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# MC33121

## Low Voltage Subscriber Loop Interface Circuit (SLIC) Thin Film Silicon Monolithic Integrated Circuit

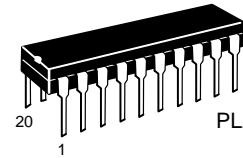
The MC33121 is designed to provide the interface between the 4-wire side of a central office, or PBX, and the 2-wire subscriber line. Interface functions include battery feed, proper loop termination AC impedance, hookswitch detection, adjustable transmit, receive, and transhybrid gains, and single/double fault indication. Additionally, the MC33121 provides a minimum of 58 dB of longitudinal balance (4-wire and 2-wire).

The transmit and receive signals are referenced to analog ground, while digital signals are referenced to digital ground, easing the interface to codecs, filters, etc. The 2 status outputs (hookswitch and faults) and the Power Down Input are TTL/CMOS compatible. The Power Down Input permits local shutdown of the circuit.

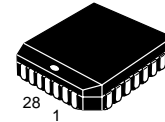
Internal drivers allow the external loop current pass transistors to be standard bipolar transistors (non-Darlington).

The MC33121 is available in a 20 pin DIP and a 28 pin PLCC surface mount package.

- 58 dB Longitudinal Balance Guaranteed; 4-wire and 2-wire
- Transmit, Receive, and Transhybrid Gains Externally Adjustable
- Return Loss Externally Adjustable
- Proper Hookswitch Detection With 30 kΩ Leakage
- Single/Double Fault Indication With Shutdown for Thermal Protection
- Critical Sense Resistors Included Internally
- Standard Power Supplies: - 21.6 V to - 42 V, and + 5.0 V, ± 10%
- On-Hook Transmission
- Power Down Input (TTL and CMOS Compatible)
- Operating Ambient Temperature: - 40°C to + 85°C
- Available in a 20 Pin DIP and 28 Pin PLCC Package



**P SUFFIX**  
PLASTIC PACKAGE  
CASE 738

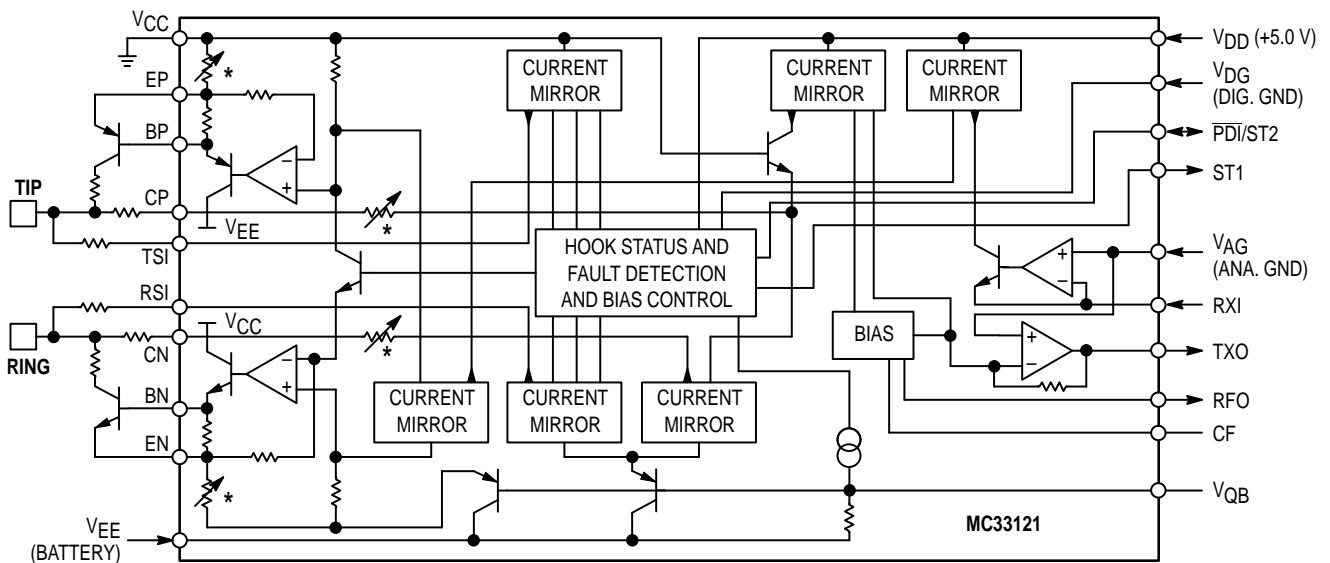


**FN SUFFIX**  
PLCC  
CASE 776

### ORDERING INFORMATION

Device	Temperature Range	Package
MC33121P	- 40° to + 85°C	Plastic DIP
MC33121FN		PLCC

### SIMPLIFIED BLOCK DIAGRAM



\* Indicates trimmed resistor

## MAXIMUM RATINGS

Characteristic	Symbol	Value	Unit
Supply Voltage (with respect to $V_{CC}$ ) (with respect to $V_{DG}$ )	$V_{EE}$ $V_{DD}$	- 60, + 0.5 - 0.5, + 7.0	Vdc
Voltage @ PDI, (with respect to $V_{DG}$ ) @ CP, CN EP, TSI BP RSI, EN BN	$V_{in}$	- 0.5, + 7.0 $V_{EE} - 0.5, V_{CC} + 0.5$ $V_{CC} - 7.0, V_{CC} + 0.5$ $V_{CC} - 14, V_{CC} + 0.5$ $V_{EE} - 0.5, V_{CC} + 14$ $V_{EE} - 1.0, V_{EE} + 21$	Vdc
Junction Temperature	$T_J$	150	°C
Storage Temperature	$T_{stg}$	- 65 to + 150	°C

\* Devices should not be operated at these limits. The "Recommended Operating Conditions" table provides for actual device operation.

## RECOMMENDED OPERATING CONDITIONS

Characteristic	Symbol	Min	Typ	Max	Unit
Supply Voltage (with respect to $V_{CC}$ ) (with respect to $V_{DG}$ )  (with respect to $V_{CC}$ ) (with respect to $V_{CC}$ ) (with respect to $V_{AG}$ )  (with respect to $V_{EE}$ ) (with respect to $V_{CC}$ and $V_{AG}$ )	$V_{EE}$ $V_{DD}$  $V_{AG}$ $V_{DG}$  $V_{DD}$	- 42 + 4.5  - 3.0 - 3.0 - 3.0  — 3.5	- 24 + 5.0  0 0 0  — —	- 21.6 + 5.5  + 10 + 7.0 + 10  47.5 —	Vdc
Loop Current	$I_L$	15	—	50	mA
PDI Input Voltage	$V_{PDI}$	0	—	$V_{DD}$	Vdc
Sink Current ST1 ST2	$I_{ST1L}$ $I_{ST2L}$	0 0	— —	1.0 1.0	mA
Transmit Signal Level at Tip & Ring Receive Signal Level at $V_{RX}$	STX SRX	- 48 - 48	— —	+ 3.0 + 3.0	dBm
Loop Resistance $V_{EE} = - 42$ V $V_{EE} = - 24$ V	$R_L$	0 0	— —	2.0 k 800	$\Omega$
External Transistor Beta	$H_{fe}$	40	—	500	A/A
Operating Ambient Temperature (See text for derating)	$T_A$	- 40	—	+ 85	°C

All limits are not necessarily functional concurrently.

## ELECTRICAL CHARACTERISTICS

( $V_{EE} = - 24$  V,  $V_{DD} = + 5.0$  V, unless otherwise noted.  $V_{CC} = V_{AG} = V_{DG} = 0$  V,  $T_A = 25^\circ\text{C}$ , see Figure 1.)

Characteristic	Symbol	Min	Typ	Max	Unit
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### POWER SUPPLIES

$V_{EE}$ Current On Hook ( $R_L > 10$ M $\Omega$ , $V_{EE} = - 42$ V) Off Hook ( $R_L = 0$ $\Omega$ , $V_{EE} = - 42$ V)*	$I_{EEN}$ $I_{EEF}$	- 2.7 - 72	- 1.0 - 55	— - 41	mA
$V_{DD}$ Current On Hook ( $R_L > 10$ M $\Omega$ , $V_{DD} = + 5.5$ V) Off Hook ( $R_L = 0$ $\Omega$ , $V_{DD} = + 5.5$ V)	$I_{DDN}$ $I_{DDF}$	— 4.0	1.4 7.0	2.7 14	
$V_{EE}$ Ripple Rejection $f = 1.0$ kHz, @ $V_{TX}$ (4-wire) $f = 1.0$ kHz, @ Tip/Ring (2-wire)	PSRR	40 40	62 52	— —	dB
$V_{DD}$ Ripple Rejection $f = 1.0$ kHz, @ $V_{TX}$ (4-wire) $f = 1.0$ kHz, @ Tip/Ring (2-wire)		37 37	52 48	— —	

\* Includes loop current.

## ELECTRICAL CHARACTERISTICS

( $V_{EE} = -24\text{ V}$ ,  $V_{DD} = +5.0\text{ V}$ , unless otherwise noted.  $V_{CC} = V_{AG} = V_{DG} = 0\text{ V}$ ,  $T_A = 25^\circ\text{C}$ , see Figure 1.)

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Characteristic	Symbol	Min	Typ	Max	Unit
<b>LOOP FUNCTIONS</b>					
Loop Current					mA
Maximum (RRF = 4.7 k, $R_L = 10\ \Omega$ )	$I_{L(max)}$	37	41	51	
Nominal (RRF = 4.7 k, $R_L = 367\ \Omega$ )	$I_L$	21	27	34	
Minimum (RRF = 4.7 k, $R_L = 796\ \Omega$ )	$I_{L(min)}$	16	17.5	—	
Battery Feed Resistance (RRF = 4.7 k, $R_L = 796\ \Omega$ )*	RBF	475	575	675	$\Omega$
Hookswitch Threshold					k $\Omega$
On-to-Off Hook	$R_{NF}$	2.0	4.1	—	
Off-to-On Hook	$R_{FN}$	—	7.7	10	
Fault Detection Threshold					$\Omega$
Ring-to-Ground ( $R_L = 367\ \Omega$ )	$R_{RG}$	600	1100	—	
Tip-to-Battery ( $R_L = 367\ \Omega$ )	$R_{TB}$	600	1100	—	

\* Calculated from  $[(24/I_{L(min)}) - 796]$

## GAIN LEVELS

Transmit Voltage Gain (CP, CN to TXO)	$G_{TX1}$	—	0.328	—	V/V
Transmit Voltage Gain ( $V_{TX}/V_L$ )	$G_{TX2}$				dB
$V_L = 0\text{ dBm}$ , $f = 1.0\text{ kHz}$		-0.3	0	0.3	
$V_L = 0\text{ dBm}$ , $f = 3.4\text{ kHz}$ , with respect to $G_{TX2}$		-0.1	0	0.1	
$V_L = +3.0\text{ dBm}$ , $f = 1.0\text{ kHz}$ , with respect to $G_{TX2}$		-0.15	0	0.15	
$V_L = -48\text{ dBm}$ , $f = 1.0\text{ kHz}$ , with respect to $G_{TX2}$		—	$\pm 0.1$	—	
Transmit Distortion (at Pin 11) ( $f = 300\text{ Hz}$ to $4.0\text{ kHz}$ , $-40\text{ dBm} \leq V_{T-R} \leq +5.0\text{ dBm}$ )	$THD_T$	—	0.05	—	%
Receive Current Gain ( $I_{EP}/I_{RXI}$ )	$G_{RX1}$	94	102	110	mA/mA
Receive Voltage Gain ( $V_L/V_{RXI}$ ) ( $R_L = 600\ \Omega$ )	$G_{RX2}$				dB
$V_{RXI} = 0\text{ dBm}$ , $f = 1.0\text{ kHz}$		-0.3	0	0.3	
$V_{RXI} = 0\text{ dBm}$ , $f = 3.4\text{ kHz}$ , with respect to $G_{RX2}$		-0.1	0	0.1	
$V_{RXI} = +3.0\text{ dBm}$ , $f = 1.0\text{ kHz}$ , with respect to $G_{RX2}$		-0.15	0	0.15	
$V_{RXI} = -48\text{ dBm}$ , $f = 1.0\text{ kHz}$ , with respect to $G_{RX2}$		—	$\pm 0.1$	—	
Receive Distortion ( $f = 300\text{ Hz}$ to $4.0\text{ kHz}$ , $-40\text{ dBm} \leq V_{RXI} \leq +5.0\text{ dBm}$ )	$THD_R$	—	0.05	—	%
Return Loss (Reference = $600\ \Omega$ resistive, $f = 1.0\text{ kHz}$ )	RL	30	> 40	—	dB
Transhybrid Rejection ( $R_L = 600\ \Omega$ resistive, $f = 1.0\text{ kHz}$ , Figure 4)	THR	—	44	—	dB

## LONGITUDINAL SIGNALS ( $V_{CM} = 1.0\text{ V}_{rms}$ , see Figures 1 and 2)

2-Wire Balance, $f = 1.0\text{ kHz}$ , $Z_{ac} = 600\ \Omega$ (@ Tip/Ring)	LB	58	64	—	dB		
4-Wire Balance, $f = 1.0\text{ kHz}$ , $Z_{ac} = 600\ \Omega$ (@ $V_{TX}$ )		58	64	—			
2-Wire Balance, $f = 330\text{ Hz}$ , $Z_{ac} = 600\ \Omega$ (@ Tip/Ring)		58	64	—			
4-Wire Balance, $f = 330\text{ Hz}$ , $Z_{ac} = 600\ \Omega$ (@ $V_{TX}$ )		58	64	—			
2-Wire Balance, $f = 3.3\text{ kHz}$ , $Z_{ac} = 600\ \Omega$ (@ Tip/Ring)		53	60	—			
4-Wire Balance, $f = 3.3\text{ kHz}$ , $Z_{ac} = 600\ \Omega$ (@ $V_{TX}$ )		53	60	—			
2-Wire Balance, $f = 1.0\text{ kHz}$ , $Z_{ac} = 900\ \Omega$ (@ Tip/Ring)	$Z_{Long}$	—	62	—	$\Omega$		
4-Wire Balance, $f = 1.0\text{ kHz}$ , $Z_{ac} = 900\ \Omega$ (@ $V_{TX}$ )		—	62	—			
Signal Balance, $f = 1.0\text{ kHz}$ (Figure 3)		40	55	—			
Longitudinal Impedance, $R_S = 9100\ \Omega$		150	180	210			
Maximum Longitudinal Current, per side $f = 1.0\text{ kHz}$ , $I_{Loop} = I_{L(min)}$ , $C_T = 0.1\ \mu\text{F}$ $V_{EE} = -42$ , $V_{CM} = 5.12\text{ V}_{rms}$		$I_{Long(max)}$	8.5	16		—	mA

## ELECTRICAL CHARACTERISTICS

( $V_{EE} = -24\text{ V}$ ,  $V_{DD} = +5.0\text{ V}$ , unless otherwise noted.  $V_{CC} = V_{AG} = V_{DG} = 0\text{ V}$ ,  $T_A = 25^\circ\text{C}$ , see Figure 1.)

