I_B. The (+) input couples only to the (−) input and then to extract from this (−) input terminal the same current (A_I, the mirror gain, is approximately equal to 1) which is entered (by the external circuitry) into the (+) input terminal. This operation is described as a "current-mirror" as the current entering the (+) input is "mirrored" or "reflected" about ground and is then extracted from the (−) input. There is a maximum or near saturation value of current which the "mirror" at the (+) input can handle. This is listed on the data sheet as "maximum mirror current" and ranges from approximately 6 mA at 25°C to 3.8 mA at 70°C.

This fact that the (+) input current modulates or affects the (−) input current causes this amplifier to pass currents between the input terminals and is the basis for many new application circuits—especially when operating with only a single power supply voltage.

The output is modeled as an active voltage source which also depends upon the open-loop voltage gain, A_v, but only the (−) input voltage, V^-, (not the differential input voltage). Finally, the output voltage of the LM3900 can swing from essentially ground (+90 mV) to within one V_{BE} of the power supply voltage.

As an example of the use of the equivalent circuit of the LM3900, the AC coupled inverting amplifier of *Figure 9a* will

be analyzed. *Figure 9b* shows the complete equivalent circuit which, for convenience, can be separated into a biasing equivalent circuit (*Figure 10*) and an AC equivalent circuit (*Figure 11*). From the biasing model of *Figure 10* we find the output quiescent voltage, V_O, is:

$$V_O = V_D{}^- + (I_B + I^+) R_2, \qquad (1)$$

and

$$I^+ = \frac{V^+ - V_D{}^+}{R_3} \qquad (2)$$

where

$$V_D{}^+ \cong V_D{}^- \cong 0.5 \ V_{DC}$$
$$I_B = \text{INPUT bias current (30 nA)}$$

and

$$V^+ = \text{Power supply voltage.}$$

If (2) is substituted into (1)

$$V_O = V_D{}^- + \left(I_B + \frac{V^+ - V_D{}^+}{R_3} \right) R_2 \qquad (3)$$

which is an exact expression for V_O.

As the second term usually dominates ($V_O \gg V_D{}^-$) and $I^+ \gg I_B$ and $V^+ \gg V_D{}^+$ we can simplify (3) to provide a more useful design relationship

$$V_O \cong \frac{R_2}{R_3} \ V^+. \qquad (4)$$

Using (4), if $R_3 = 2R_2$ we find

$$V_O \cong \frac{R_2}{2R_2} V^+ = \frac{V^+}{2}, \qquad (5)$$

which shows that the output is easily biased to one-half of the power supply voltage by using V^+ as a biasing reference at the (+) input.

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(a) A Typical Biased Amplifier
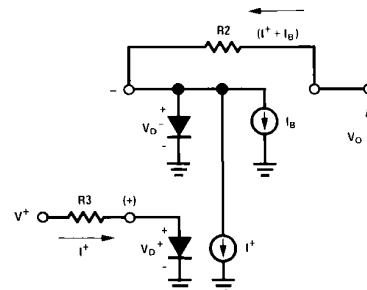
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FIGURE 10. Biasing Equivalent Circuit

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(b) Using the LM3900 Equivalent Circuit

FIGURE 9. Applying the LM3900 Equivalent Circuit
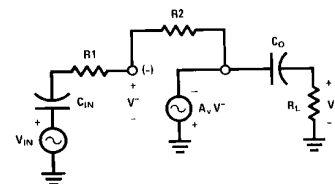
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FIGURE 11. AC Equivalent Circuit

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The AC equivalent circuit of *Figure 11* is the same as that which would result if a standard IC op amp were used with the (+) input grounded. The closed-loop voltage gain A_{VCL}, is given by:

$$A_{VCL} \equiv \frac{V_O}{V_{IN}} \cong - \frac{R_2}{R_1} \qquad (6)$$

if A_V (open-loop) $> \dfrac{R_2}{R_1}$.

The design procedure for an AC coupled inverting amplifier using the LM3900 is therefore to first select R_1, C_{IN}, R_2, and C_O as with a standard IC op amp and then to simply add $R_3 = 2R_2$ as a final biasing consideration. Other biasing techniques are presented in the following sections of this note. For the switching circuit applications, the biasing model of *Figure 10* is adequate to predict circuit operation.

Although the LM3900 has four independent amplifiers, the use of the label "¼LM3900" will be shortened to simply "LM3900" for the application drawings contained in this note.

3.0 Designing AC Amplifiers

The LM3900 readily lends itself to use as an AC amplifier because the output can be biased to any desired DC level within the range of the output voltage swing and the AC gain is independent of the biasing network. In addition, the single power supply requirement makes the LM3900 attractive for any low frequency gain application. For lowest noise performance, the (+) input should be grounded (*Figure 9a*) and the output will then bias at $+V_{BE}$. Although the LM3900 is not suitable as an ultra low noise tape pre-amp, it is useful in most other applications. The restriction to only shunt feedback causes a small input impedance. Transducers which can be loaded can operate with this low input impedance. The noise degradation which would result from the use of a large input resistor limits the usefulness where low noise and high input impedance are both required.

3.1 SINGLE POWER SUPPLY BIASING

The LM3900 can be biased in several different ways. The circuit in *Figure 12* is a standard inverting AC amplifier which has been biased from the same power supply which is used to operate the amplifier. (The design of this amplifier has been presented in the previous section). Notice that if AC ripple voltages are present on the V^+ power supply line they will couple to the output with a "gain" of ½. To eliminate this, one source of ripple filtered voltage can be provided and then used for many amplifiers. This is shown in the next section.

$$V_{ODC} = \frac{V^+}{2}$$

$$A_V \cong - \frac{R_2}{R_1}$$
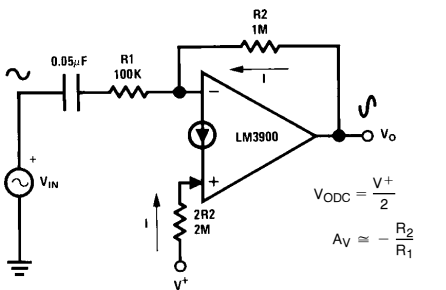
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FIGURE 12. Inverting AC Amplifier Using Single-Supply Biasing

3.2 A NON-INVERTING AMPLIFIER

The amplifier in *Figure 13* shows both a non-inverting AC amplifier and a second method for DC biasing. Once again the AC gain of the amplifier is set by the ratio of feedback resistor to input resistor. The small signal impedance of the diode at the (+) input should be added to the value of R_1 when calculating gain, as shown in *Figure 13*.

$$A_V = \frac{R3}{R1 + r_d}$$

$$r_d = \frac{0.026}{I_2} \ \Omega$$

$$V_{ODC} = \frac{V^+}{2}$$

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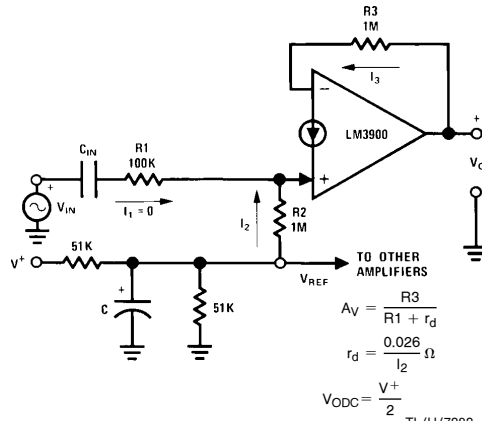
FIGURE 13. Non-inverting AC Amplifier Using Voltage Reference Biasing

By making $R_2 = R_3$, V_{ODC} will be equal to the reference voltage which is applied to the resistor R_2. The filtered $V^+/2$ reference shown can also be used for other amplifiers.

3.3 "N V_{BE}" BIASING

A third technique of output DC biasing is best described as the "N V_{BE}" method. This technique is shown in *Figure 14* and is most useful with inverting AC amplifier applications.

$$V_{BE} = 0.5 \ V_{DC}$$

$$V_{ODC} = V_{BE} \left(1 + \frac{R2}{R3} \right)$$

$$A_V \cong - \frac{R2}{R1}$$
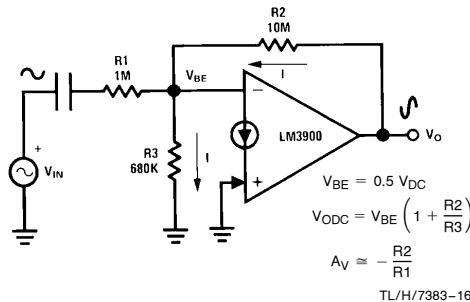
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FIGURE 14. Inverting AC Amplifier Using N V_{BE} Biasing

The input bias voltage (V_{BE}) at the inverting input establishes a current through resistor R_3 to ground. This current must come from the output of the amplifier. Therefore, V_O must rise to a level which will cause this current to flow through R_2. The bias voltage, V_O, may be calculated from the ratio of R_2 to R_3 as follows:

$$V_{ODC} = V_{BE} \left(1 + \frac{R_2}{R_3} \right)$$

When NV_{BE} biasing is employed, values for resistors R_1 and R_2 are first established and then resistor R_3 is added to provide the desired DC output voltage.

For a design example (*Figure 14*), a Z in = 1M and $A_V \cong 10$ are required.

Select R_1 = 1M.

Calculate $R_2 \cong A_V R_1 = 10M$.

To bias the output voltage at 7.5 V_{DC}, R_3 is found as:

$$R_3 = \frac{R_2}{\dfrac{V_O}{v_{BE}} - 1} = \frac{10M}{\dfrac{7.5}{0.5} - 1}$$

or

$$R_3 \cong 680 \, k\Omega.$$

3.4 BIASING USING A NEGATIVE SUPPLY

If a negative power supply is available, the circuit of *Figure 15* can be used. The DC biasing current, I, is established by the negative supply voltage via R_3 and provides a very stable output quiescent point for the amplifier.

$$V_{ODC} \cong -\frac{R2}{R3}V^-$$

$$A_V \cong -\frac{R2}{R1}$$

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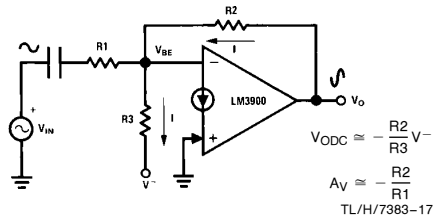
FIGURE 15. Negative Supply Biasing

3.5 OBTAINING HIGH INPUT IMPEDANCE AND HIGH GAIN

For the AC amplifiers which have been presented, a designer is able to obtain either high gain or high input impedance with very little difficulty. The application which requires both and still employs only one amplifier presents a new problem. This can be achieved by the use of a circuit similar to the one shown in *Figure 16*. When the A_V from the input to

$$V_A = \frac{V_O}{100}$$

$$A_V = -\frac{R4}{R5}$$

$$V_O = V_{REF}$$

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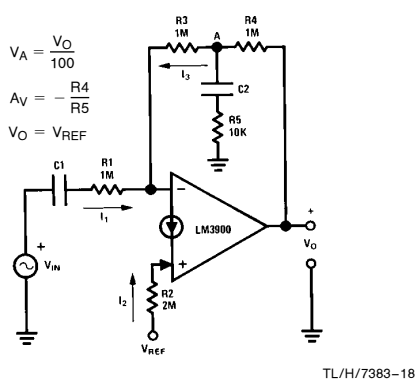
FIGURE 16. A High Z_{IN} High Gain Inverting AC Amplifier

point A is unity ($R_1 = R_3$), the A_V of the complete stage will be set by the voltage divider network composed of R_4, R_5, and C_2. As the value of R_5 is decreased, the A_V of the stage will approach the AC open loop limit of the amplifier. The insertion of capacitor C_2 allows the DC bias to be controlled by the series combination of R_3 and R_4 with no effect from R_5. Therefore, R_2 may be selected to obtain the desired output DC biasing level using any of the methods which have been discussed. The circuit in *Figure 16* has an input impedance of 1M and a gain of 100.

3.6 AN AMPLIFIER WITH A DC GAIN CONTROL

A DC gain control can be added to an amplifier as shown in *Figure 17*. The output of the amplifier is kept from being driven to saturation as the DC gain control is varied by providing a minimum biasing current via R_3. For maximum gain, CR_2 is OFF and both the current through R_2 and R_3 enter the (+) input and cause the output of the amplifier to bias at approximately 0.6 V^+. For minimum gain, CR_2 is ON and only the current through R_3 enters the (+) input to bias the output at approximately 0.3 V^+. The proper output bias for large output signal accommodation is provided for the maximum gain situation. The DC gain control input ranges from 0 V_{DC} for minimum gain to less than 10 V_{DC} for maximum gain.

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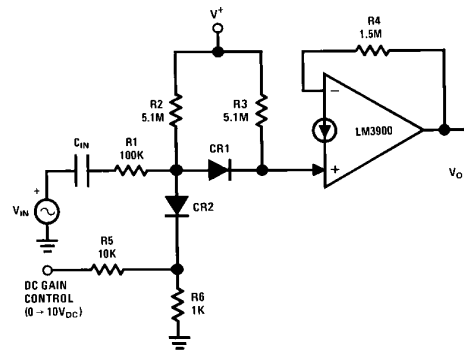
FIGURE 17. An Amplifier with a DC Gain Control

3.7 A LINE-RECEIVER AMPLIFIER

The line-receiver amplifier is shown in *Figure 18*. The use of both inputs cancels out common-mode signals. The line is terminated by R_{LINE} and the larger input impedance of the amplifier will not affect this matched loading.

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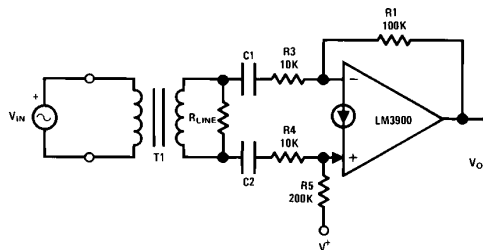
FIGURE 18. A Line-receiver Amplifier

4.0 Designing DC Amplifiers

The design of DC amplifiers using the LM3900 tends to be more difficult than the design of AC amplifiers. These difficulties occur when designing a DC amplifier which will operate from only a single power supply voltage and yet provide an output voltage which goes to zero volts DC and also will accept input voltages of zero volts DC. To accomplish this, the inputs must be biased into the linear region ($+V_{BE}$) with DC input signals of zero volts and the output must be modified if operation to actual ground (and not V_{SAT}) is required. Therefore, the problem becomes one of determining what type of network is necessary to provide an output voltage (V_O) equal to zero when the input voltage (V_{IN}) is equal to zero. (See also section 10.15, ''adding a Differential Input Stage'').

We will start with a careful evaluation of what actually takes place at the amplifier inputs. The mirror circuit demands that the current flowing into the positive input ($+$) be equaled by a current flowing into the negative input ($-$). The difference between the current demanded and the current provided by an external source must flow in the feedback circuit. The output voltage is then forced to seek the level required to cause this amount of current to flow. If, in the steady state condition $V_O = V_{IN} = 0$, the amplifier will operate in the desired manner. This condition can be established by the use of common-mode biasing at the inputs.

4.1 USING COMMON-MODE BIASING
FOR $V_{IN} = 0$ V_{DC}

Common-mode biasing is achieved by placing equal resistors between the amplifier input terminals and the supply voltage (V^+), as shown in *Figure 19*. When V_{IN} is set to 0 volts the circuit can be modeled as shown in *Figure 20*,

$$V_O \cong V_{IN}$$

$$A_V = \frac{R6}{R1}$$

$$R1 = R2$$
$$R3 = R4$$
$$R5 = R6$$

TL/H/7383–21
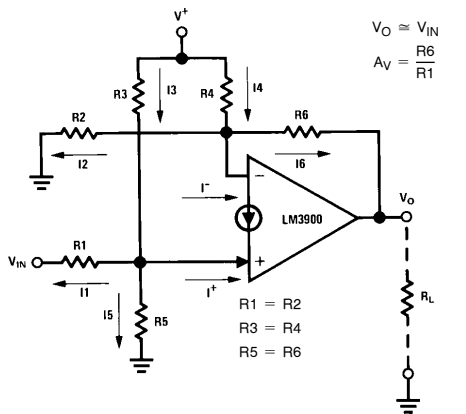
**FIGURE 19. A DC Amplifier Employing
Common-mode Biasing**

where:

$$R_{EQ_1} = R_1 \| R_5,$$
$$R_{EQ_2} = R_2 \| R_6,$$

and

$$R_3 = R_4.$$

Because the current mirror demands that the two current sources be equal, the current in the two equivalent resistors must be identical.

$$R3 = R4$$
$$I^+ = I^-$$
$$Req\ 1 = Req\ 2$$

TL/H/7383–22
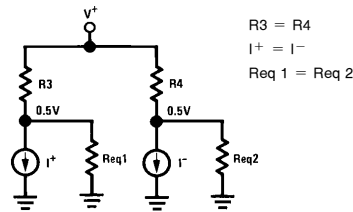
**FIGURE 20. An Ideal Circuit Model of a DC Amplifier
with Zero Input Voltage**

If this is true, both R_2 and R_6 must have a voltage drop of 0.5 volt across them, which forces V_O to go to $V_{O\ MIN}$ (V_{SAT}).

4.2 ADDING AN OUTPUT DIODE FOR $V_O = 0$ V_{DC}

For many applications a $V_{O\ MIN}$ Of 100 mV may not be acceptable. To overcome this problem a diode can be added between the output of the amplifier and the output terminal (*Figure 21*).

TL/H/7383–23
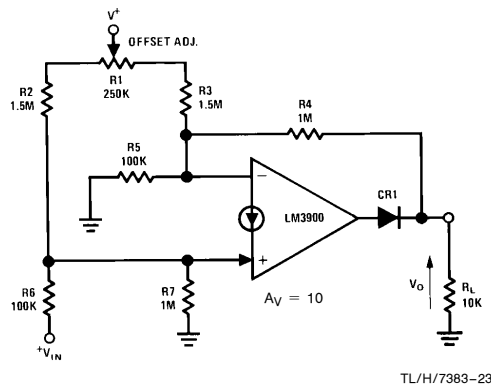
**FIGURE 21. A Non-inverting DC Amplifier with Zero
Volts Output for Zero Volts Input**

The function of the diode is to provide a DC level shift which will allow V_O to go to ground. With a load impedance (R_L) connected, V_O becomes a function of the voltage divider formed by the series connection of R_4 and R_L.

If $R_4 = 100\ R_L$, then $V_{O\ MIN} = \dfrac{0.5\ R_L}{101\ R_L}$,

or $V_{O\ MIN} \cong 5\ mV_{DC}$.

An offset voltage adjustment can be added as shown (R_1) to adjust V_O to $0V_{DC}$ with $V_{IN} = 0$ V_{DC}.

The voltage transfer functions for the circuit in *Figure 21*, both with and without the diode, are shown in *Figure 22*. While the diode greatly improves the operation around 0 volts, the voltage drop across the diode will reduce the peak output voltage swing of the state by approximately 0.5 volt.

When using a DC amplifier similar to the one in *Figure 21*, the load impedance should be large enough to avoid excessively loading the amplifier. The value of R_L may be significantly reduced by replacing the diode with an NPN transistor.
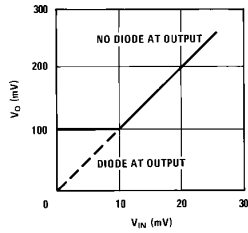
TL/H/7383–24

FIGURE 22. Voltage Transfer Function for a DC Amplifier with a Voltage Gain of 10

4.3 A DC COUPLED POWER AMPLIFIER ($I_L \leq 3$ AMPS)

The LM3900 may be used as a power amplifier by the addition of a Darlington pair at the output. The circuit shown in *Figure 23* can deliver in excess of 3 amps to the load when the transistors are properly mounted on heat sinks.
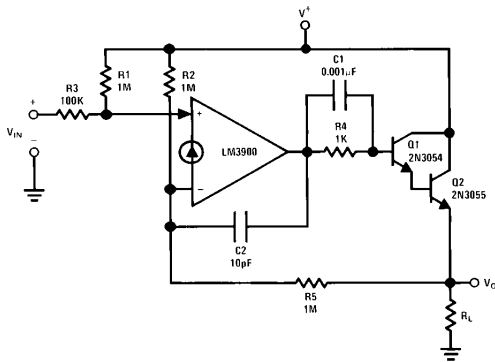
TL/H/7383–25

FIGURE 23. A DC Power Amplifier

4.4 GROUND REFERENCING A DIFFERENTIAL VOLTAGE

The circuit in *Figure 24* employs the LM3900 to ground reference a DC differential input voltage. Current I_1 is larger
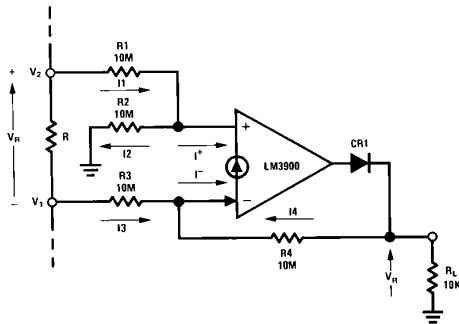
TL/H/7383–26

FIGURE 24. Ground Referencing a Differential Input DC Voltage

than current I_3 by a factor proportional to the differential voltage, V_R. The currents labeled on *Figure 24* are given by:

$$I_1 = \frac{V_1 + V_R - \phi}{R_1}$$

$$I_2 = \phi / R_2$$

$$I_3 = \frac{(V_1 - \phi)}{R_3}$$

and $\qquad I_4 = \frac{V_O - \phi}{R_4}$

where

$\phi \equiv V_{BE}$ at either input terminal of the LM3900.