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Characteristic	Symbol	Min	Typ	Max	Unit
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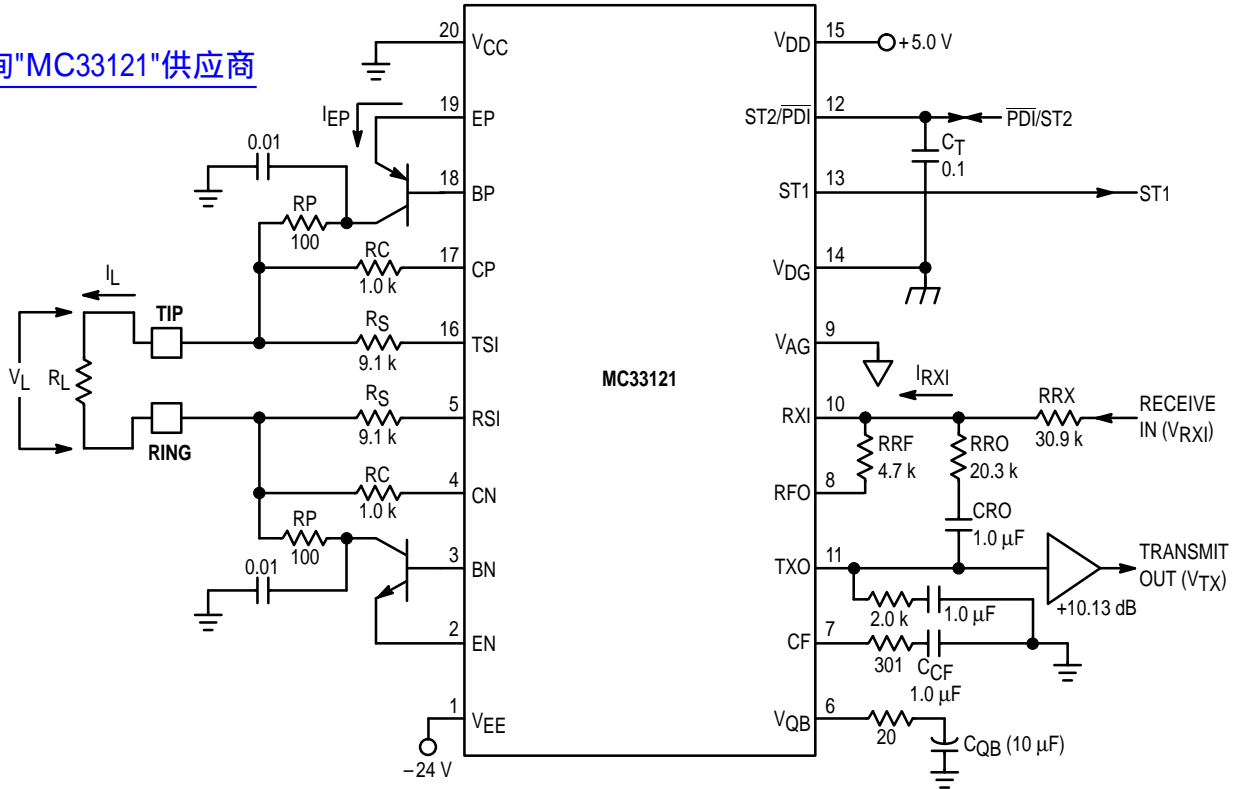
### LOGIC INTERFACE

ST1 Output Voltage Low ( $I_{ST1} = 1.0\text{ mA}$ , $V_{DD} = 5.5\text{ V}$ ) High ( $I_{ST1} = -100\text{ }\mu\text{A}$ , $V_{DD} = 4.5\text{ V}$ )	$V_{OL}$ $V_{OH}$	$V_{DG}$ 2.4	0.17 3.2	0.4 —	Vdc
ST2 Output Voltage Low ( $I_{ST2} = 1.0\text{ mA}$ , $V_{DD} = 5.5\text{ V}$ ) High ( $I_{ST2} = -100\text{ }\mu\text{A}$ , $V_{DD} = 4.5\text{ V}$ )	$V_{OL}$ $V_{OH}$	$V_{DG}$ 2.4	0.17 4.3	0.4 —	
Time Delay Hookswitch Closure to ST1 Change Hookswitch Opening to ST1 Change	$t_{ST11}$ $t_{ST12}$	— —	10 200	— —	$\mu\text{s}$
Hookswitch Closure to 90% of Loop Current ( $C_T = 0.1\text{ }\mu\text{F}$ )	$t_{HS}$	—	19	—	ms
PDI Taken High-to-Low to 10% of Loop Current PDI Taken Low-to-High to 90% of Loop Current	$t_{ST21}$ $t_{ST22}$	— —	18 10	— —	ms $\mu\text{s}$
PDI Input Current $V_{PDI} = 3.0\text{ V}$ , $R_L = 367\text{ }\Omega$ , $V_{DD} = 5.0\text{ V}$ $V_{PDI} = 0\text{ V}$ , $R_L = 367\text{ }\Omega$ , $V_{DD} = 5.5\text{ V}$	$I_{IH}$	-1250 —	-800 -800	-300 —	$\mu\text{A}$
PDI Input Voltage Low High	$V_{IL}$ $V_{IH}$	$V_{DG}$ 2.0	— —	0.8 $V_{DD}$	Vdc

### MISCELLANEOUS

$V_{QB}$ Voltage ( $V_{QB} - V_{EE}$ ) @ $I_L = 20\text{ mA}$ @ $I_L = 40\text{ mA}$	$V_{QB}$	— —	0.82 0.95	— —	Vdc
TXO Offset Voltage ( $V_{TXO} - V_{AG}$ ) @ $R_L = 600\text{ }\Omega$	$V_{TXO}$	-400	+30	+400	mVdc
TXO Output Current	$I_{TXO}$	$\pm 275$	$\pm 800$	—	$\mu\text{A pk}$
RXI Offset Voltage ( $V_{RXI} - V_{AG}$ ) @ $R_L = 600\text{ }\Omega$	$V_{RXOS}$	—	0.8	—	mVdc
$V_{AG}$ Input Current @ $R_L = 600\text{ }\Omega$	$I_{VAG}$	—	0.2	—	$\mu\text{A}$
Idle Channel Noise (with C-message filter, $R_L = 600\text{ }\Omega$ ) @ TXO (Pin 11) @ Tip/Ring	$N_{IC4}$ $N_{IC2}$	— —	-10 -5.0	— —	dBmC
Thermal Resistance — Junction to Ambient (Either package, in still air, soldered to a PC board)	$\theta_{JA}$	— —	62 36	— —	$^\circ\text{C/W}$
					(@ $T_A = +25^\circ\text{C}$ ) (@ $T_A = +85^\circ\text{C}$ )

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Components shown for a 600  $\Omega$  system.  
Three grounds are connected directly together.

Figure 1. Test Circuit

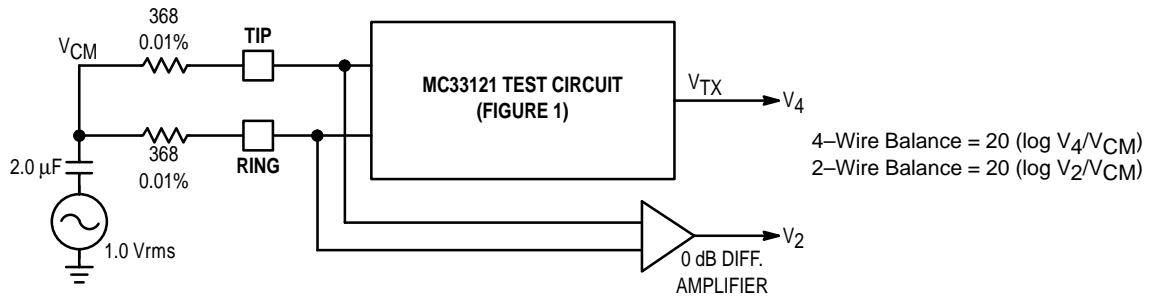


Figure 2. Longitudinal Balance Test  
(Per IEEE-455)

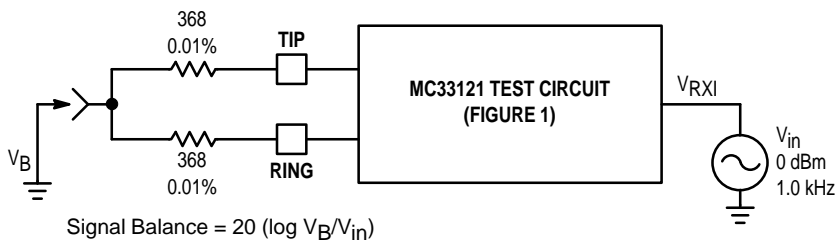
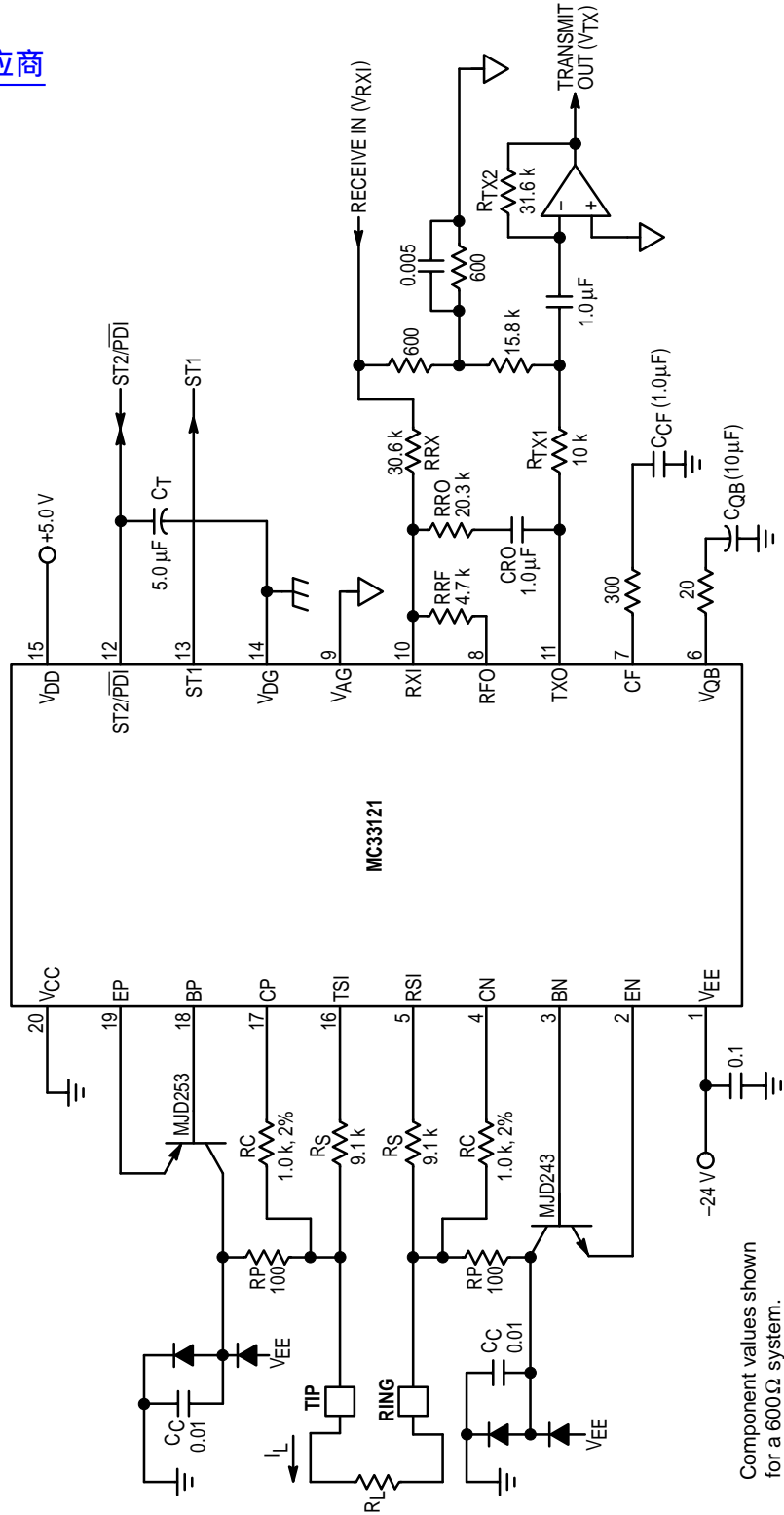


Figure 3. Signal Balance Test



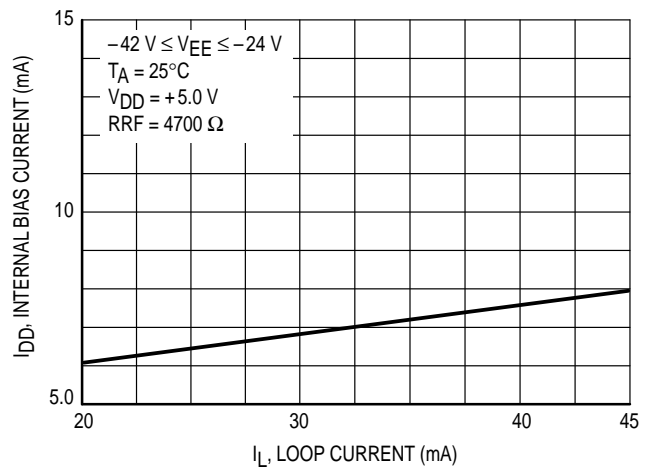
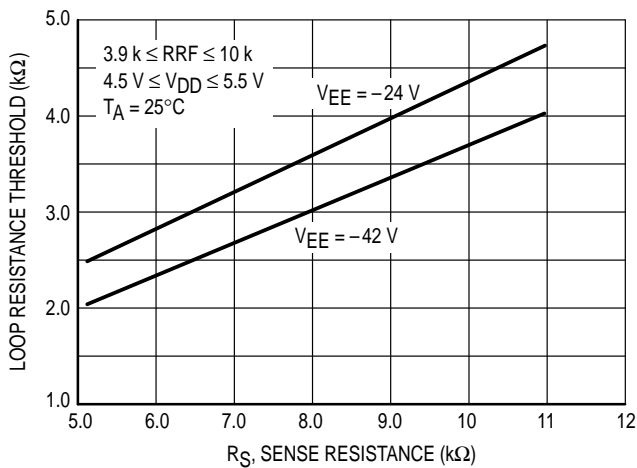
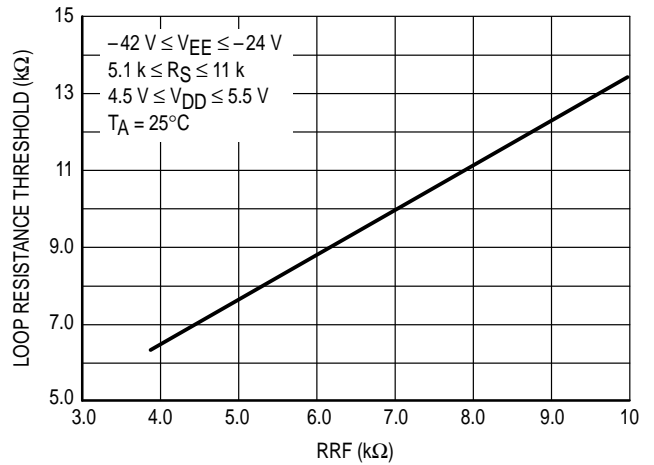
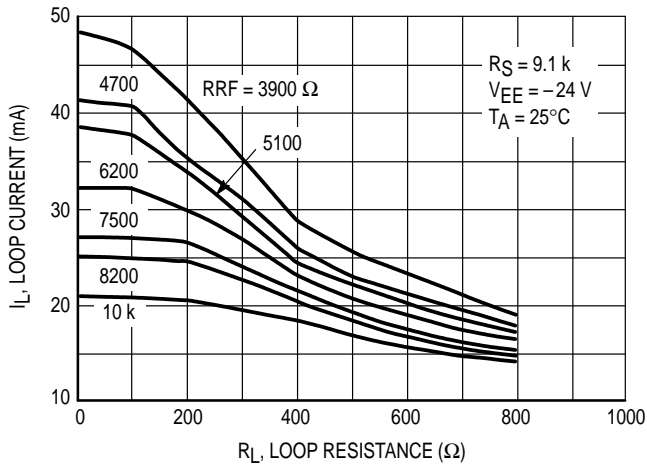
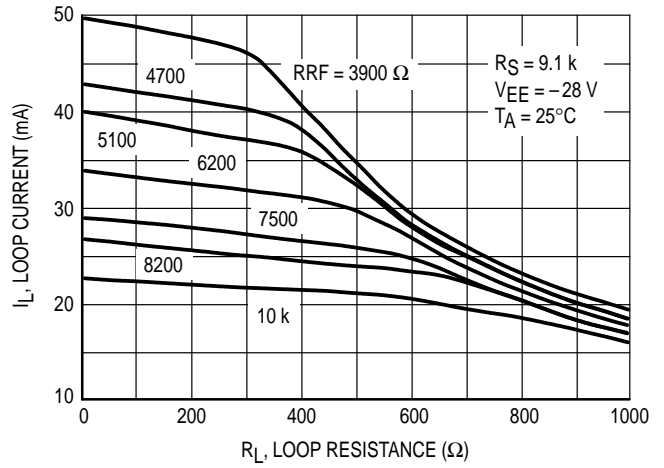
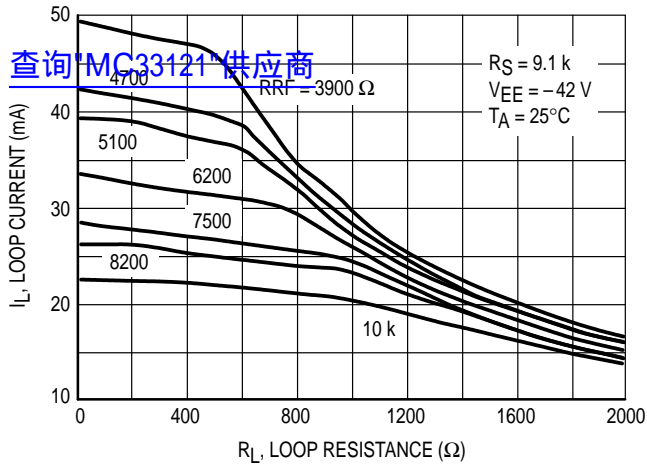
Component values shown for a 600Ω system.

Figure 4. Application Circuit

## PIN FUNCTION DESCRIPTION

Symbol	Pin		Description
	DIP	PLCC	
V <sub>CC</sub>	20	28	Connect to noise-free battery ground. Carries loop current and some bias currents.
EP	19	27	Connect to the emitter of the PNP pass transistor.
BP	18	26	Connect to the base of the PNP pass transistor.
CP	17	24	Connect to TIP through a current limiting protection resistor (R <sub>C</sub> ). CP is the noninverting input to the internal transmit amplifier (Figure 28). Input impedance is 31 kΩ.
TSI	16	23	Sense input. Connect to TIP through a current limiting protection resistor (R <sub>S</sub> ) which also sets the longitudinal impedance. Input impedance is ≈ 100 Ω to V <sub>CC</sub> .
V <sub>DD</sub>	15	22	Connect to a + 5.0 V, ± 10% supply, referenced to digital ground. Powers logic section and provides some bias currents for the loop current drivers.
V <sub>DG</sub>	14	20	Digital Ground. Reference for ST1, ST2 and V <sub>DD</sub> . Connect to system digital ground.
ST1	13	18	Status Output (TTL/CMOS). Indicates hook switch status — high when on-hook, low when off-hook, and pulse dialing information. Used with ST2 to indicate fault conditions.
ST2/PDI	12	17	Status output and an input (TTL/CMOS). As an output, ST2 can indicate hook status — Low when on-hook, high when off-hook. Used with ST1 to indicate fault conditions. As an input, it can be taken low (when off-hook) to deny subscriber loop current.
TXO	11	16	Transmit voltage output. Amplitude is ≈ 1/3 that across CP and CN. Nominally capable of 800 μA output current. DC referenced to V <sub>AG</sub> .
RXI	10	14	Receive current input. Current at this pin is multiplied by 102 at EP and EN to generate loop current. RXI is a virtual ground at V <sub>AG</sub> level. Current flow is out of this pin.
V <sub>AG</sub>	9	13	Analog ground, reference for TXO and RXI. Connect to system analog ground. Current flow is into this pin.
RFO	8	12	A resistor from this pin and RXI sets the maximum loop current and DC feed resistance. Minimum resistor value is 3.3 k (see Figures 5 to 7).
CF	7	10	A low leakage capacitor between this pin and V <sub>AG</sub> provides DC and AC signal separation. A series resistor is required for battery supply turn-on/off transient protection (Figure 4).
V <sub>QB</sub>	6	8	Quiet Battery. A capacitor between V <sub>QB</sub> and V <sub>CC</sub> filters noise and ripple from V <sub>EE</sub> , providing a quiet battery source for the speech amplifiers. A series resistor is required for battery supply turn-on/off transient protection (Figure 4).
RSI	5	7	Sense input. Connect to RING through a current limiting protection resistor which also sets the longitudinal impedance. Input impedance is ≈ 100 Ω to V <sub>QB</sub> .
CN	4	6	Connect to RING through a current limiting protection resistor. CN is the inverting input to the internal transmit amplifier (Figure 28). Input impedance is 31 kΩ.
BN	3	4	Connect to the base of the NPN pass transistor.
EN	2	3	Connect to the emitter of the NPN pass transistor.
V <sub>EE</sub>	1	2	Connect to battery voltage (– 21.6 V to – 42 V).

(Pins 1, 5, 9, 11, 15, 19, 21, and 25 are not internally connected on the PLCC package.)