Since the input current mirror demands that

$$I^- = I^+;$$

and $\qquad\qquad I^+ = I_1 - I_2$

and $\qquad\qquad I^- = I_3 + I_4$

Therefore $\qquad\quad I_4 = I_1 - I_2 - I_3.$

Substituting in from the above equation

$$\frac{V_O - \phi}{R_4} = \frac{(V_1 + V_R - \phi)}{R_1} - \frac{(\phi)}{R_2} - \frac{(V_1 - \phi)}{R_3}$$

and as $R_1 = R_2 = R_3 = R_4$

$$V_O = (V_1 + V_R - \phi) - (\phi) - V_1 + \phi + \phi$$

or

$$V_O = V_R.$$

The resistors are kept large to minimize loading. With the 10 MΩ resistors which are shown on the figure, an error exists at small values of V_1 due to the input bias current at the (−) input. For simplicity this has been neglected in the circuit description. Smaller R values reduce the percentage error or the bias current can be supplied by an additional amplifier (see Section 10.7.1).

For proper operation, the differential input voltage must be limited to be within the output dynamic voltage range of the amplifier and the input voltage V_2 must be greater than 1 volt. For example; if $V_2 = 1$ volt, the input voltage V_1 may vary over the range of 1 volt to −13 volts when operating from a 15 volt supply. Common-mode biasing may be added as shown in *Figure 25* to allow both V_1 and V_2 to be negative.
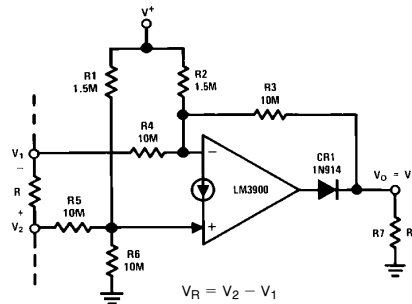
TL/H/7383–27

FIGURE 25. A Network to Invert and to Ground Reference a Negative DC Differential Input Voltage

4.5 A UNITY GAIN BUFFER AMPLIFIER

The buffer amplifier with a gain of one is the simplest DC application for the LM3900. The voltage applied to the input (*Figure 26*) will be reproduced at the output. However, the input voltage must be greater than one V_{BE} but less than the maximum output swing. Common-mode biasing can be added to extend V_{IN} to 0 V_{DC}, if desired.

13
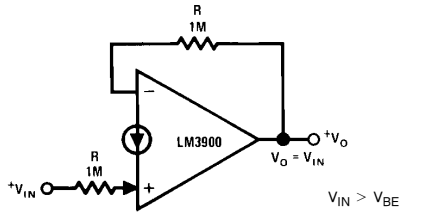
FIGURE 26. A Unity-gain DC Buffer Amplifier

TL/H/7383–28

5.0 Designing Voltage Regulators

Many voltage regulators can be designed which make use of the basic amplifier of the LM3900. The simplest is shown in *Figure 27a* where only a Zener diode and a resistor are added. The voltage at the (−) input (one $V_{BE} \cong 0.5\ V_{DC}$) appears across R and therefore a resistor value of 510Ω will cause approximately 1 mA of bias current to be drawn through the Zener. This biasing is used to reduce the noise output of the Zener as the 30 nA input current is too small for proper Zener biasing. To compensate for a positive temperature coefficient of the Zener, an additional resistor can be added, R_2, (*Figure 27b*) to introduce an arbitrary number, N, of "effective" V_{BE} drops into the expression for the output voltage. The negative temperature coefficient of these diodes will also be added to temperature compensate the DC output voltage. For a larger output current, an emitter follower (Q_1 of *Figure 27c*) can be added. This will multiply the 10 mA (max.) output current of the LM3900 by the β of the added transistor. For example, a $\beta = 30$ will provide a max. load current of 300 mA. This added transistor also reduces the output impedance. An output frequency compensation capacitor is generally not required but may be added, if desired, to reduce the output impedance at high frequencies.

The DC output voltage can be increased and still preserve the temperature compensation of *Figure 27b* by adding resistors R_A and R_B as shown in *Figure 27d*. This also can be accomplished without the added transistor, Q_1. The unregulated input voltage, which is applied to pin 14 of the LM3900 (and to the collector of Q_1, if used) must always exceed the regulated DC output voltage by approximately 1V, when the unit is not current boosted or approximately 2V when the NPN current boosting transistor is added.

5.1 REDUCING THE INPUT-OUTPUT VOLTAGE

The use of an external PNP transistor will reduce the required ($V_{IN} - V_{OUT}$) to a few tenths of a volt. This will depend on the saturation characteristics of the external transistor at the operating current level. The circuit, shown in *Figure 28*, uses the LM3900 to supply base drive to the PNP transistor. The resistors R_1 and R_2 are used to allow the output of the amplifier to turn OFF the PNP transistor. It is important that pin 14 of the LM3900 be tied to the $+V_{IN}$ line to allow this OFF control to properly operate. Larger voltages are permissible (if the base-emitter junction of Q_1 is prevented from entering a breakdown by a shunting diode, for example), but smaller voltages will not allow the output of the amplifier to raise enough to give the OFF control.

The resistor, R_3, is used to supply the required bias current for the amplifier and R_4 is again used to bias the Zener diode. Due to a larger gain, a compensation capacitor, C_O, is required. Temperature compensation could be added as was shown in *Figure 27b*.

$$V_O = V_Z + \phi$$

TL/H/7383–29

(a) Basic Current

$$V_O = V_Z + N\phi$$

TL/H/7383–30

(b) Temperature Compensating

$$V_O = V_Z + \phi$$
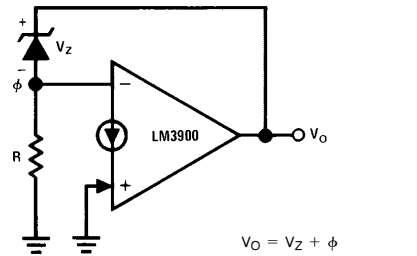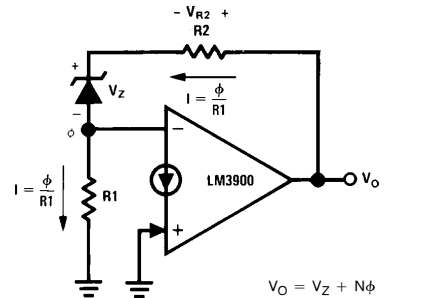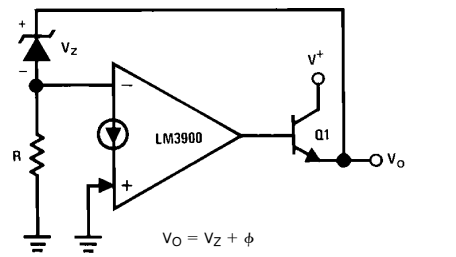
TL/H/7383–31

(c) Current Boosting
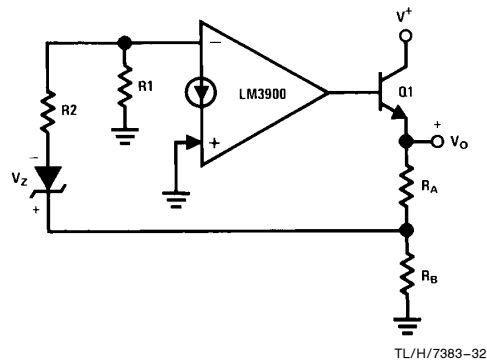
TL/H/7383–32

(d) Raising V_O Without Disturbing Temperature Compensation

FIGURE 27. Simple Voltage Regulators

14
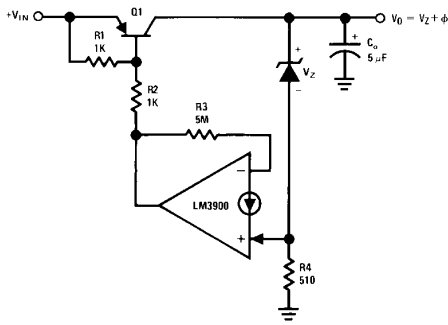
TL/H/7383–33

FIGURE 28. Reducing ($V_{IN} - V_{OUT}$)

5.2 PROVIDING HIGH INPUT VOLTAGE PROTECTION

One of the four amplifiers can be used to regulate the supply line for the complete package (pin 14), to provide protection against large input voltage conditions, and in addition, to supply current to an external load. This circuit is shown in *Figure 29*. The regulated output voltage is the sum of the Zener voltage, CR_2, and the V_{BE} of the inverting input terminal. Again, temperature compensation can be added as in *Figure 27b*. The second Zener, CR_1, is a low tolerance component which simply serves as a DC level shift to allow the output voltage of the amplifier to control the conduction of the external transistor, Q_1. This Zener voltage should be approximately one-half of the CR_2 voltage to position the DC Output voltage level of the amplifier approximately in the center of the dynamic range.
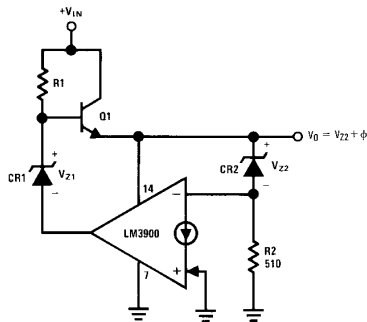
TL/H/7383–34

FIGURE 29. High V_{IN} Protection and Self-regulation

The base drive current for Q_1 is supplied via R_1. The maximum current through R_1 should be limited to 10 mA as

$$I_{MAX} = \frac{V_{IN\,(MAX)} - (V_O + V_{BE})}{R_1}$$

To increase the maximum allowed input voltage, reduce the output ripple, or to reduce the ($V_{IN} - V_{OUT}$) requirements of this circuit, the connection described in the next section is recommended.

5.3 HIGH INPUT VOLTAGE PROTECTION AND LOW ($V_{IN} - V_{OUT}$)

The circuit shown in *Figure 30* basically adds one additional transistor to the circuit of *Figure 29* to improve the performance. In this circuit both transistors (Q_1 and Q_2) absorb any high input voltages (and therefore need to be high voltage devices) without any increases in current (as with R_1 of *Figure 29*). The resistor R_1 (of *Figure 30*) provides a "start-up" current into the base of Q_2.

A new input connection is shown on this regulator (the type on *Figure 29* could also be used) to control the DC output voltage. The Zener is biased via R_4 (at approximately 1 mA). The resistors R_3 and R_6 provide gain (non-inverting) to allow establishing V_O at any desired voltage larger than V_Z. Temperature compensation of either sign ($\pm$TC) can be obtained by shunting a resistor from either the ($+$) input to ground (to add $+$ TC to V_O) or from the ($-$) input to ground (to add $-$TC to V_O). To understand this, notice that the resistor, R, from the ($+$) input to ground will add $-N\ V_{BE}$ to V_O where

$$N = 1 + \frac{R_3}{R},$$

and V_{BE} is the base emitter voltage of the transistor at the ($+$) input. This then also adds a positive temperature change at the output to provide the desired temperature correction.

The added transistor, Q_2, also increases the gain (which reduces the output impedance) and if a power device is used for Q_1 large load currents (amps) can be supplied. This regulator also supplies the power to the other three amplifiers of the LM3900.

5.4 REDUCING INPUT VOLTAGE DEPENDENCE AND ADDING SHORT-CIRCUIT PROTECTION

To reduce ripple feedthrough and input voltage dependence, diodes can be added as shown in *Figure 31* to dropout the start once start-up has been achieved. Short-circuit protection can also be added as shown in *Figure 32*.

The emitter resistor of Q_2 will limit the maximum current of Q_2 to ($V_O - 2\ V_{BE})/R_5$.
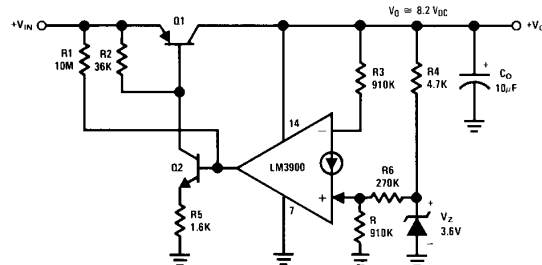
TL/H/7383–35

FIGURE 30. A High V_{IN} Protected, Low ($V_{IN} - V_{OUT}$) Regulator

15
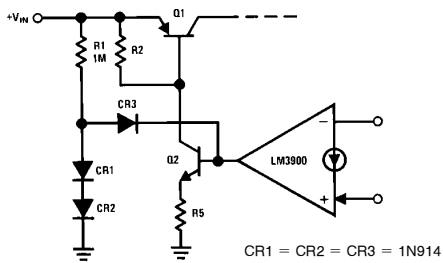
FIGURE 31. Reducing V_{IN} Dependence

CR1 = CR2 = CR3 = 1N914

TL/H/7383–36

$$I_{O\ MAX} \simeq \frac{V_{BE}}{R_{SC}}$$

TL/H/7383–37
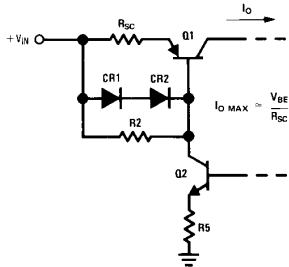
FIGURE 32. Adding Short-circuit Current Limiting

6.0 Designing RC Active Filters

Recent work in RC active filters has shown that the performance characteristics of multiple-amplifier filters are relatively insensitive to the tolerance of the RC components used. This makes the performance of these filters easier to control in production runs. In many cases where gain is needed in a system design it is now relatively easy to also get frequency selectivity.

The basis of active filters is a gain stage and therefore a multiple amplifier product is a valuable addition to this application area. When additional amplifiers are available, less component selection and trimming is needed as the performance of the filter is less disturbed by the tolerance and temperature drifts of the passive components.

The *passive components* do *control* the *performance* of the filter and for this reason carbon composition resistors are useful mainly for room temperature breadboarding or for final trimming of the more stable metal film or wire-wound resistors. Capacitors present more of a problem in range of values available, tolerance and stability (with temperature, frequency, voltage and time). For example, the disk ceramic type of capacitors are generally not suited to active filter applications due to their relatively poor performance.

The impedance level of the passive components can be scaled without (theoretically) affecting the filter characteristics. In an actual circuit; if the resistor values become too small (≤ 10 kΩ) an excessive loading may be placed on the output of the amplifier which will reduce gain or actually exceed either the output current or the package dissipation capabilities of the amplifier. This can easily be checked by calculating (or noticing) the impedance which is presented to the output terminal of the amplifier at the highest operating frequency. A second limit sets the upper range of impedance levels, this is due to the DC bias currents ($\simeq 30$ nA) and the input impedance of actual amplifiers. The solution to this problem is to reduce the impedance levels of the passive components (≤ 10 MΩ). In general, better perform-

ance is obtained with relatively low passive component impedance levels and in filters which do not demand high gain, high Q (Q $\geq$ 50) and high frequency ($f_o > 1$ kHz) simultaneously.

A measure of the effects of changes in the values of the passive components on the filter performance has been given by "sensitivity functions". These assume infinite amplifier gain and relate the percentage change in a parameter of the filter, such as center frequency (f_o), Q, or gain to a percentage change in a particular passive component. Sensitivity functions which are small are desirable (as 1 or $\frac{1}{2}$).

Negative signs simply mean an increase in the value of a passive component causes a decrease in that filter performance characteristic. As an example, if a bandpass filter listed the following sensitivity factor

$$S_{C_3}^{\omega_o} = -\frac{1}{2}.$$

This states that "if C_3 were to increase by 1%, the center frequency, ω_o, would decrease by 0.5%." Sensitivity functions are tabulated in the reference listed at the end of this section and will therefore not be included here.

A brief look at low pass, high pass and bandpass filters will indicate how the LM3900 can be applied in these areas. A recommended text (which provided these circuts) is, "Operational Amplifiers", Tobey, Graeme, and Huelsman, McGraw Hill, 1971.

6.1 BIASING THE AMPLIFIERS

Active filters can be easily operated off of a single power supply when using these multiple single supply amplifiers. The general technique is to use the ($+$) input to accomplish the biasing function. The power supply voltage, V^+, is used as the DC reference to bias the output voltage of each amplifier at approximately $V^+/2$. As shown in *Figure 33*, undesired AC components on the power supply line may have to

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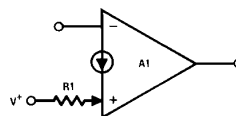
(a) Biasing From a "Noise-Free" Power Supply

TL/H/7383–39
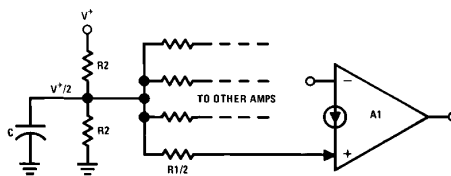
(b) Biasing From a "Noisy" Power Supply

FIGURE 33. Biasing Considerations

be removed (by a filter capacitor, *Figure 33b*) to keep the filter output free of this noise. One filtered DC reference can generally be used for all of the amplifiers as there is essentially no signal feedback to this bias point.

In the filter circuits presented here, all amplifiers will be biased at $V^+/2$ to allow the maximum AC voltage swing for any given DC power supply voltage. The inputs to these filters will also be assumed at a DC level of $V^+/2$ (for those which are direct coupled).

6.2 A HIGH PASS ACTIVE FILTER

A single amplifier high pass RC active filter is shown in *Figure 34*. This circuit is easily biased using the (+) input of the LM3900. The resistor, R_3, can be simply made equal to R_2 and a bias reference of $V^+/2$ will establish the output Q point at this value ($V^+/2$). The input is capacitively coupled (C_1) and there are therefore no further DC biasing problems.

$f_c = 1$ kHz
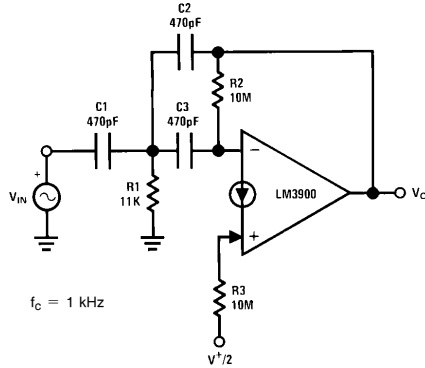
TL/H/7383–40

FIGURE 34. A High Pass Active Filter

The design procedure for this filter is to select the pass band gain, H_O, the Q and the corner frequency, f_c. A Q value of 1 gives only a slight peaking near the bandedge (<2 dB) and smaller Q values decrease this peaking. The slope of the skirt of this filter is 12 dB/octave (or 40 dB/decade). If the gain, H_O, is unity all capacitors have the same value. The design proceeds as:

Given: H_O, Q and $\omega_c = 2\pi f_c$

To find: R_1, R_2, C_1, C_2, and C_3

let $C_1 = C_3$ and choose a convenient starting value.

Then:

$$R_1 = \frac{1}{Q\,\omega_c C_1\,(2H_O + 1)} \qquad (1)$$

$$R_2 = \frac{Q}{\omega_c C_1}\,(2H_O + 1), \qquad (2)$$

and

$$C_2 = \frac{C_1}{H_O}. \qquad (3)$$

As a design example,

Require: $H_O = 1$,

$\qquad\quad Q = 10$,

and $\qquad f_c = 1$ kHz ($\omega_c = 6.28 \times 10^3$ rps).

Start by selecting $C_1 = 300$ pF and then from equation (1)

$$R_1 = \frac{1}{(10)\,(6.28 \times 10^3)\,(3 \times 10^{-10})\,(3)}$$

$\qquad R_1 = 17.7$ kΩ

and from equation (2)

$$R_2 = \frac{10 \qquad (3)}{(6.28 \times 10^3)\,(3 \times 10^{-10})}$$

$\qquad R_2 = 15.9$ MΩ

and from equation (3)

$$C_2 = \frac{C_1}{1} = C_1$$

Now we see that the value of R_2 is quite large; but the other components look acceptable. Here is where impedance scaling comes in. We can reduce R_2 to the more convenient value of 10 MΩ which is a factor of 1.59:1. Reducing R_1 by this same scaling factor gives:

$$R_{1NEW} = \frac{17.7 \times 10^3}{1.59} = 11.1 \text{ k}\Omega$$

and the capacitors are similarly reduced in impedance as:

$$\Big(C_1 = C_2 = C_3\Big)_{NEW} = (1.59)\,(300) \text{ pF}$$

$$C_{1NEW} = 477 \text{ pF}.$$

To complete the design, R_3 is made equal to R_2 (10 MΩ) and a V_{REF} of $V^+/2$ is used to bias the output for large signal accommodation.

Capacitor values should be adjusted to use standard valued components by using impedance scaling as a wider range of standard resistor values is generally available.

6.3 A LOW PASS ACTIVE FILTER

A single amplifier low pass filter is shown in *Figure 35*. The resistor, R_4, is used to set the output bias level and is selected after the other resistors have been established.

$f_c = 1$ kHz
GAIN = 1
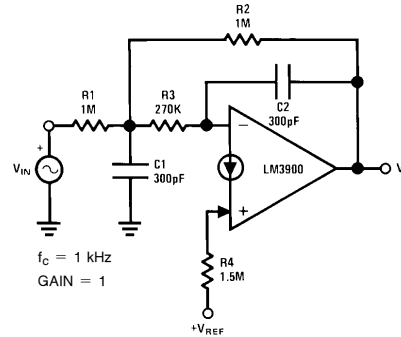
TL/H/7383–41

FIGURE 35. A Low Pass Active Filter

The design procedure is as follows:

Given: H_O, Q, and $\omega_c = 2\pi f_c$

To find: R_1, R_2, R_3, R_4, C_1, and C_2

Let C_1 be a convenient value,

then

$$C_2 = KC_1 \qquad (4)$$

where K is a constant which can be used to adjust component values. For example, with $K = 1$, $C_1 = C_2$. Larger values of K can be used to reduce R_2 and R_3 at the expense of a larger value for C_2.

$$R_1 = \frac{R_2}{H_O}, \qquad (5)$$

$$R_2 = \frac{1}{2Q\,\omega_c C_1}\left[1 \pm \sqrt{1 + \frac{4Q^2\,(H_O + 1)}{K}}\right] \qquad (6)$$

and

$$R_3 = \frac{1}{\omega_c^2 C_1^2 R_2(K)} \qquad (7)$$

As a design example:

Require: $H_O = 1$,

$\qquad\quad Q = 1$,

and $\qquad f_c = 1$ kHz ($\omega_c = 6.28 \times 10^3$ rps).

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Start by selecting $C_1 = 300$ pF and $K = 1$ so C_2 is also 300 pF (equation 4).

Now from equation (6)

$$R_2 = \frac{1}{2\,(1)\,(6.28 \times 10^3)\,(3 \times 10^{-10})}\left[1 \pm \sqrt{1 + 4\,(2)}\right]$$

$R_2 = 1.06$ MΩ

Then from equation (5)

$$R_1 = R_2 = 1.06 \text{ M}\Omega$$

and finally from equation (7)

$$R_3 = \frac{1}{(6.28 \times 10^3)^2\,(3 \times 10^{-10})^2\,(1.06 \times 10^6)\,(1)}$$

$R_3 = 266$ kΩ.

To select R_4, we assume the DC input level is 7 V_{DC} and the DC output of this filter is to also be 7 V_{DC}. This gives us the circuit of *Figure 36*. Notice that $H_O = 1$ gives us not only
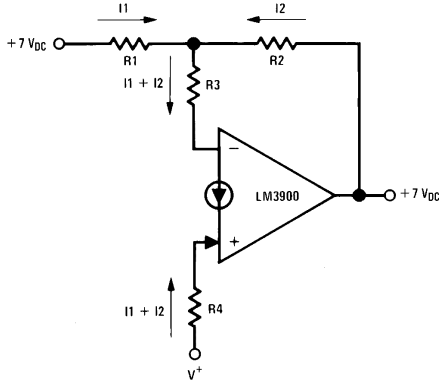
FIGURE 36. Biasing the Low Pass Filter

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equal resistor values (R_1 and R_2) but simplifies the DC bias calculation as $I_1 = I_2$ and we have a DC amplifier with a gain of -1 (so if the DC input voltage increases 1 V_{DC} the output voltage decreases 1 V_{DC}). The resistors R_1 and R_2 are in parallel so that the circuit simplifies to that shown in *Figure 37* where the actual resistance values have been added. The resistor R_4 is given by

$$R_4 = 2\left(\frac{R_1}{2} + R_3\right) + R_3$$

or, using values

$$R_4 = 2\left(\frac{1 \text{ M}\Omega}{2} + 266\text{k}\right) \cong 1.5 \text{ M}\Omega$$
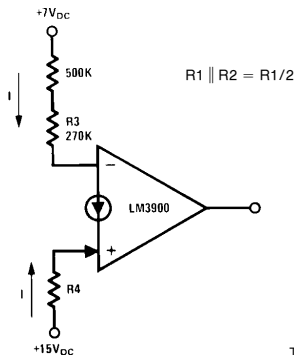
FIGURE 37. Biasing Equivalent Circuit

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6.4 A SINGLE-AMPLIFIER BANDPASS ACTIVE FILTER

The bandpass filter is perhaps the most interesting. For low frequencies, low gain and low Q (≤ 10) requirements, a single amplifier realization can be used. A one amplifier circuit is shown in *Figure 38* and the design procedure is as follows;

Given: H_O, Q and $\omega_o = 2\pi f$.

To find: R_1, R_2, R_3, R_4, C_1 and C_2.
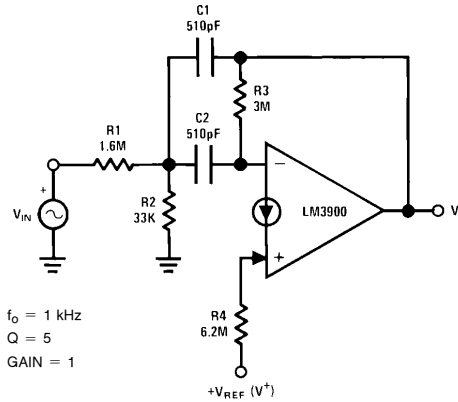
$f_o = 1$ kHz
$Q = 5$
GAIN $= 1$

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FIGURE 38. A One Op Amp Bandpass Filter

Let $C_1 = C_2$ and select a convenient starting value.
Then

$$R_1 = \frac{Q}{H_O \omega_o C_1} \tag{8}$$

$$R_2 = \frac{Q}{(2\,Q^2 - H_O)\,\omega_o\,C_1} \tag{9}$$

$$R_3 = \frac{2Q}{\omega_o C_1} \tag{10}$$

and

$$R_4 = 2R_3 \text{ (for } V_{REF} = V^+) \tag{11}$$

As a design example:

Require: $H_O = 1$

$\quad Q = 5$

$\quad f_o = 1$ kHz ($\omega_o = 6.28 \times 10^3$ rps).

Start by selecting

$\quad C_1 = C_2 = 510$ pF.

Then using equation (8)

$$R_1 = \frac{5}{(6.28 \times 10^3)\,(5.1 \times 10^{-10})}$$

$R_1 = 1.57$ MΩ,

and using equation (9)

$$R_2 = \frac{5}{[2(25) - 1]\,(6.28 \times 10^3)\,(5.1 \times 10^{-10})}$$

$R_2 = 32$ kΩ

from equation (10)

$$R_3 = \frac{2(5)}{(6.28 \times 10^3)\,(5.1 \times 10^{-10})}$$

$R_3 = 3.13$ MΩ,

and finally, for biasing, using equation (11)

$$R_4 = 6.2 \text{ M}\Omega.$$

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6.5 A TWO-AMPLIFIER BANDPASS ACTIVE FILTER

To allow higher Q (between 10 and 50) and higher gain, a two amplifier filter is required. This circuit, shown in *Figure 39*, uses only two capacitors. It is similar to the previous single amplifier bandpass circuit and the added amplifier supplies a controlled amount of positive feedback to improve the response characteristics. The resistors R_5 and R_8 are used to bias the output voltage of the amplifiers at $V^+/2$.

Again, R_5 is simply chosen as twice R_4 and R_8 must be selected after R_6 and R_7 have been assigned values. The design procedure is as follows:

Given: Q and f_o

To find: R_1 through R_7, and C_1 and C_2

Let: $C_1 = C_2$ and choose a convenient starting value and choose a value for K to reduce the spread of element values or to optimize sensitivity ($1 \leq K_{Typically} \leq 10$).

Then

$$R_1 = R_4 = R_6 = \frac{Q}{\omega_o C_1} \qquad (12)$$

$$R_2 = R_1 \frac{KQ}{(2Q - 1)} \qquad (13)$$

$$R_3 = \frac{R_1}{Q^2 - 1 - 2/K + 1/KQ} \qquad (14)$$

and

$$R_7 = KR_1 \qquad (15)$$
$$H_O = \sqrt{Q}\, K. \qquad (16)$$

As a design example:

Require: Q = 25 and f_o = 1 kHz.

Select: $C_1 = C_2 = 0.1\ \mu F$

and K = 3.

Then from equation (12)

$$R_1 = R_4 = R_6 = \frac{25}{(2\pi \times 10^3)(10^{-7})}$$

$$R_1 = 40\ k\Omega$$

and from equation (13)

$$R_2 = (40 \times 10^3) \frac{3(25)}{[2(25) - 1]}$$

$$R_2 = 61\ k\Omega$$

and from equation (14)

$$R_3 = \frac{40 \times 10^3}{(25)^2 - 1 - 2/3 + \dfrac{1}{3(25)}}$$

$$R_3 = 64\Omega$$

And R_7 is given by equation (15)

$$R_7 = 3\ (40\ k\Omega) = 120\ k\Omega,$$

and the gain is obtained from equation (16)

$$H_O = \sqrt{25}\ (3) = 15\ (23\ dB).$$

To properly bias the first amplifier

$$R_5 = 2R_4 = 80\ k\Omega$$

and the second amplifier is biased by R_8. Notice that the outputs of both amplifiers will be at $V^+/2$. Therefore R_6 and R_7 can be paralleled and

$$R_8 = 2(R_6 \parallel R_7)$$

or

$$R_8 = 2\left[\frac{(40)(120) \times 10^3}{160}\right] = 59\ k\Omega$$

These values, to the closest standard resistor values, have been added to *Figure 39*.

6.6 A THREE-AMPLIFIER BANDPASS ACTIVE FILTER

To reduce Q sensitivity to element variation even further or to provide higher Q (Q>50) a three amplifier bandpass filter can be used. This circuit, *Figure 40*, pre-dates most of the literature on RC active filters and has been used on analog computers. Due to the use of three amplifiers it often is considered too costly—especially for low Q applications. The multiple amplifiers of the LM3900 make this a very useful circuit. It has been called the "Bi-Quad" as it can produce a transfer function which is "Quad"–ratic in both numerator and denominator (to give the "Bi"). A newer real-

f_o = 1 kHz
Q = 25
GAIN = 15 (23 dB)
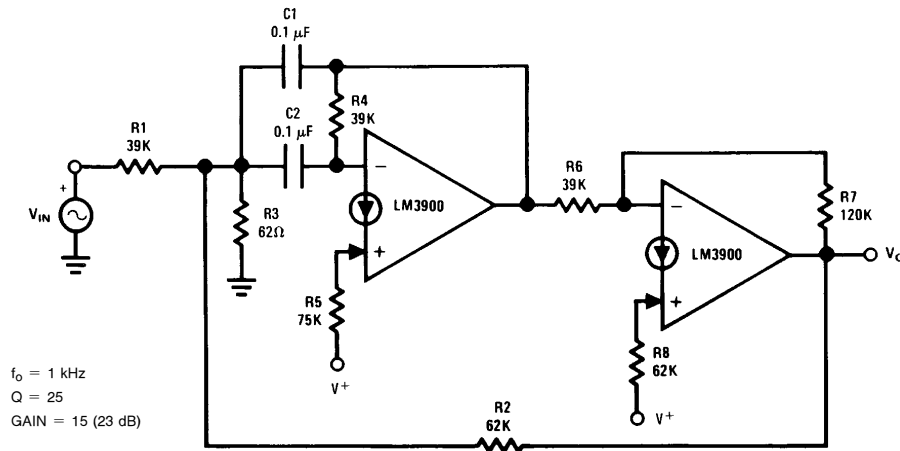
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FIGURE 39. A Two Op Amp Bandpass Filter
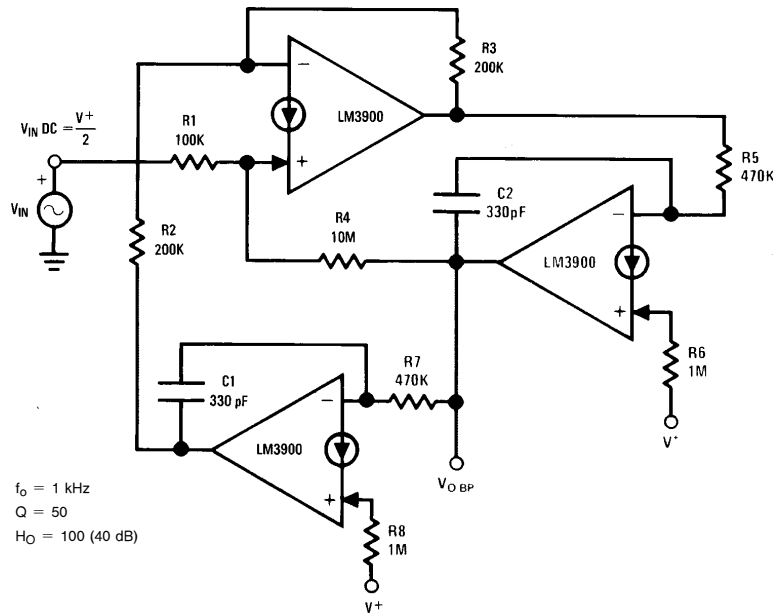
FIGURE 40. The "Bi-quad" RC Active Bandpass Filter

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ization technique for this type of filter is the "second-degree state-variable network." Outputs can be taken at any of three points to give low pass, high pass or bandpass response characteristics (see the reference cited).

The bandpass filter is shown in *Figure 40* and the design procedure is:

Given: Q and f_0.

To simplify: Let $C_1 = C_2$ and choose a convenient starting value and also let $2R_1 = R_2 = R_3$ and choose a convenient starting value.

Then:

$$R_4 = R_1 (2Q - 1), \tag{17}$$

$$R_5 = R_7 = \frac{1}{\omega_0 C_1}, \tag{18}$$

and for biasing the amplifiers we require

$$R_6 = R_8 = 2R_5. \tag{19}$$

The mid-band gain is:

$$H_O = \frac{R_4}{R_1}. \tag{20}$$

As a design example;

Require: $f_0 = 1$ kHz and Q = 50.

To find: C_1, C_2 and R_1 through R_8.

Choose: $C_1 = C_2 = 330$ pF

and $2R_1 = R_2 = R_3 = 360$ kΩ, and $R_1 = 180$ kΩ.

Then from equation (17),

$$R_4 = (1.8 \times 10^5) [2(50) - 1]$$

$$R_4 = 17.8 \text{ M}\Omega$$

From equation (18),

$$R_5 = R_7 = \frac{1}{(2\pi \times 10^3)(3.3 \times 10^{-10})}$$

$$R_5 = 483 \text{ k}\Omega.$$

And from equation (19),

$$R_6 = R_8 \cong 1 \text{ M}\Omega.$$

From equation (20) the midband gain is 100 (40 dB). The value of R_4 is high and can be lowered by scaling only R_1 through R_4 by the factor 1.78 to give:

$$2R_1 = R_2 = R_3 = \frac{360 \times 10^3}{1.78} = 200 \text{ k}\Omega, R_1 = 100 \text{ k}\Omega.$$

and

$$R_4 = \frac{17.8 \times 10^6}{1.78} = 10 \text{ M}\Omega.$$

These values (to the nearest 5% standard) have been added to *Figure 40*.

6.7 CONCLUSIONS

The unity-gain cross frequency of the LM3900 is 2.5 MHz which is approximately three times that of a "741" op amp. The performance of the amplifier does limit the performance of the filter. Historically, RC active filters started with little

concern for these practical problems. The sensitivity functions were a big step forward as these demonstrated that many of the earlier suggested realization techniques for RC active filters had passive component sensitivity functions which varied as Q or even Q^2. The Bi-Quad circuit has reduced the problems with the passive components (sensitivity functions of 1 or $\frac{1}{2}$) and recently the contributions of the amplifier on the performance of the filter are being investigated. An excellent treatment ("The Biquad: Part I — Some Practical Design Considerations," L.C. Thomas, IEEE Transactions on Circuit Theory, Vol. CT-18, No. 3, May 1971) has indicated the limits imposed by the characteristics of the amplifier by showing that the design value of Q (Q_D) will differ from the actual measured value of Q (Q_A) by the given relationship

$$Q_A = \frac{Q_D}{1 + \frac{2Q_D}{A_O \omega_a}(\omega_a - 2\omega_p)} \qquad (21)$$

where A_O is the open loop gain of the amplifier, ω_a is the dominant pole of the amplifier and ω_p is the resonant frequency of the filter. The result is that the trade-off between Q and center frequency (ω_p) can be determined for a given set of amplifier characteristics. When Q_A differs significantly from Q_D excessive dependence on amplifier characteristics is indicated. An estimate of the limitations of an amplifier can be made by arbitrarily allowing approximately a 10% effect on Q_A which results if

$$\frac{2Q_D}{A_O \omega_a}(\omega_a - 2\omega_p) = 0.1$$

or

$$\left(\frac{\omega_p}{\omega_a}\right) = 2.5 \times 10^{-2}\left(\frac{A_O}{Q_D}\right) + 0.5. \qquad (22)$$

As an example, using $A_O = 2800$ for the LM3900 we can estimate the maximum frequency where a $Q_D = 50$ would be reasonable as

$$\frac{f_p}{f_a} = 2.5 \times 10^{-2}\left(\frac{2.8 \times 10^3}{5 \times 10}\right) + 0.5$$

or

$$\frac{f_p}{f_a} = 1.9$$

therefore

$$f_p = 1.9\,f_a.$$

Again, using data of the LM3900, $f_a = 1$ kHz so this upper frequency limit is approximately 2 kHz for the assumed Q of 50. As indicated in equation (26) the value of Q_A can actually exceed the value of Q_D (Q enhancement) and, as expected, the filter can even provide its own input (oscillating). Excess phase shift in the high frequency characteristics of the amplifier typically cause unexpected oscillations. Phase compensation can be used in the Bi-Quad network to reduce this problem (see L.C. Thomas paper).

Designing for large passband gain also increases filter dependency on the characteristics of the amplifier and finally signal to noise ratio can usually be improved by taking gain in an input RC active filter (again see L.C. Thomas paper).

Somewhat larger Q's can be achieved by adding more filter sections in either a synchronously tuned cascade (filters tuned to same center frequency and taking advantage of the bandwidth shrinkage factor which results from the series connection) or as a standard multiple pole filter. All of the conventional filters can be realized and selection is based upon all of the performance requirements which the application demands. The cost advantages of the LM3900, the relatively large bandwidth and the ease of operation on a single power supply voltage make this product an excellent "building block" for RC active filters.