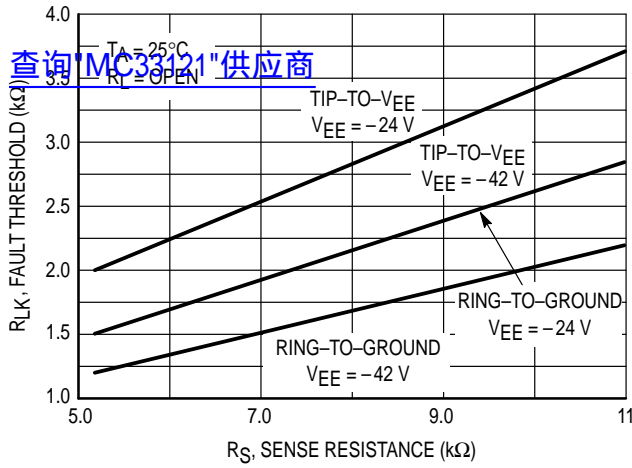


Figure 11. Fault Threshold (On-Hook) versus  $R_S$

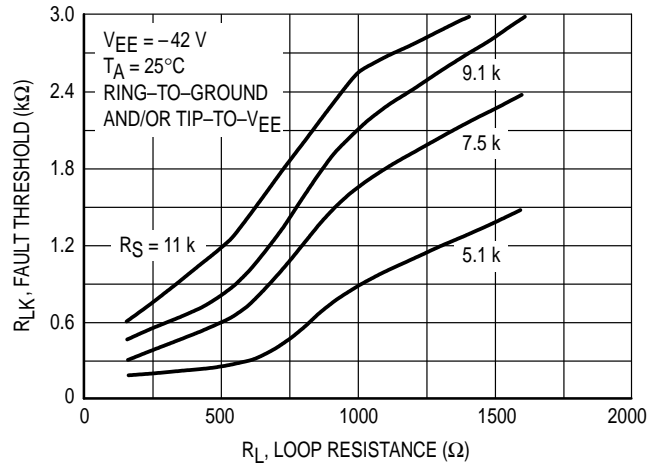


Figure 12. Fault Threshold (Off-Hook) versus Loop Resistance

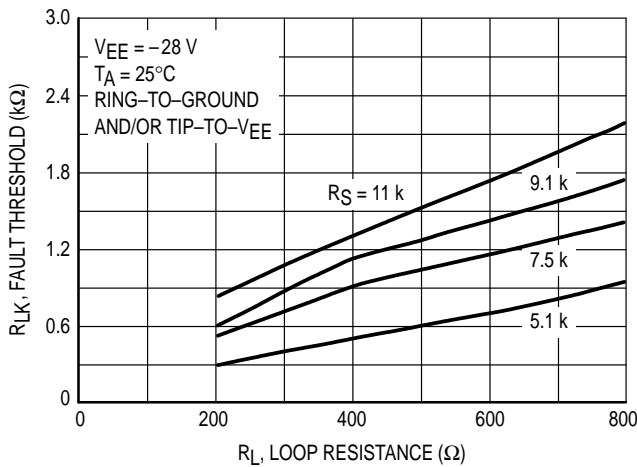


Figure 13. Fault Threshold (Off-Hook) versus Loop Resistance

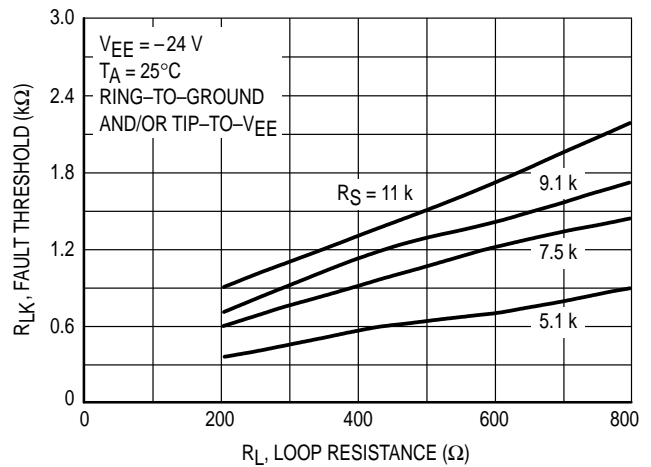


Figure 14. Fault Threshold (Off-Hook) versus Loop Resistance

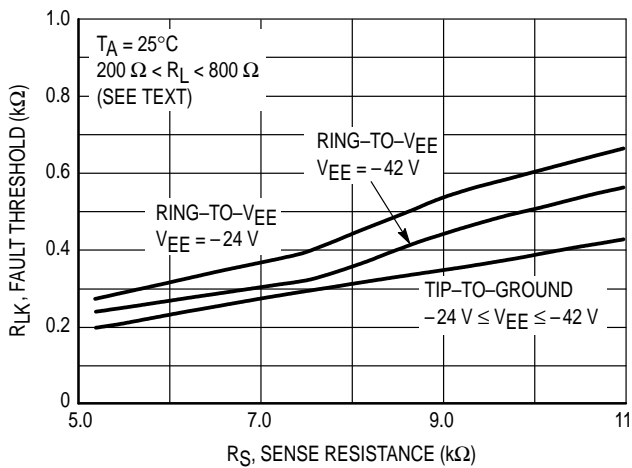


Figure 15. Fault Threshold (Off-Hook) versus  $R_S$

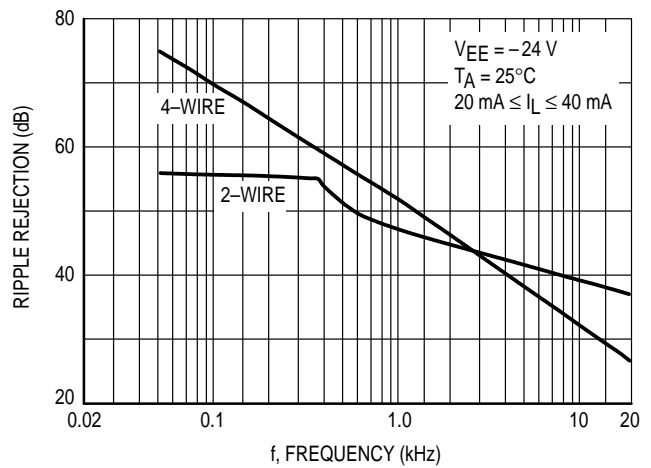


Figure 16.  $V_{DD}$  Ripple Rejection versus Frequency

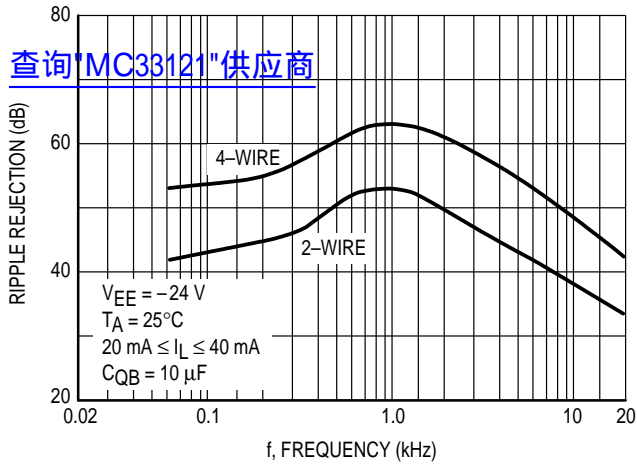


Figure 17.  $V_{EE}$  Ripple Rejection versus Frequency

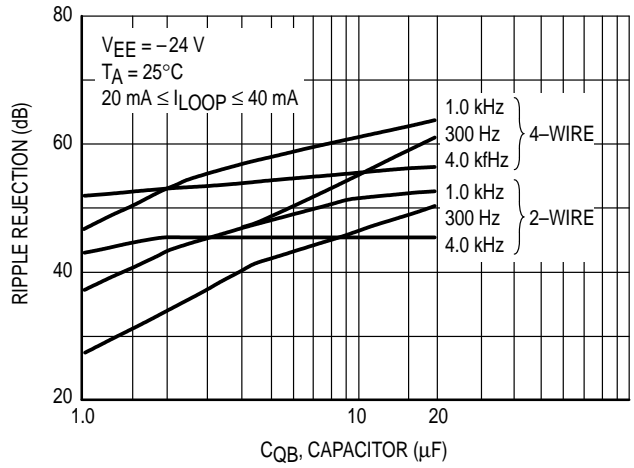


Figure 18.  $V_{EE}$  Ripple Rejection versus Frequency and  $C_{QB}$

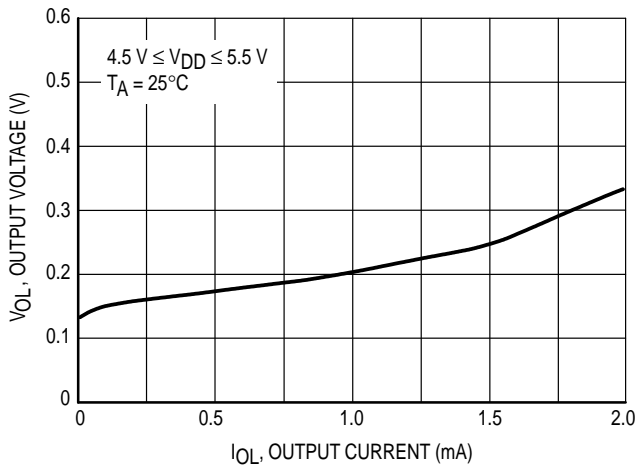


Figure 19. ST1,  $V_{OL}$  versus  $I_{OL}$

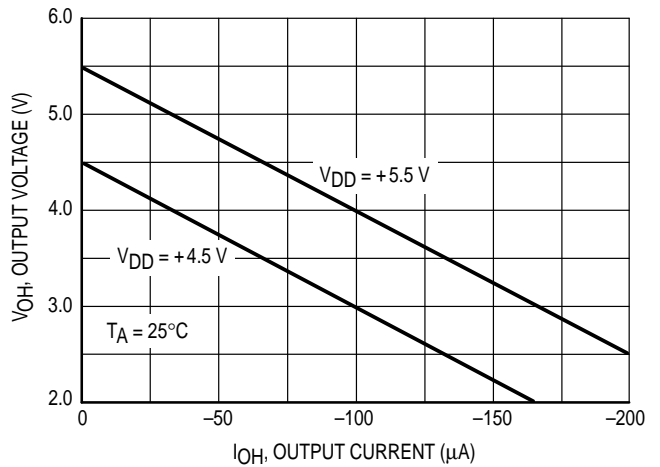


Figure 20. ST1,  $V_{OH}$  versus  $I_{OH}$

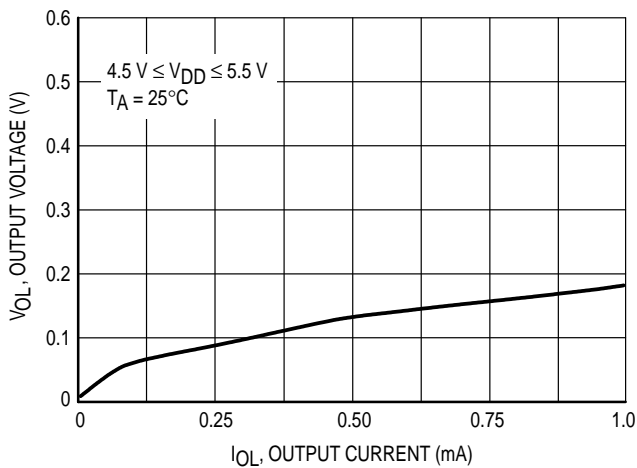


Figure 21. ST2,  $V_{OL}$  versus  $I_{OL}$

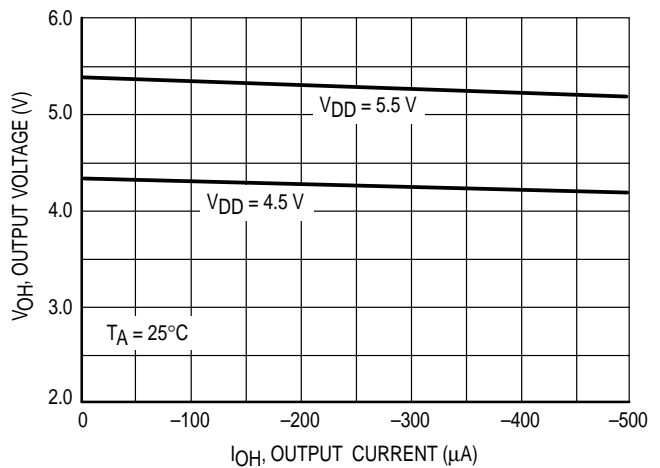


Figure 22. ST2,  $V_{OH}$  versus  $I_{OH}$

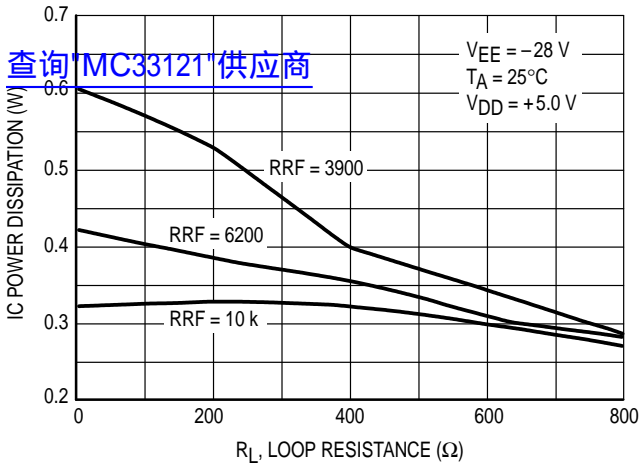


Figure 23. IC Power Dissipation versus Loop Resistance and RRF

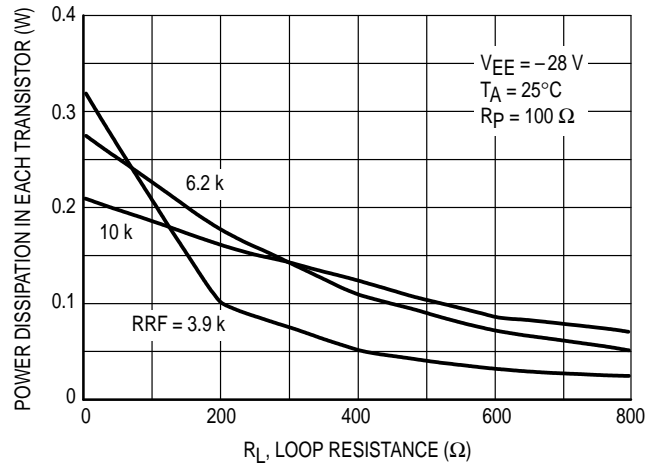


Figure 24. Transistor Power Dissipation versus Loop Resistance and RRF

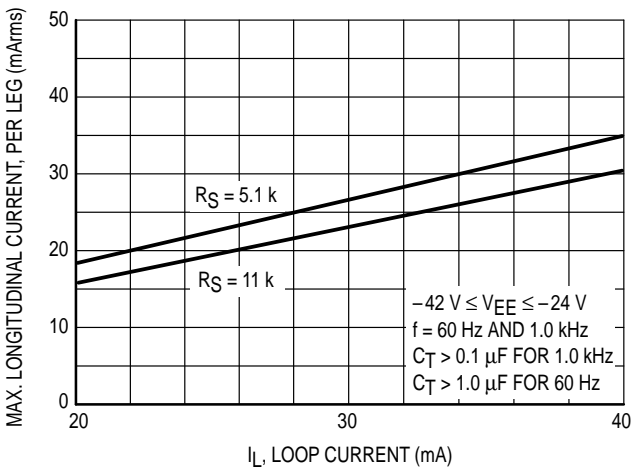


Figure 25. Maximum Longitudinal Current versus Loop Current

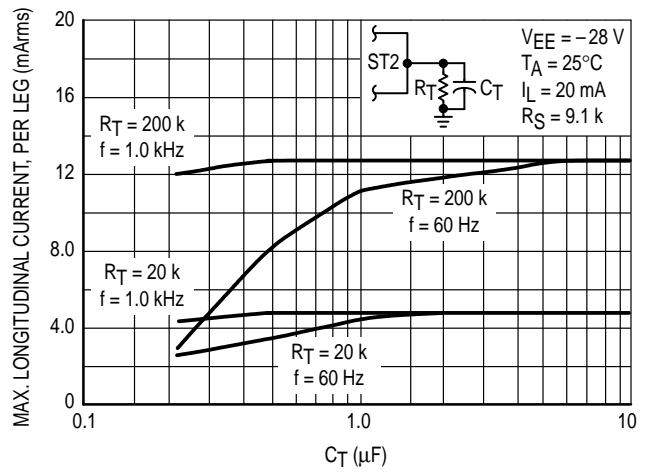


Figure 26. Maximum Longitudinal Current versus  $C_T$ ,  $R_T$  and Frequency

## FUNCTIONAL DESCRIPTION

### Introduction

The MC33121 is a solid state SLIC (Subscriber Line Interface Circuit) which provides the interface between the two wire telephone line and the four wire side of a Central Office or PBX. Most of the BORSCHT functions are provided, specifically:

- **Battery feed** of the loop current to the line, with programmable maximum current for short lines and battery feed resistance for long lines.
- **Oversvoltage protection** through internal clamp diodes and external resistors and diodes.
- **Supervision**, in that hook status is indicated in the presence of  $\geq 30\text{ k}\Omega$  leakage, and regardless of whether or not the circuit is powered down intentionally by the Central Office or PBX. Fault conditions are detected and indicated to the system. Dialing (pulse and DTMF) information is passed through the MC33121 to the 4-wire side.
- **Hybrid function**, in that the MC33121 is a 2-to-4-wire converter. Transmit, receive, return loss, and transhybrid gains are independently adjustable.

The MC33121 does not provide ring insertion, ring trip, digital coding/decoding of the speech signals, nor test functions. These must be provided external to this device.

The MC33121 controls two external transistors (one NPN and one PNP) through which the loop current flows. By appropriate circuit design, the power dissipation (which can

exceed 3.0 W under certain worst case conditions) is approximately equally distributed among the two transistors and the IC, thereby lowering junction temperatures and increasing long term reliability. In most situations, heatsinks will not be required.

The MC33121 incorporates critical sense resistors internally, which are trimmed for optimum performance. With this technique, the external resistors on the two wire side, which generally must be high wattage for transient protection reasons, can be non-precision.

Longitudinal balance is tested to a minimum of 58 dB @ 1.0 kHz (refer to Electrical Characteristics and Figure 2) for both the two-wire and four-wire side, and typically measures in the mid-60s. The longitudinal current capability is tested to a minimum of 8.5 mArms per side (refer to Electrical Characteristics and Figure 2) at a loop current of 20 mA.

Following is a description of the individual sections. Figure 4 is the reference schematic.