7.0 Designing Waveform Generators

The multiple amplifiers of the LM3900 can be used to easily generate a wide variety of waveforms in the low frequency range (f ≤ 10 kHz). Voltage controlled oscillators (VCO)'s are also possible and are presented in section 8.0 "Designing Phase-locked Loops and Voltage Controlled
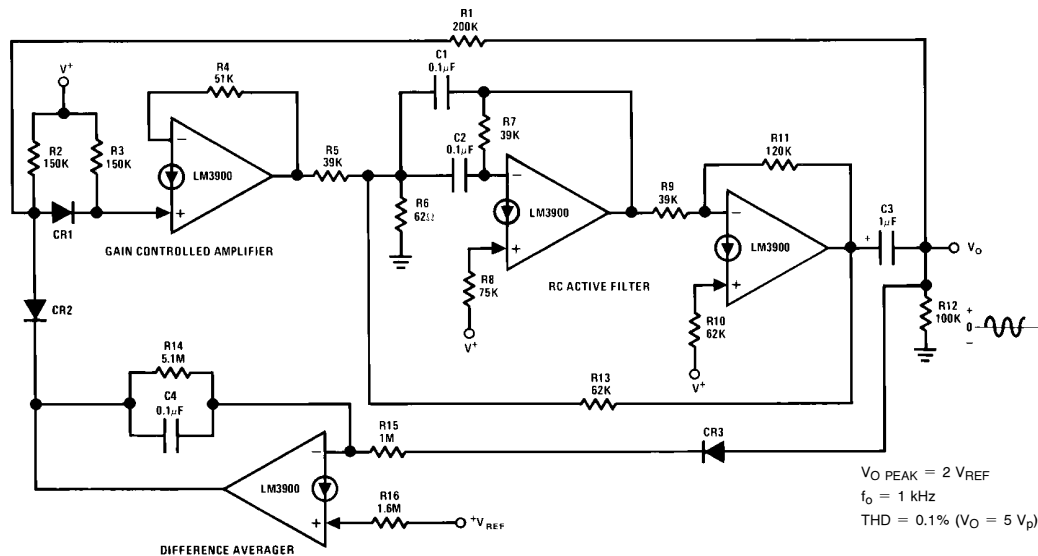

FIGURE 41. A Sinewave Oscillator

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Oscillators." In addition, power oscillators (such as noise makers, etc.) are presented in section 10.11.3. The waveform generators which will be presented in this section are mainly of the switching type, but for completeness a sinewave oscillator has been included.

7.1 A SINEWAVE OSCILLATOR

The design of a sinewave oscillator presents problems in both amplitude stability (and predictability) and output waveform purity (THD). If an RC bandpass filter is used as a high Q resonator for the oscillator circuit we can obtain an output waveform with low distortion and eliminate the problem of relative center frequency drift which exists if the active filter were used simply to filter the output of a separate oscillator.

A sinewave oscillator which is based on this principle is shown in *Figure 41*. The two-amplifier RC active filter is used as it requires only two capacitors and provides an overall non-inverting phase characteristic. If we add a non-inverting gain controlled amplifier around the filter we obtain the desired oscillator configuration. Finally, the sinewave output voltage is sensed and regulated as the average value is compared to a DC reference voltage, V_{REF}, by use of a differential averaging circuit. It can be shown that with the values selected for R_{15} and R_{16} (ratio of 0.64/1) that there is first order temperature compensation for CR_3 and the internal input diodes of the IC amplifier which is used for the "difference averager". Further, this also provides a simple way to regulate and to predict the magnitude of the output sinewave as

$$V_O \text{ peak} = 2 V_{REF}$$

which is essentially independent of both temperature and the magnitude of the power supply voltage (if V_{REF} is derived from a stable voltage source).

7.2 SQUAREWAVE GENERATOR

The standard op amp squarewave generator has been modified as shown in *Figure 42*. The capacitor, C_1, alternately
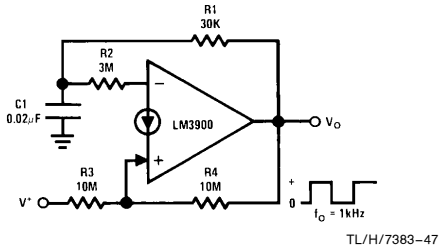

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FIGURE 42. A Squarewave Oscillator

charges and discharges (via R_1) between the voltage limits which are established by the resistors R_2, R_3 and R_4. This combination produces a Schmitt Trigger circuit and the operation can be understood by noticing that when the output is low (and if we neglect the current flow through R_4) the resistor R_2 (3M) will cause the trigger to fire when the current through this resistor equals the current which enters the ($+$) input (via R_3). This gives a firing voltage of approximately $R_2/(R_3)$ V^+ (or $V^+/3$). The other trip point, when the output voltage is high, is approximately $[2(R_2/R_3)]$ V^+, as $R_3 = R_4$, or $\frac{2}{3}(V^+)$. Therefore the voltage across the capacitor, C_1, will be the first one-half of an exponential waveform between these voltage trip limits and will have good symmetry and be essentially independent of the magnitude of the power supply voltage. If an unsymmetrical squarewave is desired, the trip points can be shifted to produce any desired mark/space ratio.

7.3 PULSE GENERATOR

The squarewave generator can be slightly modified to provide a pulse generator. The slew rate limits of the LM3900 (0.5V/μsec) must be kept in mind as this limits the ability to produce a narrow pulse when operating at a high power supply voltage level. For example, with a $+15$ V_{DC} power supply the rise time, t_r, to change 15V is given by:

$$t_r = \frac{15V}{\text{Slew Rate}} = \frac{15V}{0.5V/\mu sec}$$

$$t_r = 30 \ \mu sec.$$

The schematic of a pulse generator is shown in *Figure 43*. A diode has been added, CR_1, to allow separating the charge path to C_1 (via R_1) from the discharge path (via R_2). The
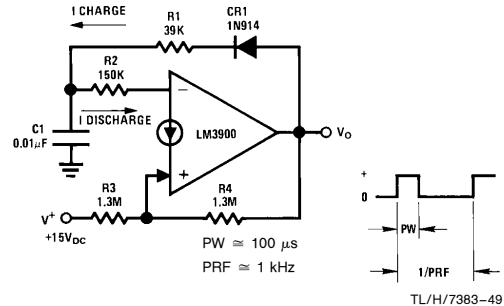

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FIGURE 43. A Pulse Generator

circuit operates as follows: Assume first that the output voltage has just switched low (and we will neglect the current flow through R_4). The voltage across C_1 is high and the magnitude of the discharge current (through R_2) is given by

$$I_{Discharge} \cong \frac{V_{C_1} - V_{BE}}{R_2}.$$

This current is larger than that entering the ($+$) input which is given by

$$I_{R_3} = \frac{V^+ - V_{BE}}{R_3}.$$

The excess current entering the ($-$) input terminal causes the amplifier to be driven to a low output voltage state (saturation). This condition remains for the long time interval (1/Pulse Repetition Frequency) until the R_2C_1 discharge current equals the I_{R_3} value (as CR_1 is OFF during this interval). The voltage across C_1 at the trip point, V_L, is given by

$$V_L = (I_{R_3}) (R_2),$$

or

$$V_L = (V^+ - V_{BE}) \left(\frac{R_2}{R_3}\right). \qquad (1)$$

At this time the output voltage will switch to a high state, V_{OHi}, and the current entering the ($+$) input will increase to

$$I_M{}^+ = \frac{V^+ - V_{BE}}{R_3} + \frac{V_{OHi} - V_{BE}}{R_4}.$$

Also CR_1 goes ON and the capacitor, C_1, charges via R_1. Some of this charge current is diverted via R_2 to ground (the (−) input is at $V_{CE_{SAT}}$ during this interval as the current mirror is demanding more current than the (−) input terminal can provide). The high trip voltage, V_H, is given by

$$V_H = (I_M{}^+) R_2 \quad \text{or}$$

$$V_H = \left(\frac{V^+ - V_{BE}}{R_3} + \frac{V_{OHi} - V_{BE}}{R_4} \right) R_2. \quad (2)$$

A design proceeds by first choosing the trip points for the voltage across C_1. The resistors R_3 and R_4 are used only for this trip voltage control. The resistor R_2 affects the discharge time (the long interval) and also both of the trip voltages so this resistor is determined first from the required pulse repetition frequency (PRF). The value of R_2 is determined by the RC exponential discharge from V_H to V_L as this time interval, T_1, controls the PRF ($T_1 = 1/\text{PRF}$). If we start with the equation for the RC discharge we have

$$V_L = V_H \, e^{-\dfrac{T_1}{R_2 C_1}}$$

or

$$\ln \frac{V_L}{V_H} = -\frac{T_1}{R_2 C_1}$$

or

$$T_1 = R_2 C_1 \ln \frac{V_H}{V_L} \quad (3)$$

To provide a low duty cycle pulse train we select small values for both V_H and V_L (such as 3V and 1.5V) and choose a starting value for C_1. Then R_2 is given by

$$R_2 = \frac{T_1}{C_1 \ln \dfrac{V_H}{V_L}}. \quad (4)$$

If R_2 from (4) is not in the range of approximately 100 kΩ to 1 MΩ, choose another value for C_1. Now equation (1) can be used to find a value for R_3 to provide the V_L which was initially assumed. Similarly equation (2) allows R_4 to be calculated. Finally R_1 is determined by the required pulse width (PW) as the capacitor, C_1, must be charged from V_L to V_H by R_1. This RC charging is given by (neglecting the loading due to R_2)

$$V_H \cong (V_{OHi} - V_D) \left(1 - e^{-\dfrac{T_2}{R_1 C_1}} \right)$$

or

$$T_2 \cong -R_1 C_1 \ln \left[1 - \frac{V_H}{V_{OHi} - V_D} \right], \text{ and finally}$$

$$R_1 \cong \frac{T_2}{-C_1 \ln \left[1 - \dfrac{V_H}{V_{OHi} - V_D} \right]} \quad (5)$$

where T_2 is the pulse width desired and V_D is the forward voltage drop across CR_1.

As a design example:

Required: Provide a 100 μs pulse every 1 ms. The power supply voltage is +15 V_{DC}

1.0 Start by choosing $V_L = 1.5V$

and $V_H = 3.0V$

2.0 Find R_2 from equation (4) assuming $C_1 = 0.01 \ \mu$F,

$$R_2 = \frac{10^{-3}}{10^{-8} \ln \left(\dfrac{3.0}{1.5} \right)}$$

$$R_2 = \frac{10^5}{0.694} = 144 \text{ k}\Omega$$

3.0 Find R_3 from equation (1)

$$R_3 = \frac{(V^+ - V_{BE}) R_2}{V_L}$$

$$R_3 = \frac{(15 - 0.5) 1.44 \times 10^5}{1.5}$$

$$R_3 = 1.39 \text{ M}\Omega$$

4.0 Find R_4 from equation (2),

$$R_4 = \frac{(V_{OHi} - V_{BE})}{\dfrac{V_H}{R_2} - \dfrac{V^+ - V_{BE}}{R_3}}$$

$$R_4 = \frac{(14.2 - 0.5)}{\dfrac{3}{1.44 \times 10^5} - \dfrac{15 - 0.5}{1.39 \times 10^6}}$$

$$R_4 = 1.32 \text{ M}\Omega$$

5.0 Find R_1 from equation (5),

$$R_1 = \frac{10^{-4}}{-10^{-8} \ln \left(1 - \dfrac{3}{(14.2 - 0.7)} \right)}$$

$$R_1 = \frac{10^4}{-\ln \left(1 - \dfrac{3}{13.5} \right)}$$

$$R_1 = \frac{10^4}{0.252} = 39.7 \text{ k}\Omega.$$

These values (to the nearest 5% standard) have been added to *Figure 43*.

7.4 TRIANGLE WAVEFORM GENERATOR

Triangle waveforms are usually generated by an integrator which receives first a positive DC input voltage, then a negative DC input voltage. The LM3900 easily provides this operation in a system which operates with only a single power supply voltage by making use of the current mirror which exists at the (+) input. This allows the generation of a triangle waveform without requiring a negative DC input voltage.

The schematic diagram of a triangle waveform generator is shown in *Figure 44*. One amplifier is doing the integration by
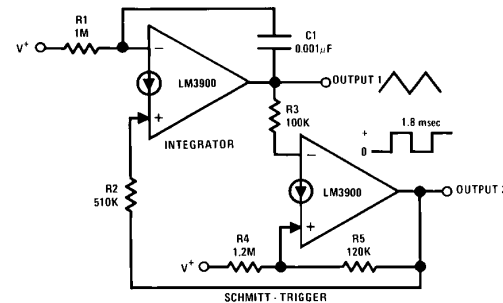

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FIGURE 44. A Triangle Waveform Generator

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operating first with the current through R_1 to produce the negative output voltage slope, and then when the output of the second amplifier (the Schmitt Trigger) is high, the current through R_2 causes the output voltage to increase. If $R_1 = 2R_2$, the output waveform will have good symmetry. The timing for one-half of the period (T/2) is given by

$$\frac{T}{2} = \frac{(R_1 C_1)\Delta V_O}{V^+ - V_{BE}}$$

or the output frequency becomes

$$f_O = \frac{V^+ - V_{BE}}{2R_1 C_1 \Delta V_O}$$

where we have assumed $R_1 = 2R_2$, V_{BE} is the DC voltage at the ($-$) input (0.5 V_{DC}), and ΔV_O is the difference between the trip points of the Schmitt Trigger. The design of the Schmitt Trigger has been presented in the section on Digital and Switching Circuits (9.0) and the trip voltages control the peak-to-peak excursion of the triangle output voltage waveform. The output of the Schmitt circuit provides a squarewave of the same frequency.

7.5 SAWTOOTH WAVEFORM GENERATOR

The previously described triangle waveform generator, *Figure 44*, can be modified to produce a sawtooth waveform. Two types of waveforms can be provided, both a positive ramp and a negative ramp sawtooth waveform by selecting R_1 and R_2. The reset time is also controlled by the ratio of R_1 and R_2. For example, if $R_1 = 10\ R_2$ a positive ramp sawtooth results and if $R_2 = 10\ R_1$ a negative ramp sawtooth can be obtained. Again, the slew rate limits of the amplifier (0.5V/μs) will limit the minimum retrace time, and the increased slew rate of a negative going output will allow a faster retrace for a positive ramp sawtooth waveform.

To provide a gated sawtooth waveform, the circuits shown in *Figure 45* can be used. In *Figure 45(a)*, a positive ramp is generated by integrating the current, I, which is entering the ($+$) input. Reset is provided via R_1 and CR_1 keeps R_1 from loading at the ($-$) input during the sweep interval. This will sweep from $V_{O\ MIN}$ to $V_{O\ MAX}$ and will remain at $V_{O\ MAX}$ until reset. The interchange of the input leads, *Figure 45(b)*, will generate a negative ramp, from $V_{O\ MAX}$ to $V_{O\ MIN}$.
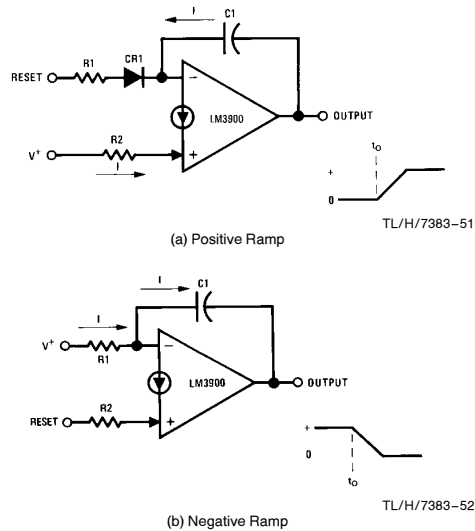

(a) Positive Ramp

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(b) Negative Ramp

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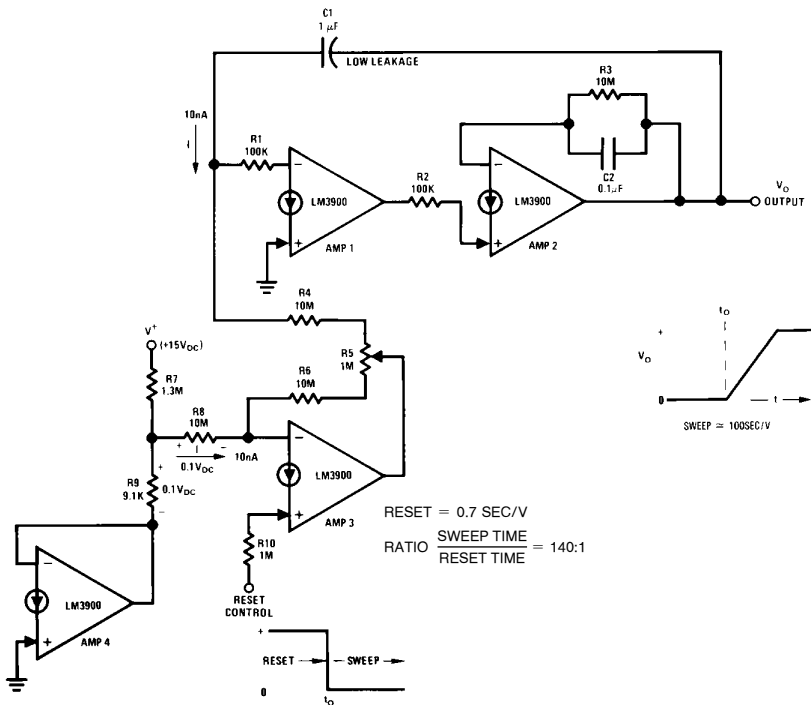**FIGURE 45. Gated Sawtooth Generators**

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FIGURE 46. Generating Very Slow Sawtooth Waveforms

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7.5.1 GENERATING A VERY SLOW SAWTOOTH WAVEFORM

The LM3900 can be used to generate a very slow sawtooth waveform which can be used to generate long time delay intervals. The circuit is shown in *Figure 46* and uses four amplifiers. Amps 1 and 2 are cascaded to increase the gain of the integrator and the output is the desired very slow sawtooth waveform. Amp 3 is used to exactly supply the bias current to Amp 1.

With resistor R_8 opened up and the reset control at zero volts, the potentiometer, R_5, is adjusted to minimize the drift in the output voltage of Amp 2 (this output must be kept in the linear range to insure that Amp 2 is not in saturation). Amp 4 is used to provide a bias reference which equals the DC voltage at the ($-$) input of Amp 3. The resistor divider, R_7 and R_9 provides a 0.1 V_{DC} reference voltage across R_9 which also appears across R_8. The current which flows through R_8, I, enters the ($-$) input of Amp 3 and causes the current through R_6 to drop by this amount. This causes an imbalance as now the current flow through R_4 is no longer adequate to supply the input current of Amp 1. The net result is that this same current, I, is drawn from capacitor C_1 and causes the output voltage of Amp 2 to sweep slowly positive. As a result of the high impedance values used, the PC component board used for this circuit must first be cleaned and then coated with silicone rubber to eliminate the effects of leakage currents across the surface of the board. The DC leakage currents of the capacitor, C_1, must also be small compared to the 10 nA charging current. For example, an insulation resistance of 100,000 MΩ will leak 0.1 nA with 10 V_{DC} across the capacitor and this leakage rapidly increases at higher temperatures. Dielectric polarization of the dielectric material may not cause problems if the circuit is not rapidly cycled. The resistor, R_8, and the capacitor, C_1, can be scaled to provide other basic sweep rates. For the values shown on *Figure 46* the 10 nA current and the 1μF capacitor establish a sweep rate of 100 sec/volt. The reset control pulse (Amp 3 (+) input) causes Amp 3 to go to the positive output saturation state and the 10 MΩ (R_4) gives a reset rate of 0.7 sec/volt. The resistor, R_1, prevents a large discharge current of C_1 from overdriving the ($-$) input and overloading the input clamp device. For larger charging currents, a resistor divider can be placed from the output of Amp 4 to ground and R_8 can tie from this tap point directly to the ($-$) input of Amp 1.

7.6 STAIRCASE WAVEFORM GENERATORS

A staircase generator can be realized by supplying pulses to an integrator circuit. The LM3900 also can be used with a squarewave input signal and a differentiating network where each transition of the input squarewave causes a step in the output waveform (or two steps per input cycle). This is shown in *Figure 47*. These pulses of current are the charge and discharge currents of the input capacitor, C_1. The charge current, I_C, enters the ($+$) input and is mirrored about ground and is "drawn into" the ($-$) input. The discharge current, I_D, is drawn through the diode at the input, CR_1, and therefore also causes a step on the output staircase.

A free running staircase generator is shown in *Figure 48*. This uses all four of the amplifiers which are available in one LM3900 package.