### DC Loop Current

The DC loop current is determined by the battery voltage ( $V_{EE}$ ), the load resistance across Tip and Ring, and the resistor at RFO. Varying the 4 resistors  $R_S$  and  $R_C$  will influence the loop current a small amount ( $< 5\%$ ). The curves of Figures 5 to 7 indicate the loop current versus loop resistance, different values of  $R_{RF}$ , and for various values of  $V_{EE}$ . The graphs represent performance at  $T_A = 25^\circ\text{C}$  and after the IC had reached a steady state temperature ( $> 5$  minutes).

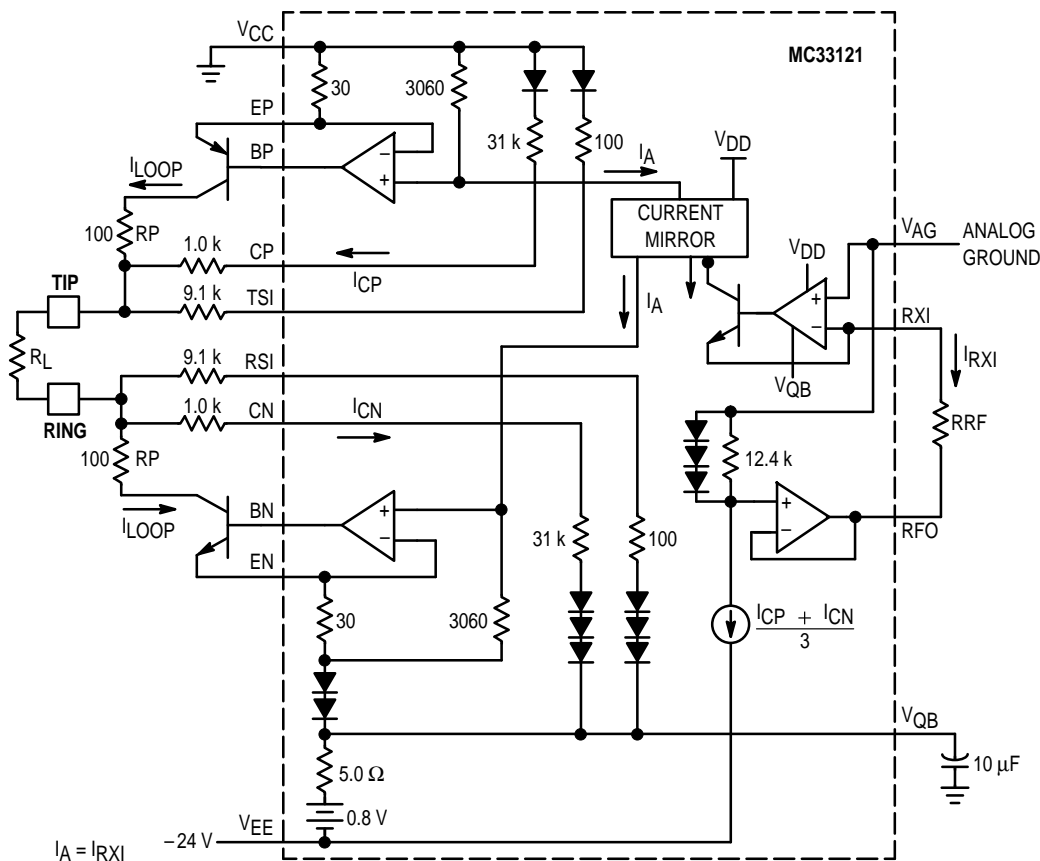


Figure 27. DC Loop Current Path

Figure 27 is representative of the DC loop current path (bold lines). On a long line ( $R_L > 400 \Omega$ ), the loop current can be determined from the following equation:

$$I_L = \frac{(|V_{EE}| - 3.6 \text{ V}) \cdot 13}{RRF + \{(R_L + 5) \cdot 13\}} \quad (1)$$

On short lines ( $R_L < 400 \Omega$ ), the three diodes across the 12.4 k resistor clamp the voltage at RFO, thereby preventing the RXI current from increasing as the load resistance is decreased. The maximum loop current is:

$$I_{L(\max)} = \frac{1.85 \text{ V} \cdot 102}{RRF} \quad (T_A = 25^\circ\text{C}) \quad (2)$$

Due to the temperature dependence of a diode's forward voltage, the maximum loop current will change with temperature by  $\approx -0.3\%/^\circ\text{C}$ .

The battery feed resistance ( $\Delta V_{TIP}/\Delta I_L$ ) is  $\approx 400 \Omega$ , but depends on the loop current,  $V_{EE}$ , RRF, and is a valid parameter only on long lines where the current limit is not in effect. On short lines, the feed resistance is high since the loop current is clamped at a near constant level. The AC impedance (Return Loss) however, is not determined nor affected by the DC parameters. See the Applications Section for Return Loss information.

### Transmit Path

The transmit path, shown in Figure 28, consists of an internal amplifier which has inputs at CP and CN, and its output at TXO. The gain is internally fixed at 0.328 V/V ( $-9.7 \text{ dB}$ ). The output is in phase with the signal at CP (normally the same as TIP), and is out of phase with the signal at CN. The signal at TXO is also out of phase with that at  $V_{RX}$ , the receive signal input, described in another section.

The TXO output can swing  $\approx 3.0 \text{ V}_p$ , with a nominal current capability of  $\pm 800 \mu\text{A}$  peak ( $\pm 275 \mu\text{A}$  minimum). The load on TXO is the parallel combination of RTX1 and the RRO network (described later). TXO is nominally internally biased at the  $V_{AG}$  DC level, but has an offset which varies with loop current.

In normal applications, the signal at CP/CN is reduced slightly from that at Tip/Ring by the voltage divider composed of the external RC resistors, and the internal 31 k resistors. The value of the RC resistors depends on the transient

protection needed, described in another section, with  $1.0 \text{ k}\Omega$  resistors being suitable for most applications. The resulting signal at TXO needs to be gained up to obtain 0 dB from Tip/Ring to  $V_{TX}$  (the 4-wire output). The common method involves an external op amp, as shown in Figure 28, with a gain of  $RTX2/RTX1$ . The gain from  $V_L$  to  $V_{TX}$  is:

$$\frac{V_{TX}}{V_L} = \frac{RTX2 \cdot 31 \text{ k} \cdot 0.328}{RTX1 \cdot (RC + 31 \text{ k})} \quad (3)$$

If a codec/filter is used, many of which include an internal op amp, a separate op amp is not needed. CTX is primarily for DC blocking (of the TXO offset), and is usually large ( $1.0 \mu\text{F}$ ) so as to not affect the gain.

### Receive Path

The receive path, shown in Figure 29, consists of the input at RXI, the transistor driver amplifiers, the external transistors, and the load at Tip/Ring.

RXI is a virtual ground (DC level =  $V_{AG}$ ) and is a current input. Current flow is out of the pin. The RXI current is mirrored to the two transistor drivers which provide a gain of 102. The two external transistors are then two current sources, in series, operating at the same value. An additional internal circuit (not shown) balances the two current sources to maintain operation in their linear region.

The load current (through  $R_L$ ) is slightly different from the transistor current due to the sense resistors RC and RS. The sense resistors add to the DC loop current, but subtract from the AC load current.

In normal operation, the current at RXI is composed of a DC current (from RFO), an AC current (from  $V_{RX}$ ) which is the receive signal, and an AC current from TXO, which is the feedback signal to set the return loss (setting the return loss is discussed in the section on AC Terminating Impedance). The resulting AC signal at Tip is inverted from that at  $V_{RX}$ , while the signal at Ring is in phase with  $V_{RX}$ .

The resistors RP are for transient protection, and their value (defined in another section) depends on the amount of protection required. A nominal value of  $100 \Omega$  is suitable for most applications.

The system receive gain, from  $V_{RX}$  to Tip/Ring, is not described in this section since in normal applications, it involves the feedback which sets the AC terminating impedance. The Applications Section discusses these in detail.

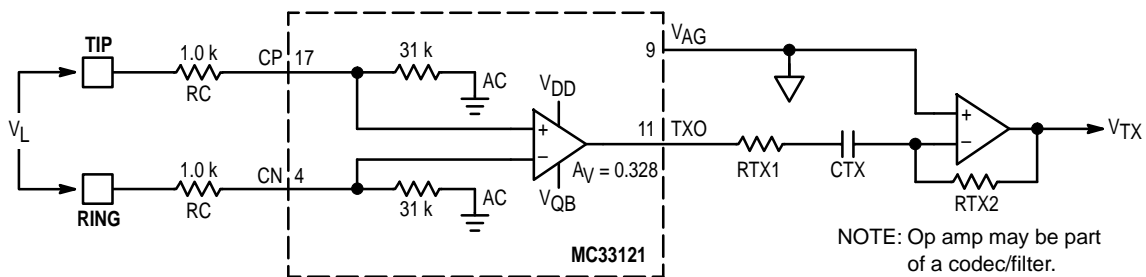


Figure 28. Transmit Path

**Logic Interface (Hook status, pulse dialing, faults)**

Figure 30 provides hookswitch status, fault information, and pulse dialing information to the 4-side of the system at the ST1 and ST2 outputs. Figure 30 is a representative diagram.

The logic outputs operate according to the truth table in Table 1:

**Table 1. Status Output Truth Table**

Hook Status	Fault Detection	Outputs		Circuit Condition
		ST1	ST2	
On-Hook	No Fault	Hi	Lo	Internally powered down
Off-Hook	No Fault	Lo	Hi	Powered up
On-Hook	Fault	Lo	Lo	Internally powered down
Off-Hook	Fault	Lo	Lo	Internally powered down

Referring to Figure 30, ST1 is configured as an active NPN pull-down with a 15 kΩ pullup resistor. ST2 has a 800 μA

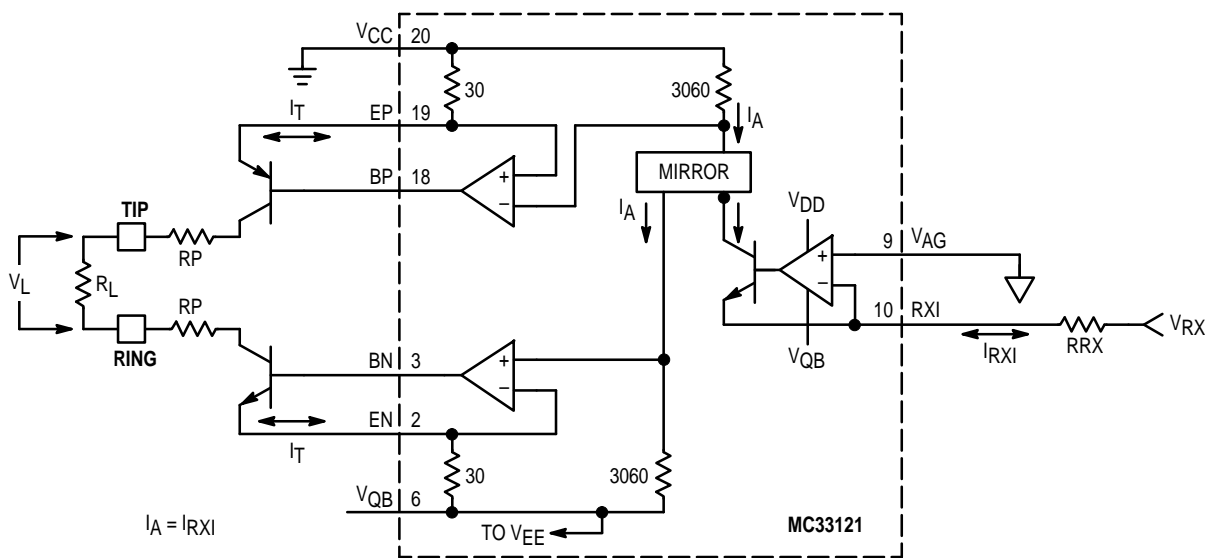
current source pullup, and a 1.0 mA current source for a pull-down. Current limiting this output controls the discharge from the external capacitor when ST2 switches low.

The condition where both ST1 and ST2 are high is not valid, but may occur momentarily during an off-hook to on-hook transition. The condition where both ST1 and ST2 are low may occur momentarily during an on-hook to off-hook transition — this should not be interpreted as a fault condition. ST1 and ST2 are TTL/CMOS compatible and are powered by the + 5.0 V supply (V<sub>DD</sub>). Refer to the Applications Section for more details.

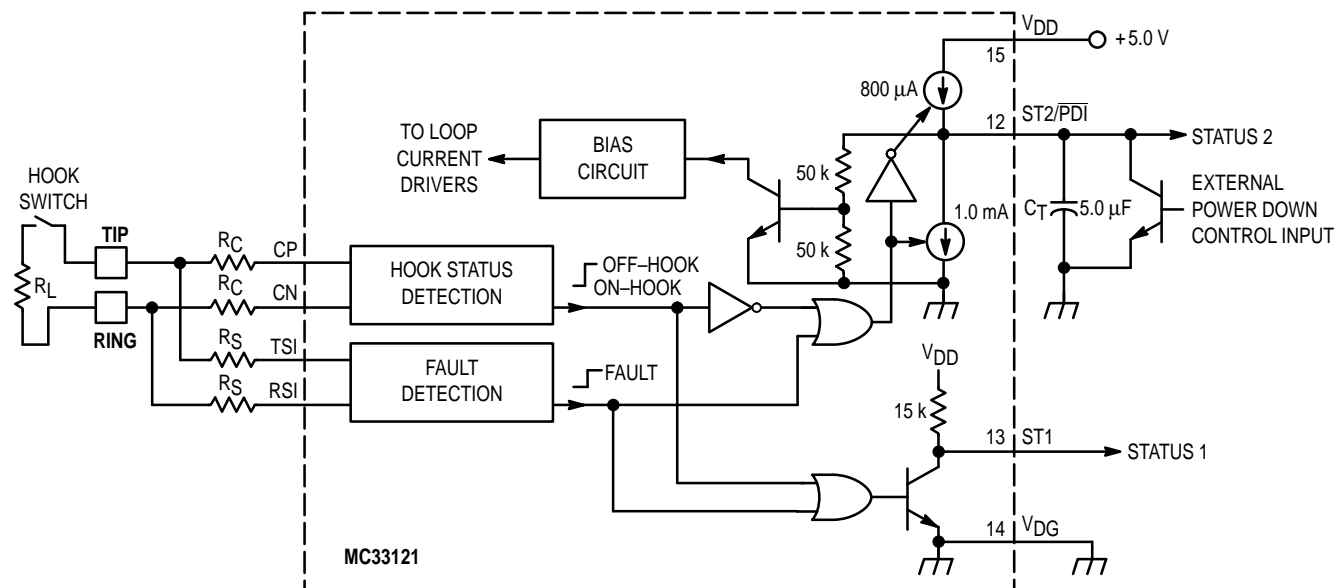
**Power Supplies, Grounds**

The MC33121 requires 2 power supplies: battery voltage between - 21.6 V and - 42 V (V<sub>EE</sub>), and an auxiliary voltage between + 4.5 V and + 5.5 V (V<sub>DD</sub>).