Amp 1 provides the input pulses which "pump up" the staircase via resistor R_1 (see section 7.3 for the design of this pulse generator). Amp 2 does the integrate and hold function and also supplies the output staircase waveform. Amps 3 and 4 provide both a compare and a one-shot multivibrator function (see the section on Digital and Switching Circuits for the design of this dual function one-shot). Resistor R_4 is used to sample the staircase output voltage and to compare it with the power supply voltage (V$^+$) via R_3. When the output exceeds approximately 80% of V$^+$ the connection of Amps 3 and 4 causes a 100 μsec reset pulse to be generated. This is coupled to the integrator (Amp 2) via R_2 and causes the staircase output voltage to fall to approximately zero volts. The next pulse out of Amp 1 then starts a new stepping cycle.

7.7 A PULSE COUNTER AND A VOLTAGE VARIABLE PULSE COUNTER

The basic circuit of *Figure 48* can be used as a pulse counter simply by omitting Amp 1 and feeding input voltage pulses directly to R_1. A simpler one-shot/comparator which requires only one amplifier can also be used in place of Amps 3 and 4 (again, see the section on Digital and Switching Circuits). To extend the time interval between pulses, an additional amplifier can be used to supply base current to Amp 2 to eliminate the tendency for the output voltage to drift up due to the 30 nA input current (see section 7.5.1). The pulse count can be made voltage variable simply by removing the comparator reference (R_3) from V$^+$ and using this as a control voltage input. Finally, the input could be derived from differentiating a squarewave input as was shown in *Figure 47* and if only one step per cycle were desired, the diode, CR_1 of *Figure 47*, can be eliminated.
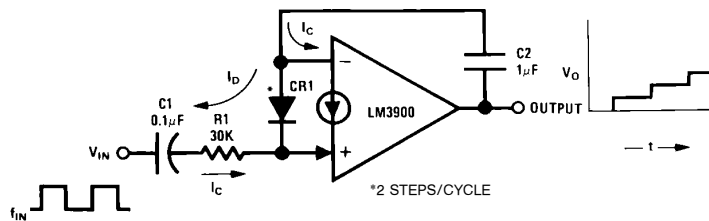

FIGURE 47. Pumping the Staircase Via Input Differentiator

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7.8 AN UP-DOWN STAIRCASE WAVEFORM GENERATOR

A staircase waveform which first steps up and then steps down is provided by the circuit shown in *Figure 49*. An input pulse generator provides the pulses which cause the output to step up or down depending on the conduction of the clamp transistor, Q_1. When this is ON, the "down" cur-

rent pulse is diverted to ground and the staircase then steps "up". When the upper voltage trip point of Amp 2 (Schmitt Trigger—see section on Digital and Switching Circuits) is reached, Q_1 goes OFF and as a result of the smaller "down" input resistor (one-half the value of the "up" resistor, R_1) the staircase steps "down" to the low voltage trip point of Amp 2. The output voltage therefore steps up and down between the trip voltages of the Schmitt Trigger.
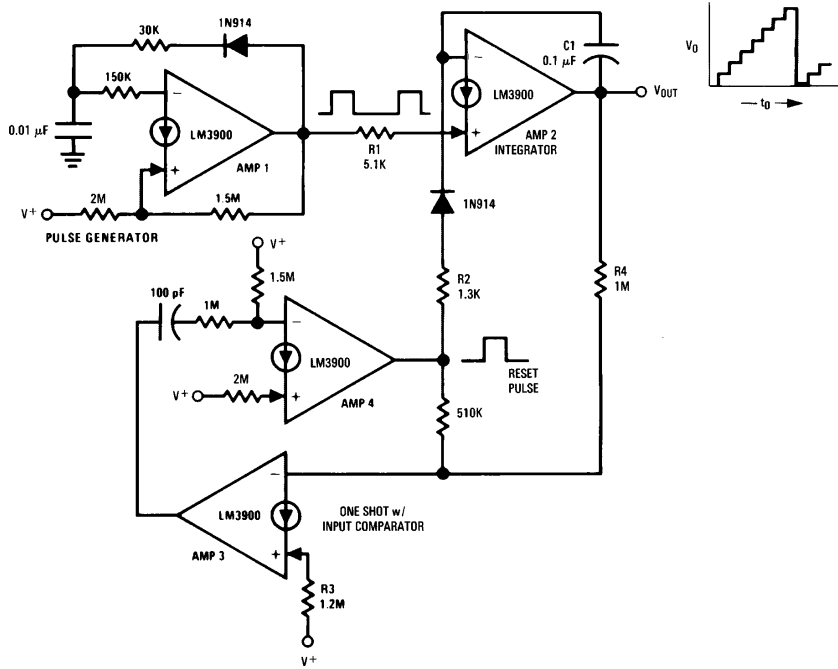

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FIGURE 48. A Free Running Staircase Generator

TL/H/7383–56

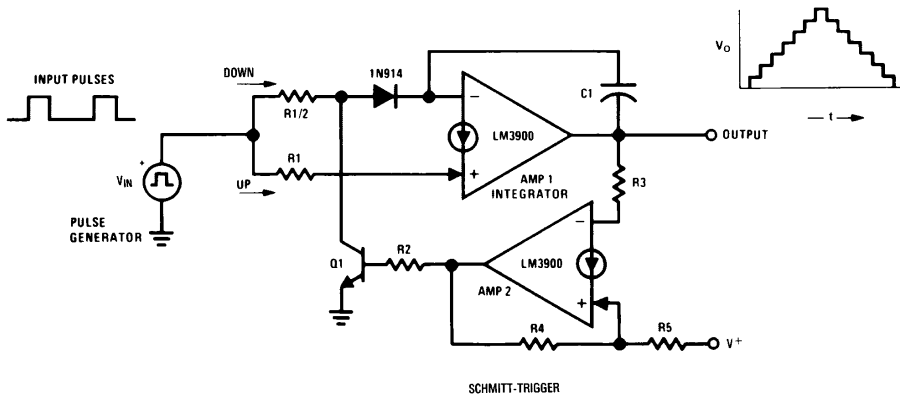**FIGURE 49. An Up-down Staircase Generator**

8.0 Designing Phase-Locked Loops and Voltage Controlled Oscillators

The LM3900 can be connected to provide a low frequency (f < 10 kHz) phase-locked loop (PL[2]). This is a useful circuit for many control applications. Tracking filters, frequency to DC converters, FM modulators and demodulators are applications of a PL[2].

8.1 VOLTAGE CONTROLLED OSCILLATORS (VCO)

The heart of a PL[2] is the voltage controlled oscillator (VCO). As the PL[2] can be used for many functions, the required linearity of the transfer characteristic (frequency out vs. DC voltage in) depends upon the application. For low distortion demodulation of an FM signal, a high degree of linearity is necessary whereas a tracking filter application would not require this performance in the VCO.

A VCO circuit is shown in *Figure 50*. Only two amplifiers are required, one is used to integrate the DC input control voltage, V_C, and the other is connected as a Schmitt-trigger which monitors the output of the integrator. The trigger circuit is used to control the clamp transistor, Q_1. When Q_1 is conducting, the input current, I_2, is shunted to ground. During this one-half cycle the input current, I_1, causes the output voltage of the integrator to ramp down. At the minimum point of the triangle waveform (output 1), the Schmitt circuit changes state and transistor Q_1 goes OFF. The current, I_2, is exactly twice the value of I_1 ($R_2 = R_1/2$) such that a charge current (which is equal to the magnitude of the discharge current) is drawn through the capacitor, C, to provide the increasing portion of the triangular waveform (output 1).

The output frequency for a given DC input control voltage depends on the trip voltages of the Schmitt circuit (V_H and V_L) and the components R_1 and C_1 (as $R_2 = R_1/2$). The

time to ramp down from V_H to V_L corresponds to one-half the period (T) of the output frequency and can be found by starting with the basic equation of the integrator

$$V_O = -\frac{1}{C} \int I_1 \, dt \qquad (1)$$

as I_1 is a constant (for a given value of V_C) which is given by

$$I_1 = \frac{V_C - V_{BE}}{R_1} \qquad (2)$$

equation (1) simplifies to

$$\Delta V_O = -\frac{I_1}{C}(\Delta t)$$

or

$$\frac{\Delta V_O}{\Delta t} = -\frac{I_1}{C} \qquad (3)$$

Now the time, Δt, to sweep from V_H to V_L becomes

$$\Delta t_1 = \frac{(V_H - V_L) C}{I_1} \quad \text{or}$$

$$T = \frac{2 (V_H - V_L) C}{I_1} \quad \text{and}$$

$$f = \frac{1}{T} = \frac{I_1}{2 (V_H - V_L) C}. \qquad (4)$$

Therefore, once V_H, V_L, R_1 and C are fixed in value, the output frequency, f, is a linear function of I_1 (as desired for a VCO).
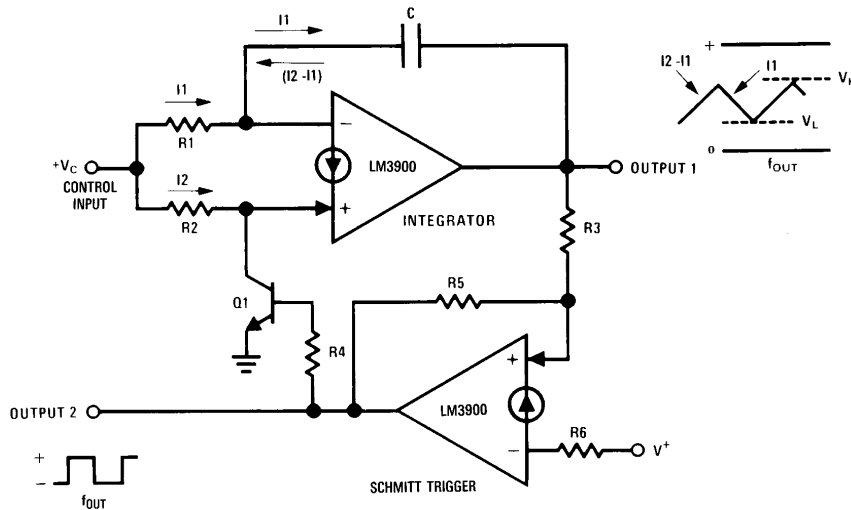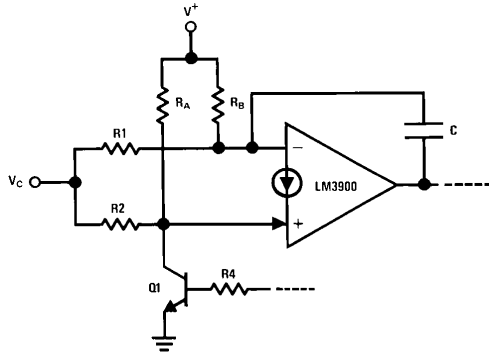

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**FIGURE 50. A Voltage Controlled Oscillator**

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FIGURE 51. Adding Input Common-mode Biasing Resistors

The circuit shown in *Figure 50* will require $V_C > V_{BE}$ to oscillate. A value of $V_C = 0$ provides $f_{OUT} = 0$, which may or may not be desired. Two common-mode input biasing resistors can be added as shown in *Figure 51* to allow $f_{OUT} = f_{MIN}$ for $V_C = 0$. In general, if these resistors are a factor of 10 larger than their corresponding resistor (R_1 or R_2) a large control frequency ratio can be realized. Actually, V_C could range outside the supply voltage limit of V^+ and this circuit will still function properly.

The output frequency of this circuit can be increased by reducing the peak-to-peak excursion of the triangle waveform (output 1) by design of the trip points of the Schmitt circuit. A limit is reached when the triangular sweep output waveform exceeds the slew rate limit of the LM3900 (0.5 V/μs). Note that the output of the Schmitt circuit has to move up only one V_{BE} to bring the clamp transistor, Q_1, ON, and therefore output slew rate of this circuit is not a limit.
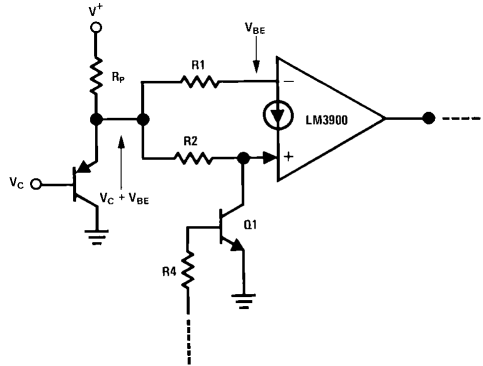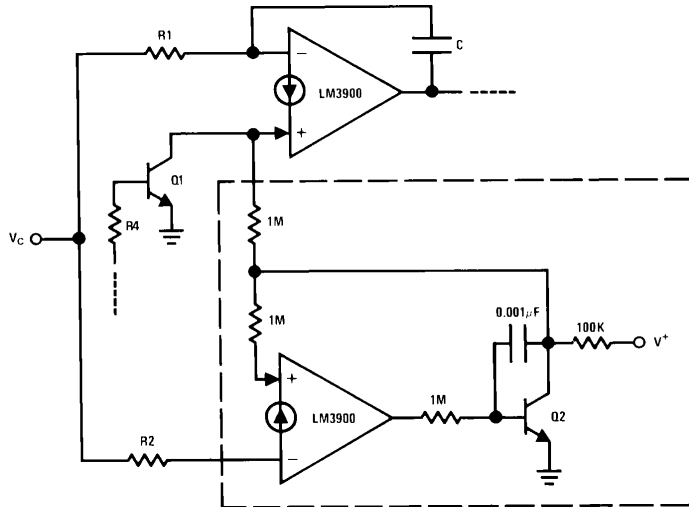

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FIGURE 52. Reducing Temperature Drift

To improve the temperature stability of the VCO, a PNP emitter follower can be used to give approximate compensation for the V_{BE}'s at the inputs to the amplifier (see *Figure 52*). Finally to improve the mark to space ratio accuracy over temperature and at low control voltages, an additional amplifier can be added such that both reference currents are applied to the same type of (inverting) inputs of the LM3900. The circuit to accomplish this is shown within dotted lines in *Figure 53*.

8.2 PHASE COMPARATOR

A basic phase comparator is shown in *Figure 54*. This circuit provides a pulse-width modulated output voltage waveform, V_O1, which must be filtered to provide a DC output voltage (this filter can be the same as the one needed in the PL[2]). The resistor R_2 is made smaller than R_1 so the ($+$) input serves to inhibit the ($-$) input signal. The center of the

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FIGURE 53. Improving Mark/Space Ratio

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dynamic range is indicated by the waveforms shown on the figure (90° phase difference between f_{IN} and f_{VCO}).
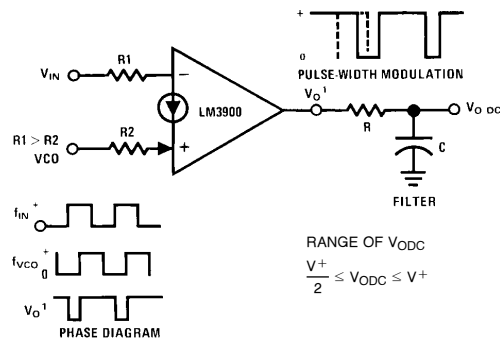

TL/H/7383–61

FIGURE 54. Phase Comparator

The filtered DC output voltage will center at $3V^+/4$ and can range from $V^+/2$ to V^+ as the phase error ranges from 0 degrees to 180 degrees.

8.3 A COMPLETE PHASE-LOCKED LOOP

A phase-locked loop can be realized with three of the amplifiers as shown in *Figure 55*. This has a center frequency of approximately 3 kHz. To increase the lock range, DC gain can be added at the input to the VCO by using the fourth amplifier of the LM3900. If the gain is inverting, the limited DC dynamic range out of the phase detector can be in-creased to improve the frequency lock range. With inverting gain, the input to the VCO could go to zero volts. This will cause the output of the VCO to go high (V^+) and will latch if applied to the ($+$) input of the phase comparator. Therefore apply the VCO signal to the ($-$) input of the phase comparator or add the common-mode biasing resistors of *Figure 51*.

8.4 CONCLUSIONS

One LM3900 package (4 amplifiers) can provide all of the operations necessary to make a phase-locked loop. In addition, a VCO is a generally useful component for other system applications.

9.0 Designing Digital and Switching Circuits

The amplifiers of the LM3900 can be over-driven and used to provide a large number of low speed digital and switching circuit applications for control systems which operate off of single power supply voltages larger than the standard $+5\ V_{DC}$ digital limit. The large voltage swing and slower speed are both advantages for most industrial control systems. Each amplifier of the LM3900 can be thought of as "a super transistor" with a β of 1,000,000 (25 nA input current and 25 mA output current) and with a non-inverting input feature. In addition, the active pull-up and pull-down which exists at the output will supply larger currents than the simple resistor pull-ups which are used in digital logic gates. Finally, the low input currents allow timing circuits which minimize the capacitor values as large impedance levels can be used with the LM3900.
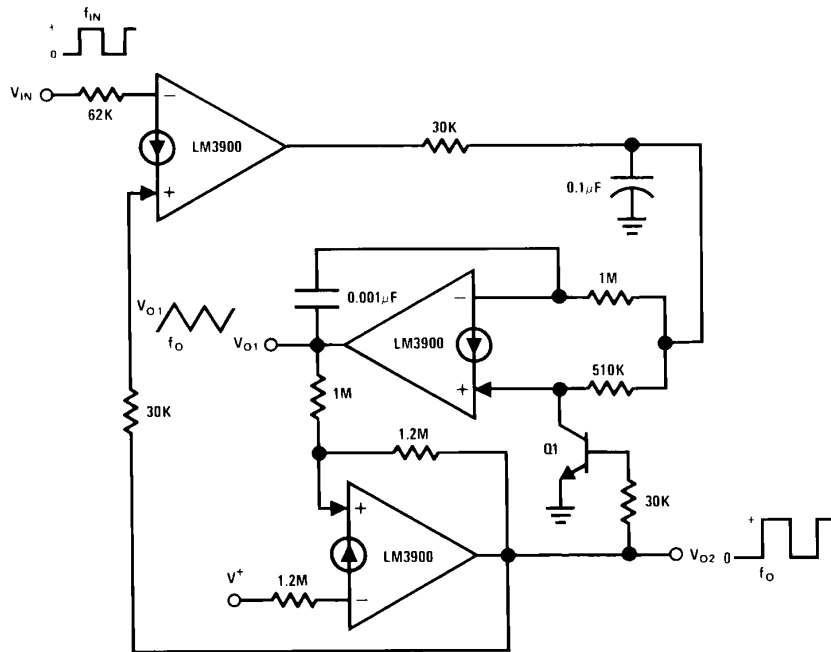

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FIGURE 55. A Phase-locked Loop

9.1 AN "OR" GATE

An OR gate can be realized by the circuit shown in *Figure 56*. A resistor (150 kΩ) from V^+ to the $(-)$ input keeps the output of the amplifier in a low voltage saturated state for all inputs A, B, and C at 0V. If any one of the input signals were to go high ($\cong V^+$) the current flow through the 75 kΩ input resistor will cause the amplifier to switch to the positive output saturation state ($V_O \cong V^+$). The current loss through the other input resistors (which have an input in the low voltage state) represents an insignificant amount of the total input current which is provided by the, at least one, high voltage input. More than three inputs can be OR'ed if desired.

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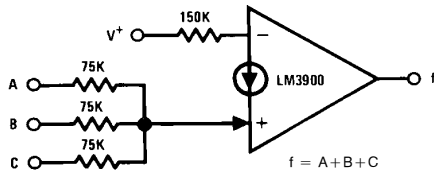
FIGURE 56. An "OR" Gate

The "fan-out" or logical drive capability is large (50 gates if each gate input has a 75 kΩ resistor) due to the 10 mA output current capability of the LM3900. A NOR gate can be obtained by interchanging the inputs to the LM3900.

9.2 AN "AND" GATE

A three input AND gate is shown in *Figure 57*. This gate requires all three inputs to be high in order to have sufficient current entering the $(+)$ input to cause the output of the amplifier to switch high. The addition of R_2 causes a smaller current to enter the $(+)$ input when only two of the inputs are high. (A two input AND gate would not require a resistor
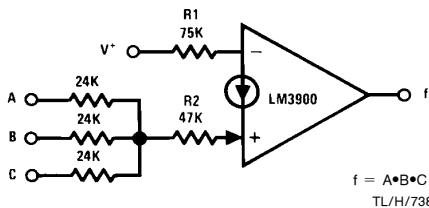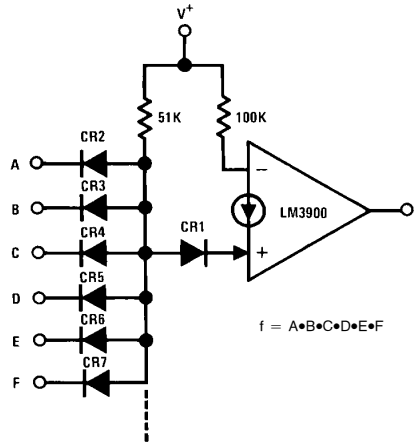

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FIGURE 57. An "AND" Gate

as R_2). More than three inputs becomes difficult with this resistor summing approach as the $(+)$ input is too close to having the necessary current to switch just prior to the last input going high. For a larger fan-in an input diode network

(similar to DTL) is recommended as shown in *Figure 58*. Interchange the inputs for a NAND gate.


All Diodes 1N914 or Equiv.   TL/H/7383–65

FIGURE 58. A Large Fan-in "AND" Gate

9.3 A BI-STABLE MULTIVIBRATOR

A bi-stable multivibrator (as asynchronous RS flip-flop) can be realized as shown in *Figure 59*. Positive feedback is provided by resistor R_4 which causes the latching. A positive pulse at the "set" input causes the output to go high and a "reset" positive pulse will return the output to essentially $0V_{DC}$.
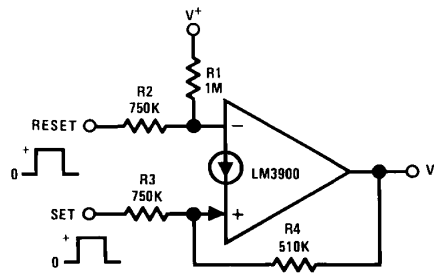
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FIGURE 59. A Bi-stable Multivibrator

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9.4 TRIGGER FLIP FLOPS

Trigger flip flops are useful to divide an input frequency as each input pulse will cause the output of a trigger flip flop to change state. Again, due to the absence of a clocking signal input, this is for an asynchronous logic application. A circuit which uses only one amplifier is shown in *Figure 60*. Steering of the differentiated positive input trigger is provided by the diode CR2. For a low output voltage state, CR2 shunts the trigger away from the (−) input and resistor R_3 couples this positive input trigger to the (+) input terminal. This causes the output to switch high. The high voltage output state now keeps CR2 OFF and the smaller value of (R_5 + R_6) compared with R_3 causes a larger positive input trigger

to be coupled to the (−) input which causes the output to switch to the low voltage state.

A second trigger flip flop can be made which consists of two amplifiers and also provides a complementary output. This connection is shown in *Figure 61*.

9.5 MONOSTABLE MULTIVIBRATORS (ONE-SHOTS)

Monostable multivibrators can be made using one or two of the amplifiers of the LM3900. In addition, the output can be designed to be either high or low in the quiescent state. Further, to increase the usefulness, a one-shot can be designed which triggers at a particular DC input voltage level to serve the dual role of providing first a comparator and then a pulse generator.
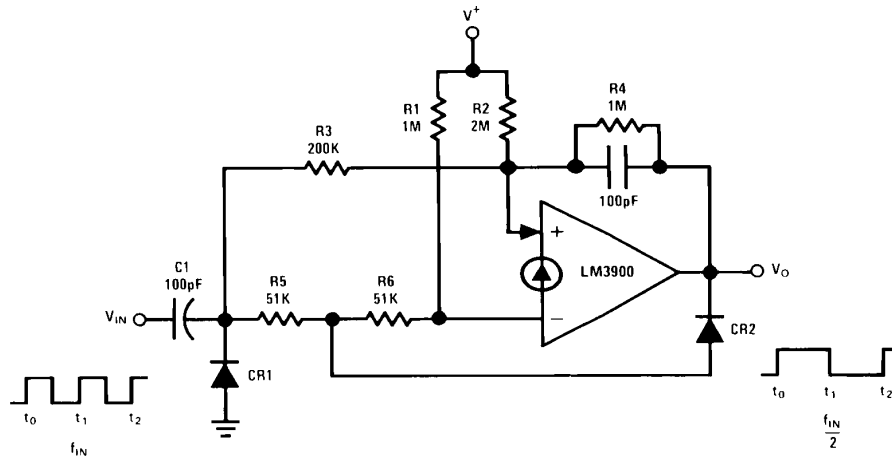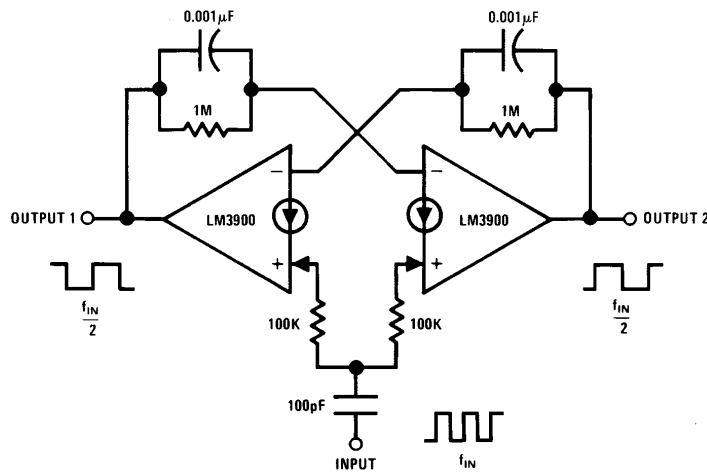

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FIGURE 60. A Trigger Flip Flop

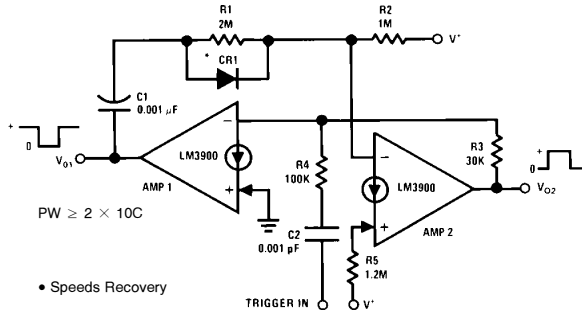

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FIGURE 61. A Two-amplifier Trigger Flip Flop

FIGURE 62. A One-shot Multivibrator

9.5.1 A TWO-AMPLIFIER ONE-SHOT

A circuit for a two-amplifier one-shot is shown in *Figure 62*. As the resistor, R_2, from V^+ to the $(-)$ input is smaller than R_5 (from V^+ to the $(+)$ input), amplifier 2 will be biased to a low-voltage output in the quiescent state. As a result, no current is supplied to the $(-)$ input of amplifier 1 (via R_3) which causes the output of this amplifier to be in the high voltage state. Capacitor C_1 therefore has essentially the full V^+ supply voltage across it ($V^+ - 2\ V_{BE}$). Now when a differentiated trigger (due to C_2) causes amplifier 1 to be driven ON (output voltage drops to essentially zero volts) this negative transient is coupled (via C_1) to the $(-)$ input of amplifier 2 which causes the output of this amplifier to be driven high (to positive saturation). This condition remains while C_1 discharges via (R_1) from approximately V^+ to approximately $V^+/2$. This time interval is the pulse width (PW). After C_1 no longer diverts sufficient current of R_2 away from the $(-)$ input of amplifier 2 (i.e., C_1 is discharged to approximately $V^+/2$ V) the stable DC state is restored—amplifier 2 output low and amplifier 1 output high.

This circuit can be rapidly re-triggered due to the action of the diode, CR_1. This re-charges C_1 as amplifier 1 drives full output current capability (approximately 10 mA) through C_1, CR_1 and into the saturated $(-)$ input of amplifier 2 to ground. The only time limit is the 10 mA available from amplifier 1 and the value of C_1. If a rapid reset is not required, CR_1 can be omitted.
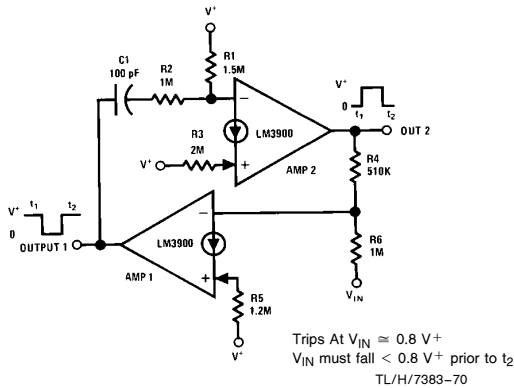
9.5.2 A COMBINATION ONE-SHOT/COMPARATOR CIRCUIT

In many applications a pulse is required if a DC input signal exceeds a predetermined value. This exists in free-running oscillators where after a particular output level has been reached a reset pulse must be generated to recycle the oscillator. This double function is provided with the circuit of *Figure 63*. The resistors R_5 and R_6 of amplifier 1 provide the inputs to a comparator and, as shown, an input signal, V_{IN}, is compared with the supply voltage, V^+. The output voltage of amplifier 1 is normally in a high voltage state and will fall and initiate the generation of the output pulse when V_{IN} is $R_6/R_5\ V^+$ or approximately 80% of V^+. To keep V_{IN} from disturbing the pulse generation it is required that V_{IN} fall to less than the trip voltage prior to the termination of the output pulse. This is the case when this circuit is used to generate a reset pulse and therefore this causes no problems.

9.5.3 A ONE-AMPLIFIER ONE-SHOT (POSITIVE PULSE)

A one-shot circuit can be realized using only one amplifier as shown in *Figure 64*.

The resistor R_2 keeps the output in the low voltage state. A differentiated positive trigger causes the output to switch to the high voltage state and resistor R_5 latches this state. The capacitor, C_1, charges from essentially ground to approximately $V^+/4$ where the circuit latches back to the quiescent state. The diode, CR_1, is used to allow a rapid re-triggering.
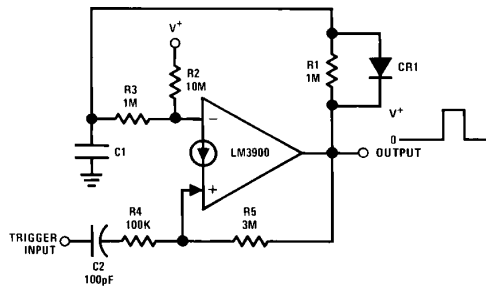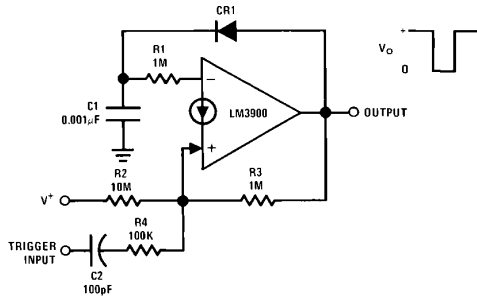

Trips At $V_{IN} \cong 0.8\ V^+$
V_{IN} must fall $< 0.8\ V^+$ prior to t_2

FIGURE 63. A One-shot Multivibrator with an Input Comparator

FIGURE 64. A One-amplifier One-shot (Positive Output)

9.5.4 A ONE-AMPLIFIER ONE-SHOT (NEGATIVE PULSE)

A one-amplifier one-shot multivibrator which has a quiescent state with the output high and which falls to zero volts for the pulse duration is shown in *Figure 65*.

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**FIGURE 65. A One-Amplifier One-Shot
(Negative Output)**

The sum of the currents through R_2 and R_3 keeps the $(-)$ input at essentially ground. This causes V_O to be in the high voltage state. A differentiated negative trigger waveform causes the output to switch to the low voltage state. The large voltage across C_1 now provides input current via R_1 to keep the output low until C_1 is discharged to approximately $V^+/10$. At this time the output switches to the stable high voltage state.

If the R_4C_2 network is moved to the $(-)$ input terminal, the circuit will trigger on a differentiated positive trigger waveform.

9.6 COMPARATORS

The voltage comparator is a function required for most system operations and can easily be performed by the LM3900. Both an inverting and a non-inverting comparator can be obtained.

9.6.1 A COMPARATOR FOR POSITIVE INPUT VOLTAGES

The circuit in *Figure 66* is an inverting comparator. To insure proper operation, the reference voltage must be larger than
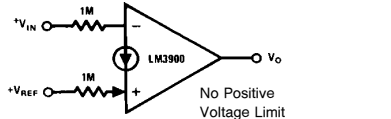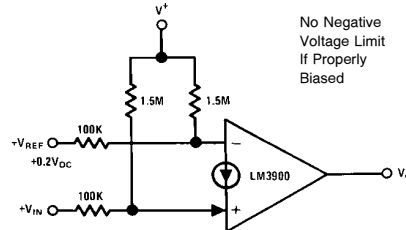

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FIGURE 66. An Inverting Voltage Comparator

V_{BE}, but there is no upper limit as long as the input resistor is large enough to guarantee that the input current will not exceed 200 μA.

9.6.2 A COMPARATOR FOR NEGATIVE INPUT VOLTAGES

Adding a common-mode biasing network to the comparator in *Figure 66* makes it possible to compare voltages between zero and one volt as well as the comparison of rather large negative voltages, *Figure 67*. When working with negative voltages, the current supplied by the common-mode network must be large enough to satisfy both the current drain demands of the input voltages and the bias current requirement of the amplifier.
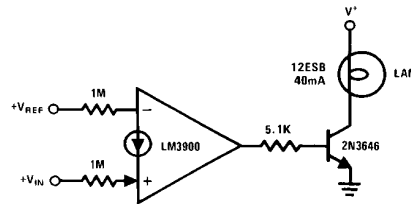

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FIGURE 67. A Non-inverting Low-voltage Comparator

9.6.3 A POWER COMPARATOR

When used in conjunction with an external transistor, this power comparator will drive loads which require more current than the IC amplifier is capable of supplying. *Figure 68* shows a non-inverting comparator which is capable of driving a 12V, 40 mA panel lamp.
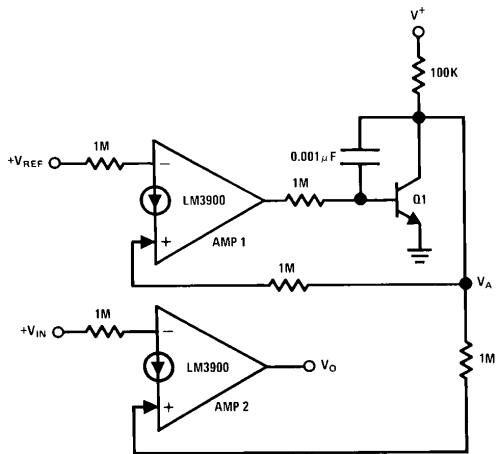

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FIGURE 68. A Non-inverting Power Comparator

9.6.4 A MORE PRECISE COMPARATOR

A more precise comparator can be designed by using a second amplifier such that the input voltages of the same type of inputs are compared. The $(-)$ input voltages of two amplifiers are naturally more closely matched initially and track well with temperature changes. The comparator of *Figure 69* uses this concept.
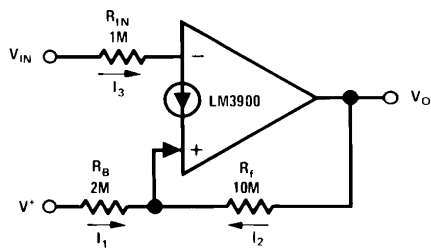
FIGURE 69. A More Precise Comparator

The current established by V_{REF} at the inverting input of amplifier 1 will cause transistor Q_1 to adjust the value of V_A to supply this current. This value of V_A will cause an equal current to flow into the non-inverting input of amplifier 2. This current corresponds more exactly to the reference current of amplifier 1.
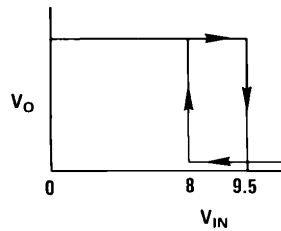
A differential input stage can also be added to the LM3900 (see section 10.16) and the resulting circuit can provide a precision comparator circuit.
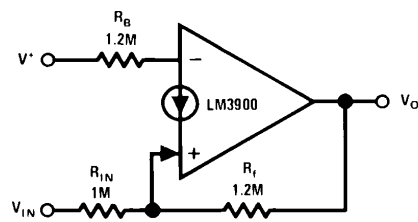
9.7 SCHMITT TRIGGERS

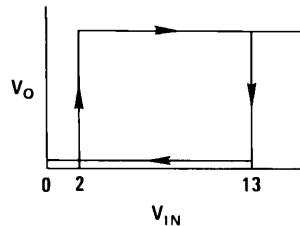Hysteresis may be designed into comparators which use the LM3900 as shown in *Figure 70*.

The lower switch point for the inverting Schmitt-Trigger is determined by the amount of current flowing into the positive input with the output voltage low. When the input current, I_3, drops below the level required by the current mirror, the output will switch to the high limit. With V_O high, the current demanded by the mirror is increased by a fixed amount, I_2. As a result, the I_3 required to switch the output increases this same amount. Therefore, the switch points are determined by selecting resistors which will establish the required currents at the desired input voltages. Reference current (I_1) and feedback current (I_2) are set by the following equation.

$$I_1 = \frac{V^+ - \phi}{R_B}$$

$$I_2 = \frac{V_{O\,MAX} - \phi}{R_F}$$

By adjusting the values of R_B, R_F, and R_{IN}, the switching values of V_{IN} may be set to any levels desired.

The non-inverting Schmitt Trigger works in the same way except that the input voltage is applied to the (+) input. The range of V_{IN} may be very large when compared with the operating voltage of the amplifier.

10.0 Some Special Circuit Applications

This section contains various special circuits which did not fit the order of things or which are one-of-a-kind type of applications.

10.1 CURRENT SOURCES AND SINKS

The amplifiers of the LM3900 can be used in feedback loops which regulate the current in external PNP transistors to provide current sources or in external NPN transistors to provide current sinks. These can be multiple sources or single sources which are fixed in value or made voltage variable.

(a) Inverting

(b) Non-inverting

FIGURE 70. Schmitt Triggers

10.1.1 A FIXED CURRENT SOURCE

A multiple fixed current source is provided by the circuit of *Figure 71*. A reference voltage (1 V_{DC}) is established across resistor R_3 by the resistive divider (R_3 and R_4). Negative feedback is used to cause the voltage drop across R_1 to also be 1 V_{DC}. This controls the emitter current of transistor Q_1 and if we neglect the small current diverted into the $(-)$ input via the 1M input resistor (13.5 μA) and the base current of Q_1 and Q_2 (an additional 2% loss if the β of these transistors is 100), essentially this same current is available out of the collector of Q_1.

Larger input resistors can be used to reduce current loss and a Darlington connection can be used to reduce errors due to the β of Q_1.


$$I_2 = \frac{R1}{R2} I_1$$
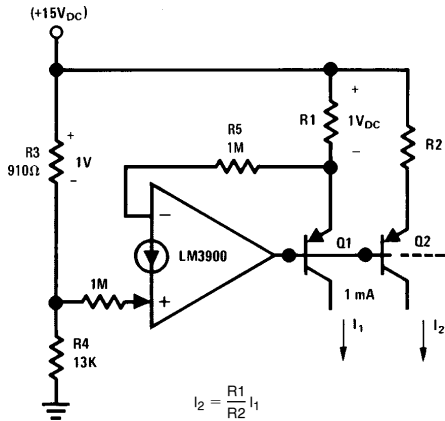
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FIGURE 71. Fixed Current Sources

The resistor, R_2, can be used to scale the collector current of Q_2 either above or below the 1 mA reference value.