V<sub>EE</sub> is nominally - 24 V, with a typical range of - 21.6 V to - 42 V, and must be referenced to V<sub>CC</sub> (battery ground). A 0.1 μF bypass capacitor should be provided between V<sub>CC</sub>



**Figure 29. Receive Path**



**Figure 30. Logic Interface**

and VEE. The VEE current (IEE) is nominally 1.0 mA when on-hook, 8.0 to 12 mA more than the loop current when off-hook, and ~5.0 mA when off-hook but powered down by using the PDI pin. Ripple and noise rejection from VEE is a minimum of 40 dB (with a 10 µF capacitor at VQB), and is dependent on the size and quality of the VQB capacitor (CQB) since VQB is the actual internal supply voltage for the speech amplifiers. The absolute maximum for VEE is -60 V, and should not be exceeded by the combination of the battery voltage, its tolerance, and its ripple.

VDD is normally supplied from the line card's digital +5.0 V supply, and is referenced to VDG (digital ground). A 0.1 µF capacitor should be provided between VDD and VDG. The VDD current (IDD) is nominally 1.7 mA when on-hook and between 6.0 and 8.0 mA when off-hook (see Figure 10). When the MC33121 is intentionally powered down using the PDI pin, IDD changes by < 1.0 mA from the normal off-hook value.

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VAG is the analog ground for the MC33121, and is the reference for the speech signals (RXI and TXO). Current flow is into the pin, and is typically < 0.5 µA.

Normally, VCC, VDG and VAG are to be at the same DC level. However, if strong transients are expected at Tip and Ring, as in a Central Office application, or any application where the phone line is outdoors, VCC should not be connected directly to VDG and VAG in order to prevent possible damage to the +5.0 V system. The MC33121 is designed to tolerate as much as ±30 V between VCC and the other two grounds on a transient basis only. This feature permits VCC and the other grounds to be kept separate (on an AC basis) on the line card by transient suppressors, or to be connected together farther into the system (at the power supplies). See the Applications Section on ground arrangements and transient protection for further information on connecting the MC33121 to the system supplies.

## APPLICATIONS INFORMATION

This section contains information on the following topics:

<b>Design Procedure</b> .....	pg. 15
<b>Power Dissipation Calculations and Considerations</b> .....	pg. 22
<b>Selecting the Transistors</b> .....	pg. 23
<b>Longitudinal Current Capability</b> .....	pg. 23
<b>PC Board Layout Considerations</b> .....	pg. 23
<b>Alternate Circuit Configurations</b> .....	pg. 26

### Design Procedure

This section describes the step-by-step sequence for designing in the MC33121 SLIC into a typical line card application for either a PBX or Central Office. The sequence is important so that each new component value which is calculated does not affect components previously determined. Figure 4 (Typical Application Circuit) is the reference circuit for most of this discussion. The recommended sequence (detailed below), consists of establishing the DC aspects first, and then the AC aspects:

- 1) Determine the maximum loop current for the shortest line, select RRF. Power dissipation must be considered here.
- 2) Select the main protection resistors (RP), and diodes, based on the expected transient voltages. Transient protection configuration must also be considered here.
- 3) Select RC based on the expected transient voltages.
- 4) Select RS based on the desired longitudinal impedance at Tip and Ring. Transient voltages are also a factor here.
- 5) Calculate RRO based on the desired AC terminating impedance (return loss).
- 6) Calculate RRR based on the desired receive gain.
- 7) Calculate RTX2 and RTX1 based on the desired transmit gain.
- 8) Calculate the balance resistor (RB), or network, as appropriate for desired transhybrid rejection.
- 9) Logic Interface

### Preliminary

There is a primary AC feedback loop which has its main sense points at CP and CN (see Figure 34). The loop extends from there to TXO, through RRO to RXI, through the internal amplifiers to the transistor drivers, through RP to Tip

and Ring, and through the RCs to CP and CN. Components within this loop, such as RP, RC, the transistors, and the compensation capacitors need not be tightly matched to each other in order to maintain good longitudinal balance. The tolerance requirements on these components, and others, are described in subsequent sections. Any components, however, which are placed outside the loop for additional line card functions, such as test relay contacts, fuses, resistors in series with Tip and Ring, etc. will affect longitudinal balance, signal balance, and gains if their values and mismatch is not carefully considered. The MC33121 cannot compensate for mismatch among components outside the loop.

The compensation capacitors (0.01 µF) shown at the transistor collectors (Figure 4) compensate the transistor driver amplifiers, providing the required loop stability. The required tolerance on these capacitors can be determined from the following guidelines:

- A 10% mismatch (±5% tolerance) will degrade the longitudinal balance by ≈ 1.0 dB on a 60 dB device, and by ≈ 3.0 dB on a 70 dB device.
- A 20% mismatch (±10% tolerance) will degrade the longitudinal balance by ≈ 3.0 dB on a 60 dB device, and by ≈ 6.0 dB on a 70 dB device.

High quality ceramic capacitors are recommended since they serve the secondary function of providing a bleedoff path for RF signals picked up on the phone line. These capacitors should be connected to a good quality RF ground.

The capacitors used at CQB and CF must be low leakage to obtain proper performance. Leakage at the CQB capacitor will affect the DC loop current characteristics, while leakage at the CF capacitor will affect the AC gain parameters, and possibly render the IC inoperative.

### 1) Maximum Loop Current and Battery Feed Resistance

The maximum loop current (at RL = 0) is determined by the RRF resistor between RFO and RXI. The current limit is accomplished by three internal series diodes (see Figure 27) which clamp the voltage across RRF as the loop resistance decreases, thereby limiting the current at RXI. Since the loop current is 102 × IRXI, the loop current is therefore clamped. The graphs of Figures 5 to 7 indicate the maximum loop



current at an ambient temperature of + 25°C, and after the IC has reached thermal equilibrium (approx. 10 minutes).

Although the maximum loop current is primarily a function of the RRF resistor, it is also affected by ambient temperature, and slightly by  $V_{EE}$ . The ambient temperature effects are due to the temperature dependence of the diodes' forward voltage drop, causing the maximum loop current to change by  $\approx -0.3\%/^{\circ}\text{C}$ . Changing  $V_{EE}$  affects the maximum current in that the power dissipation is changed, thereby changing the die temperature, which affects the diodes' voltage.

The maximum loop current is affected slightly (< 5%) by the choice of the  $R_S$  and  $R_C$  resistors, since the sense currents through those resistors add to the current supplied by the transistors.

The battery feed resistance is determined by RRF, and is not adjustable independently of the current limit. Defined as  $\Delta V_{TIP}/\Delta I_L$ , it is  $\approx 400\ \Omega$ , and is a valid parameter only on long lines where the current limit is not in effect. On short lines, the feed resistance is high since the loop current is clamped at a near constant level. The AC impedance (Return Loss) however, is not determined nor affected by these DC parameters. Return loss is discussed in another section.

If the application requires that the current limit value have a low temperature dependence, refer to the section following this design sequence which describes an alternate configuration.

## 2) Main Protection Resistors (RP) and Transient Currents

The purpose of the protection resistors (RP), along with the 4 clamp diodes shown in Figure 4, is to absorb the bulk of the transient energy when transient voltages come in from the phone line. The resistor value must be selected to limit the transient current to a value which can be tolerated by the diodes, while dissipating the energy. The recommended value shown (100  $\Omega$ ) will limit the current from a 1500 V transient to 15 A, which can be carried by 1N4002 diodes under surge conditions. The resistors must be of a type which can tolerate the high instantaneous energy associated with transients. Resistor manufacturers should be consulted for this information.

Referring to Figure 4, a positive transient on either Tip or Ring, or both, will cause the transient current to be delivered to Ground. A negative transient will cause the transient current to come from the  $V_{EE}$  supply line. Therefore, the PC board track supplying  $V_{CC}$  and  $V_{EE}$  to the MC33121 must be designed to carry the transient currents as well as the normal operating currents. Additionally, since a negative transient will cause a current flow **out** of the power supply's negative output, which is opposite to the normal flow of current, provisions must be made for this reverse current flow. One suggested method is to place a zener transient suppressor (1N6287 for - 42 V, 1N6282 for - 28 V and - 24 V) across the battery supply pins ( $V_{CC}$  to  $V_{EE}$ ) physically adjacent to the MC33121. The inductance associated with PC board tracks and wiring will result in insufficient protection for the MC33121 if the suppressor is located at the opposite end of the line card, or at the power supplies.

Transient currents can be reduced by increasing the value of RP, with an upper limit determined by the DC conditions on the longest line (highest loop resistance) and minimum  $V_{EE}$  supply voltage. These conditions determine the minimum DC

voltage across the transistors, which must be sufficient to handle the largest AC (transmit and receive) signals. If too large a value is selected for RP, the AC signals will be clipped. It is recommended that each transistor have no less than one volt (DC) across their collector to emitter. System AC specifications may require more than this.

Since the RP resistors are within the loop, their tolerance can be  $\pm 5\%$  with no substantial degradation of longitudinal balance. A  $\pm 10\%$  tolerance (20% mismatch) will degrade balance by  $\approx 4.0\ \text{dB}$  on a 65 dB device.

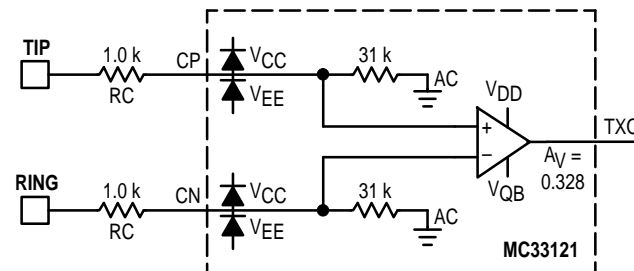


Figure 32. RC Protection Resistors

### 3) Selecting the RC Resistors

The primary purpose of the RC resistors is to protect the CP and CN pins from transient voltages and destructive currents. Internally, these pins have clamp diodes to  $V_{CC}$  and  $V_{EE}$  rated for a maximum of 1.0 A under surge conditions only (Figure 32). The 1.0 k $\Omega$  resistors shown in the figures, for example, will provide protection against surges up to 1.0 kV. Resistor manufacturers must be consulted for the proper type of resistor for this environment.

The RC resistors are in series with internal 31 k $\Omega$  resistors, and therefore form a voltage divider to the inputs of the transmit amplifier, as shown in Figure 32. This will affect the transmit gain, receive gain, return loss, and transhybrid rejection (described in subsequent sections). The tolerance of the RC resistors depends on the value selected for them, since any mismatch between them will create a differential voltage at CP and CN when longitudinal voltages are present on Tip and Ring. To ensure a minimum of 58 dB of longitudinal balance, the resistors' absolute value must not differ by more than 39  $\Omega$ . With a nominal value of 1.0 k $\Omega$ , their tolerance must be  $\pm 2\%$ , or less. If their nominal value is 390  $\Omega$  or less, their tolerance can be  $\pm 5\%$ .

### 4) Longitudinal Impedance ( $Z_{Long}$ ) — Selecting the $R_S$ Resistors

The longitudinal impedance is determined by the  $R_S$  resistors at the TSI and RSI pins according to the following equation:

$$Z_{Long} = \frac{R_S + 100}{51} \quad (4)$$

$Z_{Long}$  is defined as  $V_{Long}/I_{Long}$  as shown in Figure 33; for  $R_S = 9.1\ \text{k}\Omega$ ,  $Z_{Long} = 180\ \Omega$ . The calculated value of  $Z_{Long}$  includes the fact that the  $R_S$  resistors are in parallel with the synthesized impedance. The tolerance of the  $R_S$  resistors therefore depends on how much mismatch can be tolerated between the longitudinal impedances at Tip and at Ring. Calculations indicate the two  $R_S$  resistors can have a  $\pm 5\%$  tolerance, and still comfortably provide a minimum of 58 dB longitudinal balance.



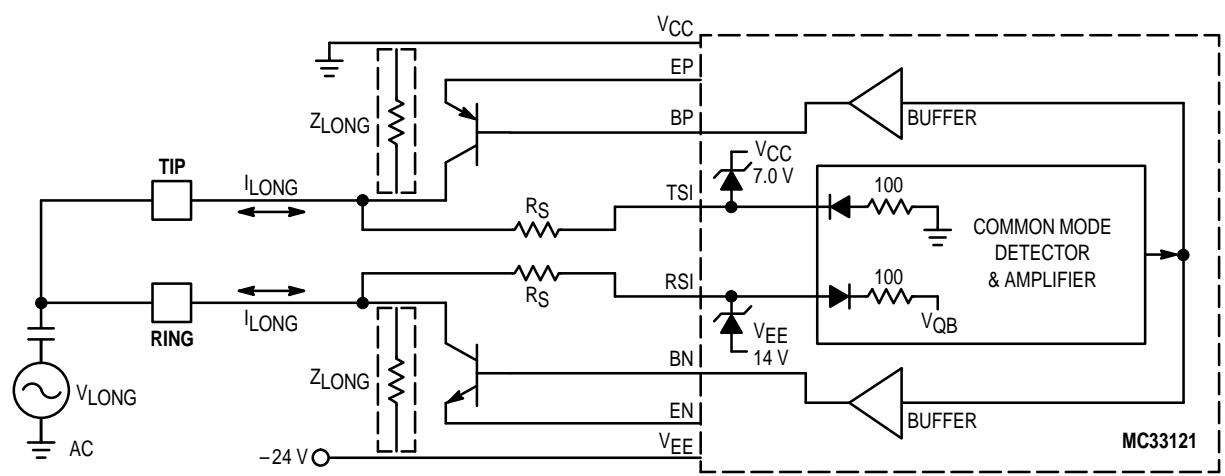
The resistors must be able to withstand transient voltages expected at Tip and Ring. The TSI and RSI pins have internal clamp diodes rated for a maximum of 1.0 A under surge conditions only (Figure 33). Resistor manufacturers must be consulted for the proper type of resistor for this environment.

**5) AC Terminating Impedance and Source Impedance ( $Z_{ac}$ ) — Return Loss**

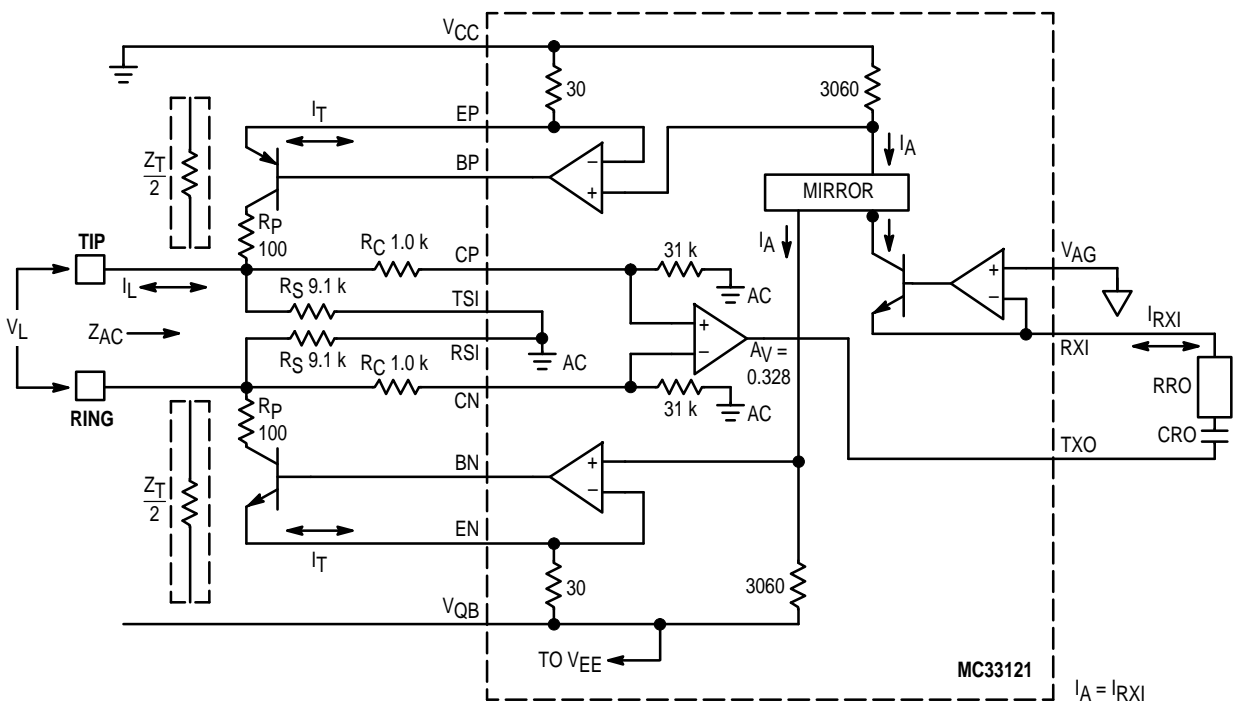
The return loss measurement is a measure of how closely the AC impedance of the SLIC circuit matches the characteristic impedance of the phone line, or a reference imped-

ance. The reference impedance can be, in some cases, a pure resistance (commonly 600  $\Omega$  or 900  $\Omega$ ), a series resistor and capacitor (900  $\Omega$  + 2.16  $\mu$ F), or a more complex network. To achieve proper return loss with the MC33121, the RRO impedance shown in Figure 34 is to have the same configuration as the reference impedance, but with values scaled according to the equations mentioned below.

CRO, used primarily for DC blocking, is generally a large value (1.0  $\mu$ F) so as to not affect the impedance of RRO. However, it can be included in the RRO network if a complex network is required.



**Figure 33. Longitudinal Impedance**



**Figure 34. AC Terminating Impedance**

$Z_{ac}$  is the impedance looking into the circuit from Tip and Ring (set by RRO), and is defined as  $V_L/I_L$ . Half of  $Z_{ac}$  is from Tip to VCC, and the other half is from Ring to VQB (an AC ground). Each half is made up of a synthesized impedance ( $Z_T/2$ ) in parallel with  $R_S$  and  $(RC + 31 k)$ . Therefore,  $Z_{ac}$  is equal to:

$$Z_{ac} = [Z_T/2 // R_S // (RC + 31 k)] \cdot 2 \quad (5)$$

$$\text{and } \frac{Z_T}{2} = \frac{\{R_S // (RC + 31 k)\} \cdot (Z_{ac}/2)}{\{R_S // (RC + 31 k)\} - (Z_{ac}/2)} \quad (6)$$

The synthesized impedance  $Z_T$  is created as follows:

An incoming signal  $V_L$  produces a differential voltage at CP and CN, and therefore at TXO equal to:

$$V_{TXO} = \frac{V_L \cdot 31 k \cdot 0.328}{(RC + 31 k)} \quad (7)$$

The signal at TXO creates an AC current  $I_{RXI}$  through RRO. RXI is a virtual ground, and CRO is insignificant for first order calculations.

$I_{RXI}$  is gained up by a factor of 102 to produce the current  $I_T$  through the transistors.

$Z_T$  is therefore  $V_L/I_T$ . The relationship between  $Z_T$  and RRO is:

$$RRO = \frac{Z_T \cdot 1.037 \cdot 10^6}{(31 k + RC)} \quad (8)$$

While equation 8 gives the exact value for RRO, a first order approximation is  $Z_{ac} \cdot 33.5$ .

a) Resistive Loads (with  $RC = 1.0 k$ ,  $R_S = 9.1 k$ ):

For a  $600 \Omega$  resistive system,  $Z_T$  calculates to  $626 \Omega$ , and RRO calculates to  $20.3 k\Omega$ .

For a  $900 \Omega$  resistive system,  $Z_T$  calculates to  $961 \Omega$ , and RRO calculates to  $31.14 k\Omega$ .

b) Complex Loads

For complex (nonresistive) loads, the MC33121 must be made to look like a termination impedance equal to that complex load. This is accomplished by configuring RRO the same as the complex load, but with all impedance values increased according to the scaling factor of Equation 9.

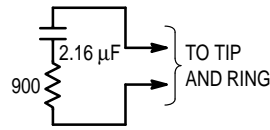
$$SF = \frac{[(RC + 31 k) // R_S] \cdot 1.037 \cdot 10^6}{(RC + 31 k) \cdot [(RC + 31 k) // R_S - (Z_{ac}/2)]} \quad (9)$$

$Z_{ac}$  is computed at a nominal frequency of interest. A first order approximation of Equation 9 is:

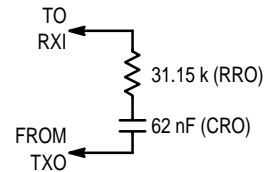
$$SF = 1.037 \cdot 10^6 / (RC + 31 k) \quad (9a)$$

For example:

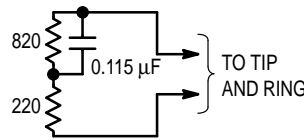
IF THE AC LOAD IS:



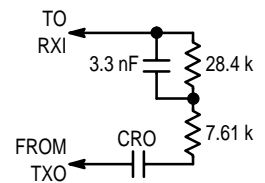
THEN RRO SHOULD BE:



IF THE AC LOAD IS:



THEN RRO SHOULD BE:



CRO must remain in series with the network to provide DC blocking. If the load network does not include a series capacitor (as in the second example above), CRO should be large ( $1.0 \mu F$ ) so its impedance does not affect the RRO network. The above procedure will yield a return loss measurement which is constant with respect to frequency. The RRO resistor, or network, must have a tolerance equal to or better than the required system tolerance for return loss and receive gain.

## 6) Receive Gain ( $G_{RX}$ )

The receive gain involves the same circuit as Figure 34, but with the addition of the RRX resistor (or network) which sets the receive gain. See Figure 35.

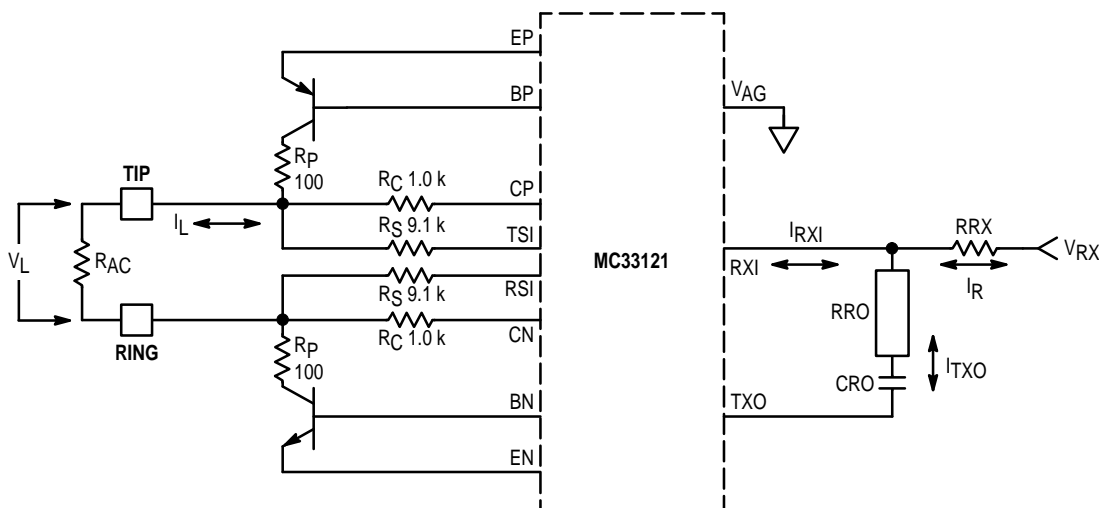


Figure 35. Receive Gain

The receive gain ( $G_{RX}$ ), defined as the voltage gain from  $V_{RX}$  to  $V_L$ , is calculated as follows:

$R_{X1}$  is a virtual ground, and  $R_{ac}$  is the AC impedance of the load (phone line).

The AC current generated in the transistors is  $102 \cdot |I_{RX}|$ , which is equal to  $102 \cdot (I_R - I_{TXO})$ .

$I_R = V_{RX}/RRX$ , and

$$I_{TXO} = \frac{V_{TXO}}{RRO} = \frac{V_L \cdot 31 \text{ k} \cdot 0.328}{RRO \cdot (31 \text{ k} + RC)} \quad (10)$$

Using equations 5 and 8, involving  $Z_{ac}$ ,  $R_S$  and  $R_C$ , and the above equations yields:

$$\frac{V_L}{V_{RX}} = G_{RX} = \frac{102 \cdot (R_{ac}/Z_{ac})}{RRX} \quad (11)$$

$$\text{Therefore, } RRX = \frac{102 \cdot (R_{ac}/Z_{ac})}{G_{RX}} \quad (12)$$

Equation 12 applies **only** for the case where  $R_{ac}$  and  $Z_{ac}$  have the same configuration. If they also have the same magnitude, then set  $RRX = 51 \cdot R_{ac}$  to set a receive gain of 0 dB. The AC source impedance of the above circuit to Tip and Ring is  $Z_{ac}$ . For the case where  $R_{ac} \neq Z_{ac}$ , use the following equation:

$$\frac{V_L}{V_{RX}} = \frac{102}{RRX \cdot \left[ \frac{1}{Z_L} + \frac{1.037 \cdot 10^6}{(31 \text{ k} + RC) \cdot RRO} \right]} \quad (13)$$

$$\text{where } Z_L = \left[ \frac{R_{ac}}{2} // R_S // (RC + 31 \text{ k}) \right] \cdot 2 \quad (14)$$

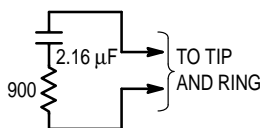
#### a) Resistive Loads

For a 600  $\Omega$  resistive system, set  $RRX = 30.6 \text{ k}\Omega$ , and for a 900  $\Omega$  resistive system, set  $RRX = 45.9 \text{ k}\Omega$ .

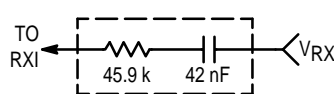
#### b) Complex Loads

For complex (nonresistive) loads, the  $RRX$  resistor needs to be replaced with a network having the same configuration as the complex load, but with all impedance values scaled up by a factor of 51 (for 0 dB gain). If a gain other than 0 dB is desired, the scaling factor is determined from Equation 12. This method applies **only** if the  $RRO$  network has been made complex comparable to the load according to the procedure in the previous section (Equations 5 – 9a), such that  $R_{ac} = Z_{ac}$ . Using a scaling factor of 51, and the previous examples, yields:

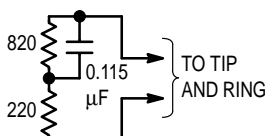
IF THE AC LOAD IS:



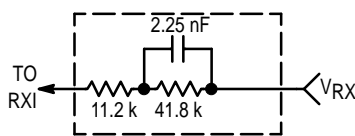
THEN  $RRX$  SHOULD BE:



IF THE AC LOAD IS:



THEN  $RRX$  SHOULD BE:



The preceding procedure will yield a receive gain which is constant with respect to frequency. The  $RRX$  resistor, or network, must have a tolerance equal to or better than the required system tolerance for receive gain.