10.1.2 A VOLTAGE VARIABLE CURRENT SOURCE

A voltage variable current source is shown in *Figure 72*. The transconductance is $-(1/R_2)$ as the voltage gain from the input terminal to the emitter of Q_1 is -1. For a $V_{IN} = 0$ V_{DC} the output current is essentially zero mA DC. The resistors R_1 and R_6 guarantee that the amplifier can turn OFF transistor Q_1.
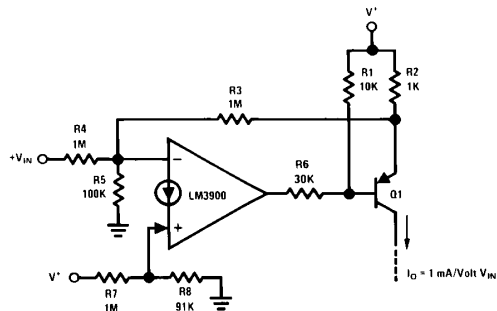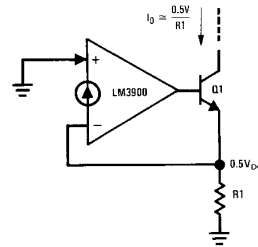

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FIGURE 72. A Voltage Controlled Current Source
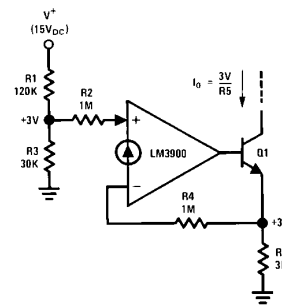
10.1.3 A FIXED CURRENT SINK

Two current sinks are shown in *Figure 73*. The circuit of *Figure 73(a)* requires only one resistor and supplies an out-

put current which is directly proportional to this R value. A negative temperature coefficient will result due to the 0.5 V_{DC} reference being the base-emitter junction voltage of the $(-)$ input transistor. If this temperature coefficient is objectionable, the circuit of *Figure 73(b)* can be employed.
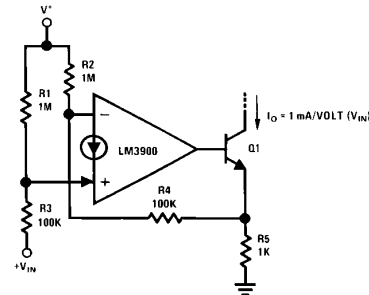

TL/H/7383–81

(a) A Simple Current Sink

TL/H/7383–82

(b) Reducing Temperature Drift Of I_O

FIGURE 73. Fixed Current Sinks

10.1.4 A VOLTAGE VARIABLE CURRENT SINK

A voltage variable current sink is shown in *Figure 74*. The output current is 1 mA per volt of V_{IN} (as $R_5 = 1$ kΩ and the gain is $+1$). This circuit provides approximately 0 mA output current for $V_{IN} = 0$ V_{DC}.
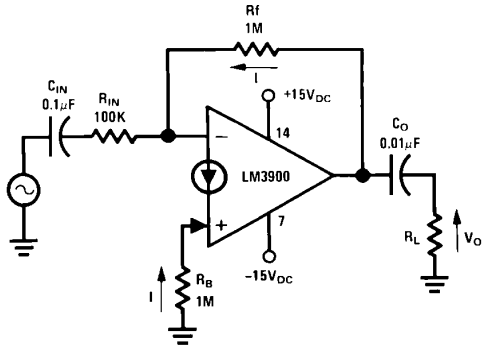

TL/H/7383–83

FIGURE 74. A Voltage Controlled Current Sink

10.2 OPERATION FROM ± 15 V_{DC} POWER SUPPLIES

If the ground pin (no. 7) is returned to a negative voltage and some changes are made in the biasing circuits, the LM3900 can be operated from ± 15 V_{DC} power supplies.

10.2.1 AN AC AMPLIFIER OPERATING WITH ± 15 V_{DC} POWER SUPPLIES

An AC coupled amplifier is shown in *Figure 75*. The biasing resistor, R_B, is now returned to ground and both inputs bias at one V_{BE} above the $-V_{EE}$ voltage (approximately -15 V_{DC}).

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FIGURE 75. An AC Amplifier Operating With ± 15 V_{DC}

With $R_f = R_B$, V_O will bias at approximately 0 V_{DC} to allow a maximum output voltage swing. As pin 7 is common to all four of the amplifiers which are in the same package, the other amplifiers are also biased for operation off of ±15 V_{DC}.

10.2.2 A DC AMPLIFIER OPERATING WITH ± 15 V_{DC} POWER SUPPLIES

Biasing a DC amplifier is more difficult and requires that the ± power supplies be complementary tracking (i.e., $|+V_{CC}| = |-V_{EE}|$). The operation of this biasing can be understood if we start by first considering the amplifier without including the feedback resistors, as shown in *Figure 76*. If $R_1 = R_2 = R_3 + R_4 = 1$ MΩ and $|+V_{CC}| = |-V_{EE}|$,
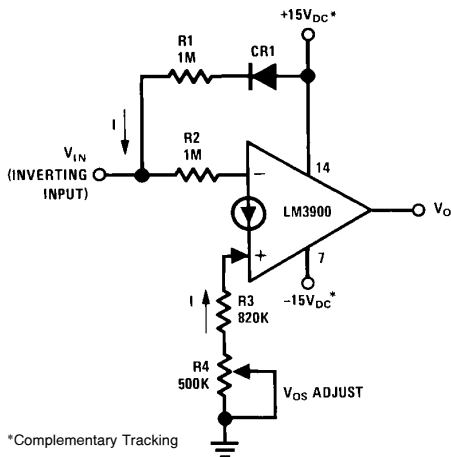

*Complementary Tracking

TL/H/7383-85

FIGURE 76. DC Biasing for ± 15 V_{DC} Operation

then the current, I, will bias V_{IN} at zero volts DC (resistor R_4 can be used to adjust this). The diode, CR_1, has been added for temperature compensation of this biasing. Now, if we include these biasing resistors, we have a DC amplifier with the input biased at approximately zero volts. If feedback resistors are added around this biased amplifier we get the schematic shown in *Figure 77*.
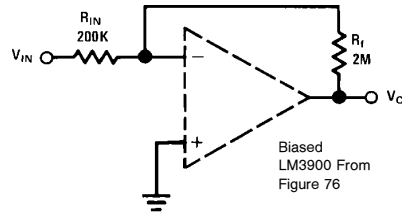
TL/H/7383-86

FIGURE 77. A DC Amplifier Operating with ± 15 V_{DC}

This is a standard inverting DC amplifier connection. The (+) input is "effectively" at ground and the biasing shown in *Figure 76* is used to take care of DC levels at the inputs.

10.3 TACHOMETERS

Many pulse averaging tachometers can be built using the LM3900. Inputs can be voltage pulses, current pulses or the differentiated transitions of squarewaves. The DC output voltage can be made to increase with increasing input frequency, can be made proportional to twice the input frequency (frequency doubling for reduced output ripple), and can also be made proportional to either the sum or the difference between two input frequencies. Due to the small bias current and the high gain of the LM3900, the transfer function is linear between the saturation states of the amplifier.

10.3.1 A BASIC TACHOMETER

If an RC averaging network is added from the output to the (−) input, the basic tachometer of *Figure 78* results. Current pulse inputs will provide the desired transfer function shown on the figure. Each input current pulse causes a small change in the output voltage. Neglecting the effects of R we have

$$\Delta V_O \cong \frac{I\Delta t}{C}$$

The inclusion of R gives a discharge path so the output voltage does not continue to integrate, but rather provides the time dependency which is necessary to average the input pulses. If an additional signal source is simply placed in parallel with the one shown, the output becomes proportional to the sum of these input frequencies. If this additional source were applied to the (−) input, the output voltage would be proportional to the difference between these input frequencies. Voltage pulses can be converted to current pulses by using an input resistor. A series isolating diode should be used if a signal is applied to the (−) input to prevent loading during the low voltage state of this input signal.
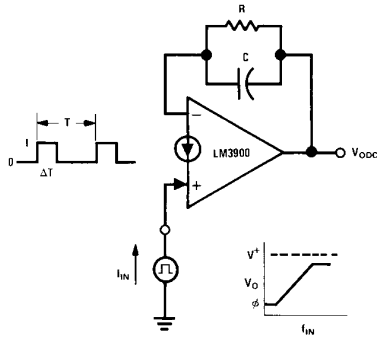
FIGURE 78. A Basic Tachometer

10.3.2 EXTENDING V_{OUT} (MINIMUM) TO GROUND

The output voltage of the circuit of *Figure 78* does not go to ground level but has a minimum value which is equal to the V_{BE} of the ($-$) input (0.5 V_{DC}). If it is desired that the output voltage go exactly to ground, the circuit of *Figure 79* can be used. Now with $V_{IN} = 0$ V_{DC}, $V_O = 0$ V_{DC} due to the addition of the common-mode biasing resistors (180 kΩ).
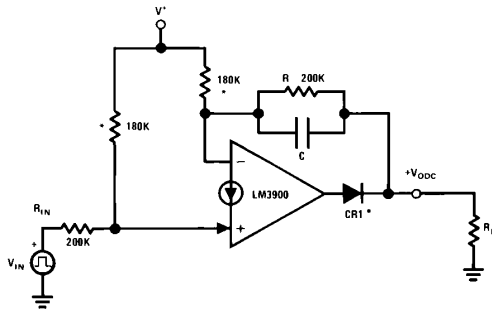

FIGURE 79. Adding Biasing to Provide $V_O = 0$ V_{DC}

The diode, CR_1, allows the output to go below $V_{CE\ SAT}$ of the output, if desired (a load is required to provide a DC path for the biasing current flow via the R of the averaging network).

10.3.3 A FREQUENCY DOUBLING TACHOMETER

To reduce the ripple on the DC output voltage, the circuit of *Figure 80* can be used to effectively double the input frequency. Input pulses are not required, a squarewave is all that is needed. The operation of the circuit is to average the charge and discharge transient currents of the input capacitor, C_{IN}. The resistor, R_{IN}, is used to convert the voltage pulses to current pulses and to limit the surge currents (to approximately 200 μA peak—or less if operating at high temperatures).

When the input voltage goes high, the charging current of C_{IN}, I_{CHG} enters the ($+$) input, is mirrored about ground and is drawn from the RC averaging network into the ($-$) input terminal. When the input voltage goes back to ground, the

discharge current of C_{IN}, $I_{DISCHARGE}$ will also be drawn from the RC averaging network via the now conducting diode, CR_1. This full wave action causes two current pulses to be drawn through the RC averaging network for each cycle of the input frequency.
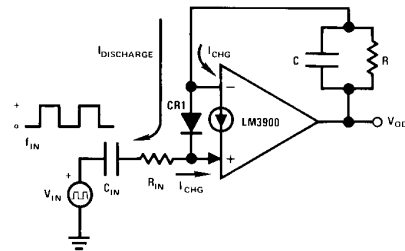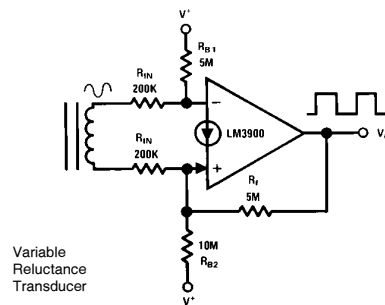

FIGURE 80. A Frequency Doubling Tachometer

10.4 A SQUARING AMPLIFIER

A squaring amplifier which incorporates symmetrical hysteresis above and below the zero output state (for noise immunity) is often needed to amplify the low level signals which are provided by variable reluctance transducers. In addition, a high frequency roll-off (low pass characteristic) is desirable both to reduce the natural voltage buildup at high frequencies and to also filter high frequency input noise disturbances. A simple circuit which accomplishes this function is shown in *Figure 81*. The input voltage is converted to


FIGURE 81. A Squaring Amplifier with Hysteresis

input currents by using the input resistors, R_{IN}. Common-mode biasing is provided by R_{B1} and R_{B2}. Finally positive feedback (hysteresis) is provided by R_f. The large source resistance, R_{IN}, provides a low pass filter due to the "Miller-effect" input capacitance of the amplifier (approximately 0.002 μF). The amount of hysteresis and the symmetry about the zero volt input are controlled by the positive feedback resistor, R_f, and R_{B1} and R_{B2}. With the values shown in *Figure 81* the trip voltages are approximately $\pm$150 mV centered about the zero output voltage state of the transducer (at low frequencies where the low pass filter is not attenuating the input signal).

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10.5 A DIFFERENTIATOR

An input differentiating capacitor can cause the input of the LM3900 to swing below ground and actuate the input clamp circuit. Again, common-mode biasing can be used to prevent this negative swing at the input terminals of the LM3900. The schematic of a differentiator circuit is shown in *Figure 82*. Common-mode biasing is provided by R_{B1} and
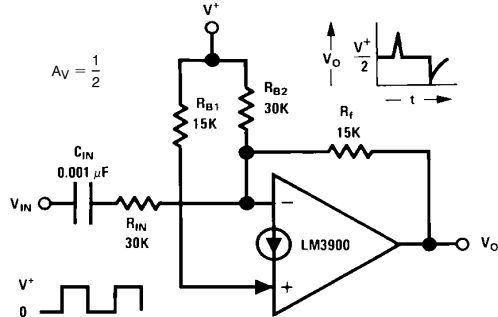
FIGURE 82. A Differentiator Circuit

R_{B2}. The feedback resistor, R_f, is one-half the value of R_{IN} so the gain is 1/2. The output voltage will bias at $V^+/2$ which thereby allows both a positive and a negative swing above and below this bias point. The resistor, R_{IN}, keeps the negative swing isolated from the $(-)$ input terminal and therefore both inputs remain biased at $+V_{BE}$.

10.6 A DIFFERENCE INTEGRATOR

A difference integrator is the basis of many of the sweep circuits which can be realized using the LM3900 operating on only a single power supply voltage. This circuit can also be used to provide the time integral of the difference between two input waveforms. The schematic of the difference integrator is shown in *Figure 83*.
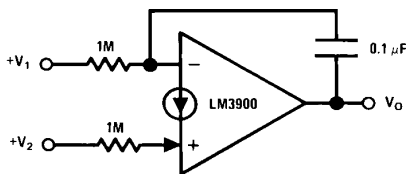
FIGURE 83. A Difference Integrator

This is a useful component for DC feedback loops as both the comparison to a reference and the integration take place in one amplifier.

10.7 A LOW DRIFT SAMPLE AND HOLD CIRCUIT

In sample and hold applications a very low input biasing current is required. This is usually achieved by using a FET transistor or a special low input current IC op amp. The existence of many matched amplifiers in the same package allows the LM3900 to provide some interesting low "equivalent" input biasing current applications.

10.7.1 REDUCING THE "EFFECTIVE" INPUT BIASING CURRENT

One amplifier can be used to bias one or more additional amplifiers as shown in *Figure 84*.
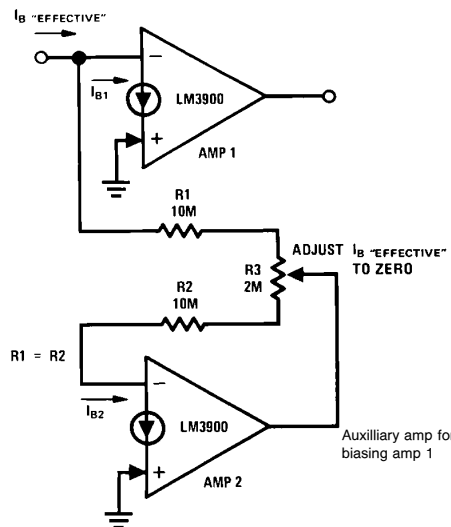
FIGURE 84. Reducing I_B "Effective" to Zero

The input terminal of Amp. 1 will only need to supply the signal current if the DC biasing current, I_{B1}, is accurately supplied via R_1. The adjustment, R_3, allows a zeroing of "I_B effective" but simply omitting R_3 and letting $R_1 = R_2$ (and relying on amplifier symmetry) can cause I_B "effective" to be less than $I_B/10$ (3 nA). This is useful in circuit applications such as sample and hold, where small values of I_B "effective" are desirable.

10.7.2 A LOW DRIFT RAMP AND HOLD CIRCUIT

The input current reduction technique of the previous section allows a relatively simple ramp and hold circuit to be built which can be ramped up or down or allowed to remain at any desired output DC level in a "hold" mode. This is shown in *Figure 85*. If both inputs are at 0 V_{DC} the circuit is in a hold mode. Raising either input will cause the DC output voltage to ramp either up or down depending on which one goes positive. The slope is a function of the magnitude of the input voltage and additional inputs can be placed in parallel, if desired, to increase the input control variables.
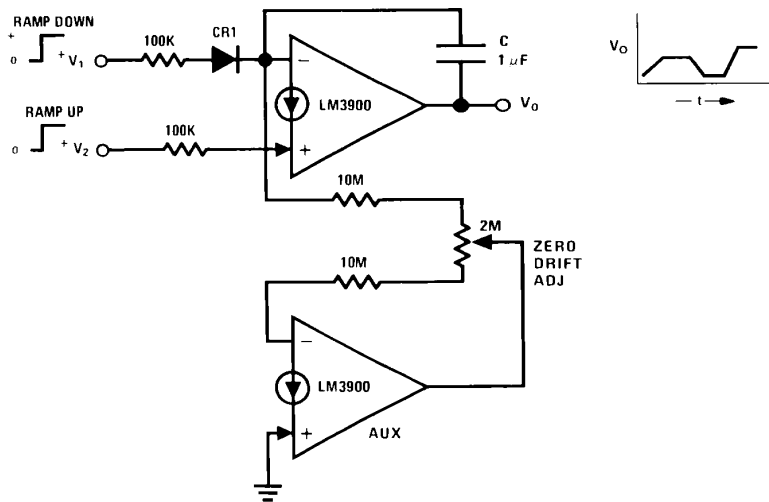
FIGURE 85. A Low-Drift Ramp and Hold Circuit

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10.7.3 SAMPLE-HOLD AND COMPARE WITH NEW $+V_{IN}$

An example of using the circuit of the previous section is shown in *Figure 86* where clamping transistors, Q_1 and Q_2, put the circuit in a hold mode when they are driven ON. When OFF the output voltage of Amp. 1 can ramp either up or down as needed to guarantee that the output voltage of Amp. 1 is equal to the DC input voltage which is applied to Amp. 3. Resistor R_1 provides a fixed "down" ramp current which is balanced or controlled via the comparator, Amp. 3, and the resistor R_4. When Q_1 and Q_2 are OFF a feedback loop guarantees that V_{01} (from Amp. 1) is equal to $+V_{IN}$ (to Amp. 3). Amplifier 2 is used to supply the input biasing current to Amp. 1.

$$V_{02} = A_{OL}\,[V_{IN}\,(+) - V_{IN(HOLD)}]$$
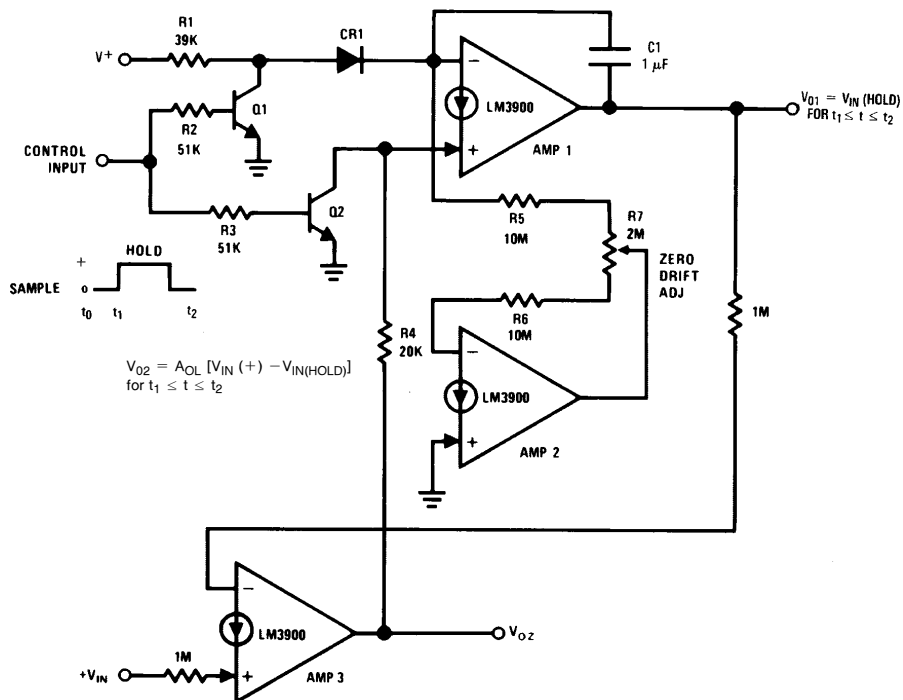for $t_1 \leq t \leq t_2$

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FIGURE 86. Sample-Hold and Compare with New $+V_{IN}$

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The stored voltage appears at the output, V_{01} of Amp. 1, and as Amp. 3 is active, a continued comparison is made between V_{01} and V_{IN} and the output of Amp. 3 fully switches based on this comparison. A second loop could force V_{IN} to be maintained at the stored value (V_{01}) by making use of V_{02} as an error signal for this second loop. Therefore, a control system could be manually controlled to bring it to a particular operating condition; then, by exercising the hold control, the system would maintain this operating condition due to the analog memory provided by V_{01}.

10.8 AUDIO MIXER OR CHANNEL SELECTOR

The multiple amplifiers of the LM3900 can be used for audio mixing (many amplifiers simultaneously providing signals which are added to generate a composite output signal) or for channel selection (only one channel enabled at a time).

Three amplifiers are shown being summed into a fourth amplifier in *Figure 87*.

If a power amplifier were available, all four amplifiers could feed the single input of the power amplifier. For audio mixing all amplifiers are simultaneously active. Particular amplifiers can be gated OFF by making use of DC control signals which are applied to the (+) inputs to provide a channel select feature. As shown on *Figure 87*, Amp. 3 is active (as sw 3 is closed) and Amps. 1 and 2 are driven to positive output voltage saturation by the 5.1M which is applied to the (+) inputs. The DC output voltage bias level of the active amplifier is approximately 0.8 V_{DC} and could be raised if larger signal levels were to be accommodated. Frequency shaping networks can be added either to the individual amplifiers or to the common amplifier, as desired. Switching transients may need to be filtered at the DC control points if the output amplifier is active during the switching intervals.
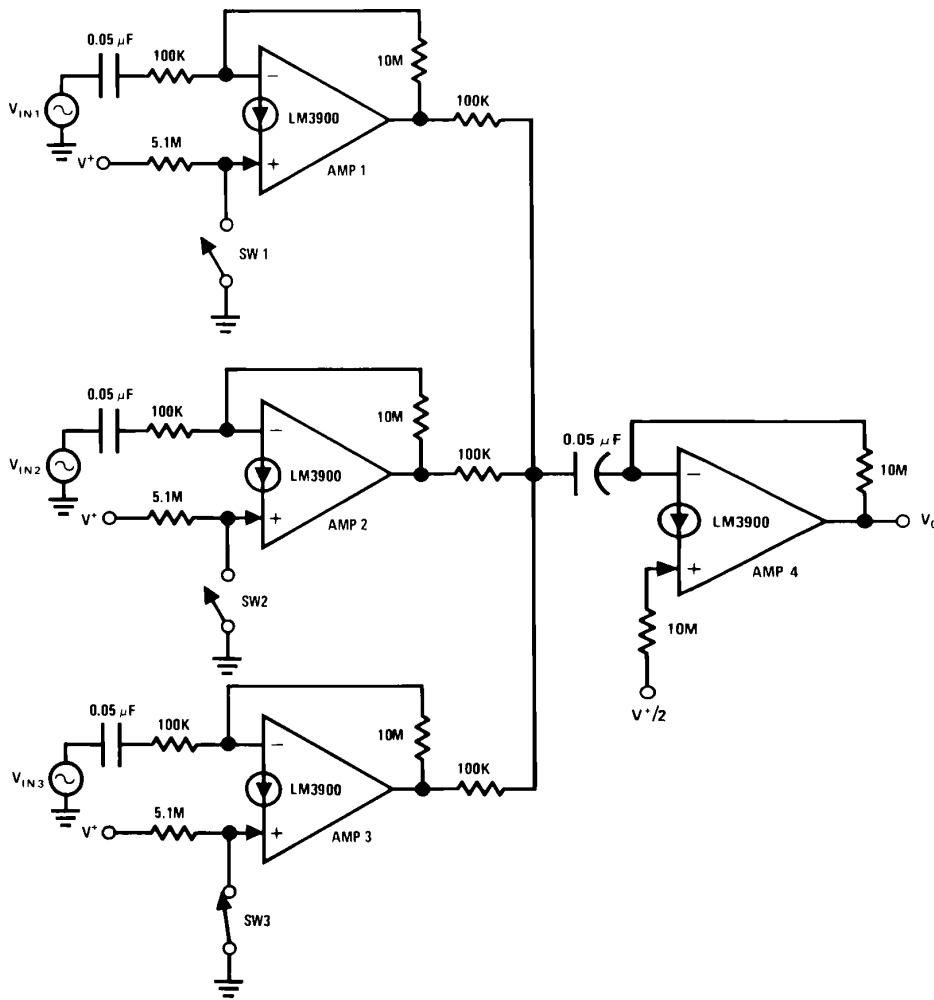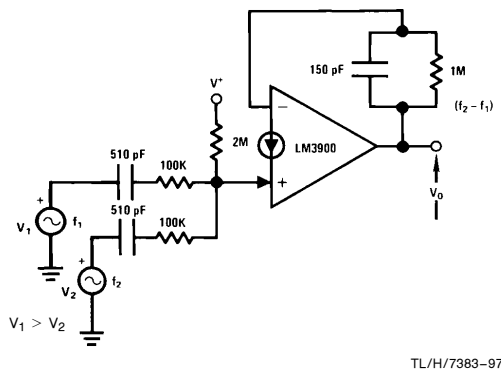

TL/H/7383–96

FIGURE 87. Audio Mixing or Selection

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10.9 A LOW FREQUENCY MIXER

The diode which exists at the (+) input can be used for non-linear signal processing. An example of this is a mixer which allows two input frequencies to produce a sum and difference frequency (in addition to other high frequency components). Using the amplifier of the LM3900, gain and filtering can also be accomplished with the same circuit in addition to the high input impedance and low output impedance advantages. The schematic of *Figure 88* shows a mixer with a gain of 10 and a low pass single pole filter (1M and 150 pF feedback elements) with a corner frequency of 1 kHz. With one signal larger in amplitude, to serve as the local oscillator input (V_1), the transconductance of the input diode is gated at this rate (f_1). A small signal (V_2) can now be added at the second input and the difference frequency is filtered from the composite resulting waveform and is made available at the output. Relatively high frequencies can be applied at the inputs as long as the desired difference frequency is within the bandwidth capabilities of the amplifier and the RC low pass filter.
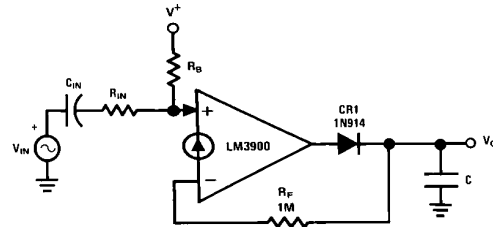
TL/H/7383–97

FIGURE 88. A Low Frequency Mixer

10.10 A PEAK DETECTOR

A peak detector is often used to rapidly charge a capacitor to the peak value of an input waveform. The voltage drop across the rectifying diode is placed within the feedback loop of an op amp to prevent voltage losses and temperature drifts in the output voltage. The LM3900 can be used as a peak detector as shown in *Figure 89*. The feedback resistor, R_f, is kept small (1 MΩ) so that the 30 nA base current will cause only a +30 mV error in V_O. This feedback resis-

tor is constantly loading C in addition to the current drawn by the circuitry which samples V_O. These loading effects must be considered when selecting a value for C.

The biasing resistor, R_B, allows a minimum DC voltage to exist across the capacitor and the input resistor, R_{IN}, can be selected to provide gain to the input signal.
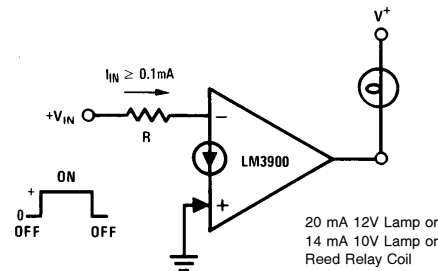
TL/H/7383–98

FIGURE 89. A Peak Detector

10.11 POWER CIRCUITS

The amplifier of the LM3900 will source a maximum current of approximately 10 mA and will sink maximum currents of approximately 80 mA (if overdriven at the (−) input). If the output is driven to a saturated state to reduce device dissipation, some interesting power circuits can be realized. These maximum values of current are typical values for the unit operating at 25°C and therefore have to be de-rated for reliable operation. For fully switched operation, amplifiers can be paralleled to increase current capability.

10.11.1 LAMP AND/OR RELAY DRIVERS (≤ 30 mA)

Low power lamps and relays (as reed relays) can be directly controlled by making use of the larger value of sink current than source current. A schematic is shown in *Figure 90* where the input resistor, R, is selected such that V_{IN} supplies at least 0.1 mA of input current.

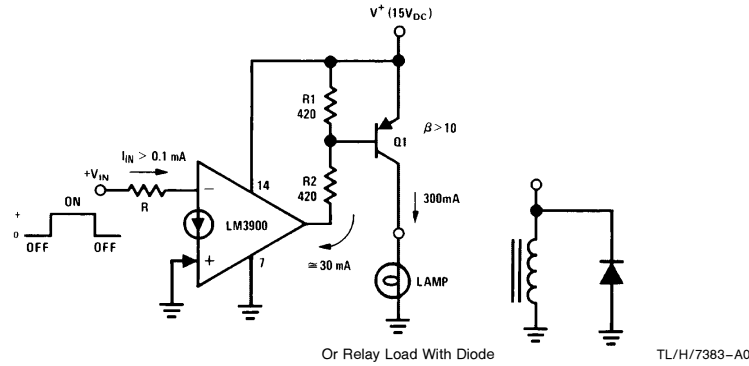TL/H/7383–99

FIGURE 90. Sinking 20 to 30 mA Loads

FIGURE 91. Boosting to 300 mA Loads

10.11.2 LAMP AND/OR RELAY DRIVERS (≤ 300 mA)

To increase the power capability, an external transistor can be added as shown in *Figure 91*. The resistors R_1 and R_2 hold Q_1 OFF when the output of the LM3900 is high. The resistor, R_2, limits the base drive when Q_1 goes ON. It is required that pin 14 tie to the same power supply as the emitter of Q_1 to guarantee that Q_1 can be held OFF. If an inductive load is used, such as a relay coil, a backswing diode should be added to prevent large inductive voltage kicks during the switching interval, ON to OFF.

10.11.3 POSITIVE FEEDBACK OSCILLATORS

If the LM3900 is biased into the active region and a resonant circuit is connected from the output to the (+) input, a positive feedback oscillator results. A driver for a piezoelectric transducer (a warning type of noise maker) is shown in *Figure 92*. The resistors R_1 and R_2 bias the output voltage at $V^+/2$ and keep the amplifier active. Large currents can be entered into the (+) input and negative currents (or currents out of this terminal) are provided by the epi-substrate diode of the IC fabrication.
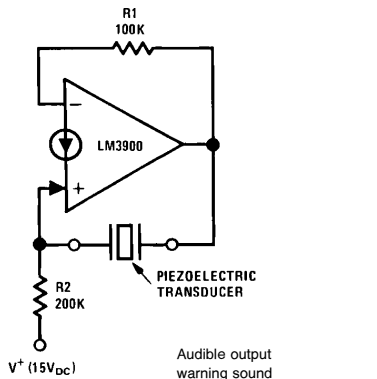

FIGURE 92. Positive Feedback Power Oscillators

When one of the amplifiers is operated in this large negative input current mode, the other amplifiers will be disturbed due to interaction. Multiple sounds may be generated as a result of using two or more transducers in various combinations, but this has not been investigated. Other two-terminal RC, RLC or piezoelectric resonators can be connected in this circuit to produce an oscillator.

10.12 HIGH VOLTAGE OPERATION

The amplifiers of the LM3900 can drive an external high voltage NPN transistor to provide a larger output voltage swing (as for an electrostatic CRT deflection system) or to operate off of an existing high voltage power supply (as the $+98 \text{ V}_{DC}$ rectified line). Examples of both types of circuits are presented in this section.

10.12.1 A HIGH VOLTAGE INVERTING AMPLIFIER

An inverting amplifier with an ouput voltage swing from essentially 0 V_{DC} to $+300$ V_{DC} is shown in *Figure 93*. The transistor, Q_1, must be a high breakdown device as it will have the full HV supply across it. The biasing resistor R_3 is used to center the transfer characteristic and the gain is the ratio of R_2 to R_1. The load resistor, R_L, can be increased, if desired, to reduce the HV current drain.
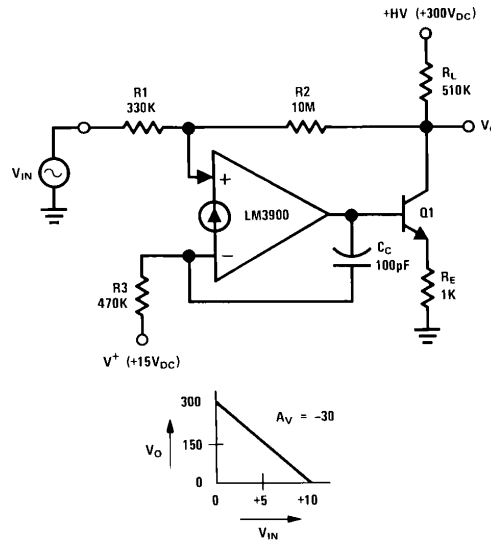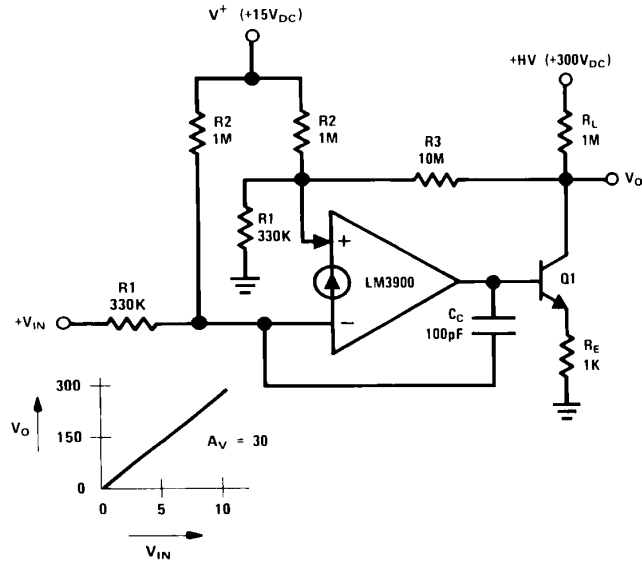

**FIGURE 93. A High Voltage Inverting Amplifier**

TL/H/7383–A3

FIGURE 94. A High Voltage Non-Inverting Amplifier

10.12.2 A HIGH VOLTAGE NON-INVERTING AMPLIFIER

A high voltage non-inverting amplifier is shown in *Figure 94*. Common-mode biasing resistors (R_2) are used to allow V_{IN} to go to 0 V_{DC}. The output voltage, V_O, will not actually go to zero due to R_E, but should go to approximately 0.3 V_{DC}. Again, the gain is 30 and a range of the input voltage of from 0 to +10 V_{DC} will cause the output voltage to range from approximately 0 to +300 V_{DC}.

10.12.3 A LINE OPERATED AUDIO AMPLIFIER

An audio amplifier which operates off a +98 V_{DC} power supply (the rectified line voltage) is often used in consumer products. The external high voltage transistor, Q_1 of *Figure 95*, is biased and controlled by the LM3900. The magnitude of the DC biasing voltage which appears across the emitter resistor of Q_1 is controlled by the resistor which is placed from the ($-$) input to ground.
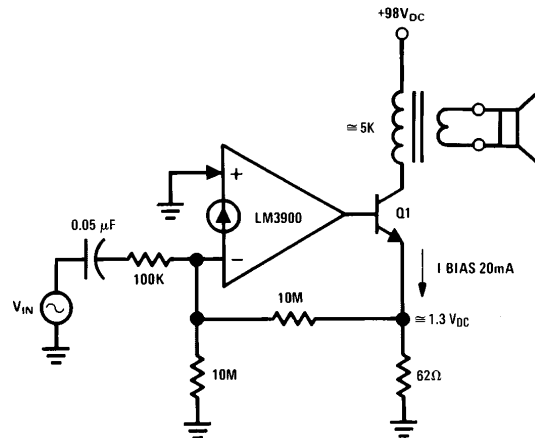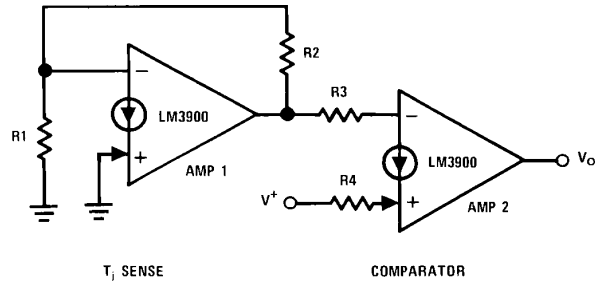

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FIGURE 95. A Line Operated Audio Amplifier
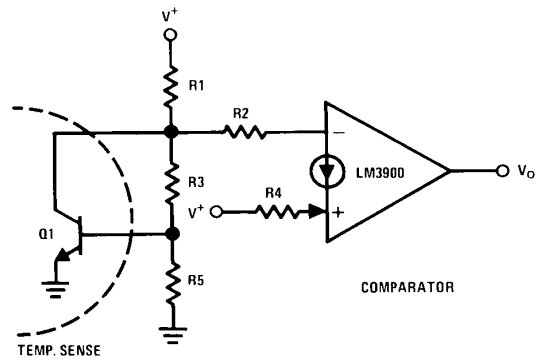
10.13 TEMPERATURE SENSING

The LM3900 can be used to monitor the junction temperature of the monolithic chip as shown in *Figure 96(a)*. Amp. 1 will generate an output voltage which can be designed to undergo a large negative temperature change by design of R_1 and R_2. The second amplifier compares this temperature dependent voltage with the power supply voltage and goes high at a designed maximum T_j of the IC.

For remote sensing, an NPN transistor, Q_1 of *Figure 96(b)*, is connected as an N V_{BE} generator (with R_3 and R_5) and biased via R_1 from the power supply voltage, V^+. The LM3900 again compares this temperature dependent voltage with the supply voltage and can be designed to have V_O go high at a maximum temperature of the remote temperature sensor, Q_1.
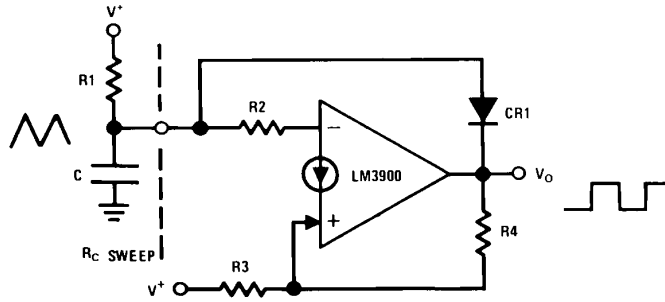

(a) IC T_j Monitor

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(b) Remote Temperature Sense

TL/H/7383–A7

FIGURE 96. Temperature Sensing

FIGURE 97. A "Programmable Unijunction"

TL/H/7383–A8

10.14 A "PROGRAMMABLE UNIJUNCTION"

If a diode is added to the Schmitt Trigger, a "programmable unijunction" function can be obtained as shown in *Figure 97*. For a low input voltage, the output voltage of the LM3900 is high and CRI is OFF. When the input voltage rises to the high trip voltage, the output falls to essentially 0V and CRI goes ON to discharge the input capacitor, C.

The low trip voltage must be larger than approximately 1V to guarantee that the forward drop of CRI added to the output voltage of the LM3900 will be less than the low trip voltage. The discharge current can be increased by using smaller values for R_2 to provide pull-down currents larger than the 1.3 mA bias current source. The trip voltages of the Schmitt Trigger are designed as shown in section 9.7.

10.15 ADDING A DIFFERENTIAL INPUT STAGE

A differential amplifier can be added to the input of the LM3900 as shown in *Figure 98*. This will increase the gain and reduce the offset voltage. Frequency compensation can be added as shown. The BV_{EBO} limit of the input transistors must not be exceeded during a large differential input condition, or diodes and input limiting resistors should be added to restrict the input voltage which is applied to the bases of Q_1 and Q_2 to $\pm V_D$.

The input common-mode voltage range does not go exactly to ground as a few tenths of a volt are needed to guarantee that Q_1 or Q_2 will not saturate and cause a phase change (and a resulting latch-up). The input currents will be small, but could be reduced further, if desired, by using FETS for Q_1 and Q_2. This circuit can also be operated off of ± 15 V_{DC} supplies.
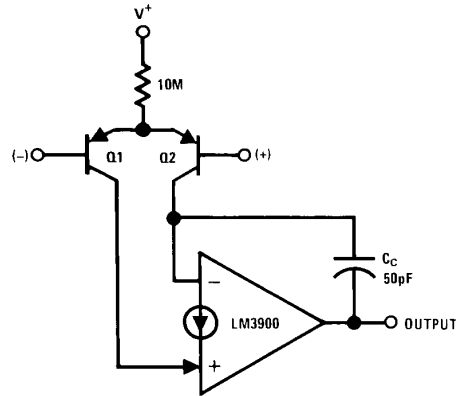
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FIGURE 98. Adding a Differential Input Stage

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