## 7) Transmit Gain ( $G_{TX}$ )

Setting the transmit gain involves selecting  $RTX1$  and  $RTX2$  in Figure 28. The voltage gain from  $V_L$  to  $V_{TX}$  is calculated from the following:

$$G_{TX} = \frac{V_{TX}}{V_L} = \frac{RTX2 \cdot 31 \text{ k} \cdot 0.328}{RTX1 \cdot (RC + 31 \text{ k})} \quad (15)$$

For 0 dB gain, set  $RTX2 = 3.15 \times RTX1$  (for  $RC = 1.0 \text{ k}$ ). The actual values of  $RTX2$  and  $RTX1$  are not critical — only their ratio so as to provide the proper gain at the op amp. Once the ratio is established, the two resistors can be selected from a set of standard resistor values. The minimum value for  $RTX1$  is limited by the drive capability of  $TXO$ , which is a nominal  $\pm 800 \mu\text{A}$  peak ( $\pm 275 \mu\text{A}$  minimum). As a general rule,  $RTX1$  should be between 5.0  $\text{k}\Omega$  and 20  $\text{k}\Omega$ . The load on  $TXO$  is the parallel combination of  $RTX1$  and  $RRO$ .

$CTX$  is for DC blocking, and is typically a large value (1.0  $\mu\text{F}$ ) so as to not be a significant impedance. In general, it should **not** be used for low frequency rolloff as that will affect the transhybrid rejection (discussed in the next section). Low frequency rolloff should be done after the op amp. High frequency roll-off can be set by placing a capacitor across  $RTX2$ .

For complex loads (at Tip and Ring), if  $RRO$  and  $RRX$  have been made complex comparable to the load as described in the previous sections, neither  $RTX1$  nor  $RTX2$  needs to be complex since both the transmit and receive signals which appear at  $TXO$  will be flat with respect to frequency.

$RTX1$  and  $RTX2$  must have a tolerance equal to or better than the required system tolerance for the transmit gain.

## 8) Balance Network (RB) — Transhybrid Rejection

When a receive signal is applied to  $V_{RX}$  to produce a signal at Tip and Ring, the two-to-four wire arrangement of a hybrid (the MC33121) results in a reflected signal at  $TXO$ . Transhybrid rejection involves canceling that reflected signal before it appears at  $V_{TX}$ . The method used is to insert the RB resistor (or network) as shown in Figure 36. The current  $I_B$ , supplied from  $V_{RX}$ , cancels the current  $I_{TX1}$  supplied from  $TXO$  (Node A is a virtual ground). Good transhybrid cancellation requires that the currents be equal in magnitude **and**  $180^\circ$  out of phase at Node A.

Using the equations for transmit and receive gains, the current  $I_{TX1}$  is equal to:

$$I_{TX1} = \frac{33.5 \cdot V_{RX} \cdot Z_{ac} \cdot Z_L \cdot 31 \text{ k}}{RRX \cdot [Z_{ac} + Z_L] \cdot RTX1 \cdot (RC + 31 \text{ k})} \quad (16)$$

a) For the case where  $RRO$  and  $RRX$  are comparable in configuration to  $Z$ :

Since  $I_B = V_{RX}/RB$ , then  $RB$  can be determined from:

$$RB = \frac{RRX \cdot RTX1 \cdot (RC + 31 \text{ k})}{33.5 \cdot [Z_{ac}/Z_L] \cdot 31 \text{ k}} \quad (17)$$



feedback aspect within this output circuit which provides considerable hysteresis for stability reasons. Its output characteristics are shown in Figures 21 and 22. Due to this configuration, any external pull-up resistance which is applied to this pin must be greater than 15 k $\Omega$ , or the output may not reliably switch from high to low. Any external pull-down resistance does not affect this output's ability to switch from low-to-high, but does affect the maximum longitudinal currents which can be accepted by the circuit (see the section on Longitudinal Current Capability). The capacitor ( $C_T$ ) is required to provide a time delay, for stability reasons, during transitions between off-hook and on-hook. This capacitor additionally affects maximum longitudinal currents, as well as stability during pulse dialing (explained below).

#### b) Hook Status

The MC33121 uses the sense currents at CP and CN to activate the hook status circuit. The sensing is configured such that the circuit monitors the impedance across Tip/Ring, which results in the hookswitch thresholds are minimally affected by the battery voltage. The off-hook to on-hook threshold is affected by the choice of RRF according to the graph of Figure 8, but is not affected by the value of  $R_S$ . The on-hook to off-hook threshold is affected by the value of  $R_S$  according to the graph of Figure 9, but is not affected by RRF. Varying the RC resistors does not affect the thresholds significantly.

When the telephone is on-hook (ST1 = High, ST2 = Low), the MC33121 is internally powered down, the external transistors are shut off, and power consumption is at a minimum. Upon closure of the phone's hookswitch, ST1 will switch low within 10  $\mu$ s. ST2 will then change state slowly due to the external capacitor ( $C_T = 5.0 \mu$ F). There is a  $\approx 8.0$  ms delay for ST2 to reach the threshold necessary to activate the internal bias circuit, which in turn activates the external drive transistors to supply loop current. This delay is necessary to prevent instabilities during the transition to off-hook.

Upon opening the telephone's hookswitch, ST1 will switch high within  $\approx 200 \mu$ s. ST2 then requires  $\approx 60$  ms to reach the threshold to switch off the internal bias circuit, which in turn shuts down the external drive transistors.

#### c) Pulse Dialing

During pulse dialing, ST1 will change state concurrent with the hookswitch. ST2 is kept from switching during pulse dialing by the external capacitor ( $C_T$ ), which keeps the MC33121 in a powered up condition and stable. If the  $C_T$  capacitor is too small, the voltage at ST2 could drop to the PDI threshold (see section e below) during each pulse. This could cause the MC33121 to create additional noise on the line as it would cycle between a power-up and power-down condition with each dialing pulse.

#### d) Fault Detection

Faults are defined as excessive leakage from Tip to  $V_{EE}$  and/or ground, and from Ring to  $V_{EE}$  and/or ground. A single fault is any one of the above conditions, while a double fault is defined as excessive leakage from Tip to  $V_{EE}$  and from Ring to  $V_{CC}$ , as depicted in Figure 38. Refer to Figures 11 – 15 for the resistance,  $R_{LK}$ , which will cause the MC33121 to switch to a power-down condition. If the leakage resistance is less than that indicated in the graphs, the MC33121 will power-down itself and the two external transistors, thereby protecting them from overheating. Both status outputs (ST1 and ST2) will be at a logic low, indicating a fault condition. A fault condition is detected by monitoring an imbalance in the magnitudes of the currents at TSI and RSI, and/or a polarity reversal at Tip and Ring.

The MC33121 will detect the following conditions:

##### 1) When on-hook (see Figure 11):

- < 2.6 k $\Omega$  between Ring and  $V_{CC}$  (depending on  $R_S$  and  $V_{EE}$ ), with no hysteresis at this threshold, or
- < 3.7 k $\Omega$  between Tip and  $V_{EE}$  (depending on  $R_S$  and  $V_{EE}$ ), with no hysteresis at this threshold, or
- Both a and b simultaneously.

Leakage from Tip to  $V_{CC}$  and/or Ring to  $V_{EE}$  are not detected as faults while the MC33121 is on-hook.

##### 2) When off-hook (367 $\Omega$ between Tip and Ring):

- < 400  $\Omega$  between Tip and  $V_{CC}$  ( $R_S = 6.2$  k $\Omega$ ), or
- < 1800  $\Omega$  between Tip and  $V_{EE}$ , or
- < 400  $\Omega$  between Ring and  $V_{EE}$  ( $R_S = 6.2$  k $\Omega$ ), or
- < 1800  $\Omega$  between Ring and  $V_{CC}$ , or
- Both b and d simultaneously

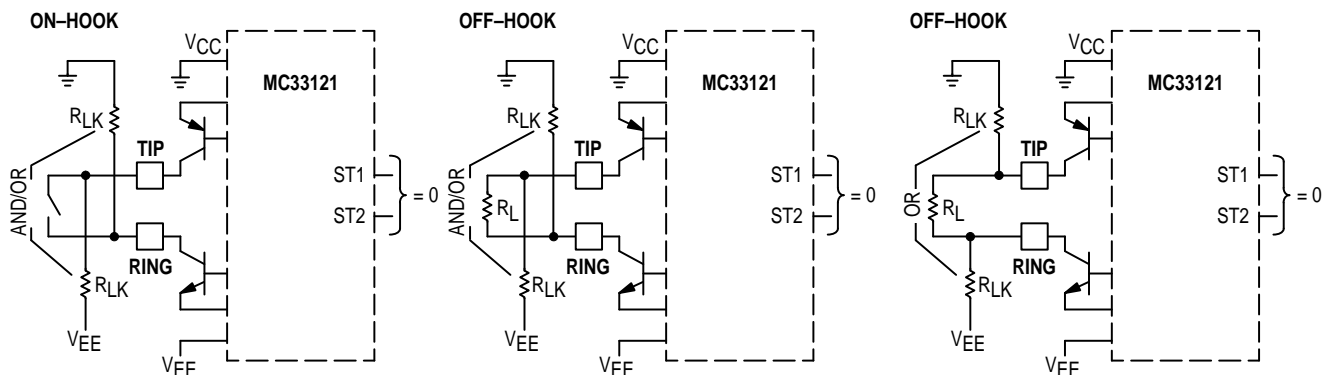


Figure 38. Fault Detection



A simultaneous occurrence of conditions a) and c) is not detected as a fault. See Figures 12 to 15 for the threshold variation with  $R_L$  and  $V_{EE}$ . Resetting of the fault detection circuit requires that the leakage resistance be increased to a value between 10 k $\Omega$  and 20 k $\Omega$ , depending on  $V_{EE}$ ,  $R_L$ , and  $R_S$ . Both ST1 and ST2 should be monitored for hookswitch status to preclude not detecting a fault condition.

Figure 15 indicates the variation in fault thresholds for Tip-to- $V_{CC}$  and Ring-to-Battery faults, and is valid only for loop resistances of 200  $\Omega$  to 800  $\Omega$ . On loops larger than 800  $\Omega$ , the MC33121 does not reliably indicate the fault condition at ST1 and ST2, but may indicate on-hook status instead. This does not apply to Tip-to-Battery and Ring-to- $V_{CC}$  faults which are correctly detected for lines beyond 800  $\Omega$ .

#### e) PDI Input

The ST2 output can also be used as an input (PDI Input) to power down the circuit, denying loop current to the subscriber (by shutting off the external pass transistors), regardless of the hookswitch position. Powering down is accomplished by pulling PDI to a logic low with an open collector output, or an NPN transistor as shown in Figure 30. The switching threshold is  $\approx 1.5$  V. The current out of PDI, when pulled low, is  $\approx 800$   $\mu$ A. Releasing PDI allows the MC33121 to resume normal operation.

If the external telephone is off-hook while the MC33121 is powered down, sense currents at CP and TSI will result in some loop current flowing through the loop and back into CN and RSI. This current is generally on the order of 1.0 to 3.0 mA, determined primarily by the  $R_S$  resistors, loop resistance, and  $V_{EE}$ . ST1 will continue to indicate the telephone's actual hook status while PDI is held low. The on-to-off hook threshold is the same as that during normal operation, but the off-to-on hook threshold is  $> 250$  k $\Omega$ .

When powered down with the PDI pin, the receive gain ( $V_{RXI}$  to Tip/Ring) is muted by  $> 90$  dB, and the transmit gain (Tip/Ring to TXO) is muted by  $> 30$  dB.

### Power Dissipation, Calculation and Considerations

#### a) Reliability

The maximum power dissipated by the MC33121 must be considered, and managed, so as to not exceed the junction temperature listed in the Maximum Ratings Table. Exceeding this temperature on a recurring basis will reduce long term reliability, and possibly degrade performance. The junction temperature also affects the statistical lifetime of the device, due to long term thermal effects within the package. Today's plastic integrated circuit packages are as reliable as ceramic packages under most environmental conditions. However, when the ultimate in system reliability is required, thermal managements must be considered as a prime system design goal.

Modern plastic package assembly technology utilizes gold wire bonded to aluminum bonding pads throughout the electronics industry. When exposed to high temperatures for protracted periods of time an intermetallic compound can form in the bond area resulting in high impedance contacts and degradation of device performance. Since the formation of intermetallic compounds is directly related to device junction temperature, it is incumbent on the designer to determine that the device junction temperature is consistent with system reliability goals.

Based on the results of almost ten years of  $+ 125^\circ\text{C}$  operating life testing, Table 2 has been derived indicating the relationship between junction temperature and time to 0.1% wire bond failure.

**Table 2. Statistical Lifetime**

Junction Temperature ( $^\circ\text{C}$ )	Time (Hours)	Time (Years)
80	1,032,200	117.8
90	419,300	47.9
100	178,700	20.4
110	79,600	9.4
120	37,000	4.2
130	17,800	2.0
140	8,900	1.0

*Motorola MECL Device Data, DL122*

The "Time" in Table 2 refers to the time the device is operating at that junction temperature. Since the MC33121 is at a low power condition (nominally 40 mW) when on-hook, the duty cycle must be considered. For example, if a statistical duty cycle of 20% off-hook time is used, operation at  $130^\circ\text{C}$  junction temperature (when off-hook) would result in a statistical lifetime of  $\approx 10$  years.

#### b) Power and Junction Temperature Calculation

The power within the IC is calculated by subtracting the power dissipated in the two-wire side (the transistors and the load) from the power delivered to the IC by the power supplies. Refer to Figure 4 and 27.

$$P_D = |V_{DD} \cdot I_{DD}| + |V_{EE} \cdot I_{EE}| - (I_L \cdot |V_{EP} - V_{EN}|) \quad (20)$$

The terms  $V_{EP}$  and  $V_{EN}$  are the DC voltages, with respect to ground, at the EP and EN pins. These voltages can be measured, or can be approximated by:

$$V_{EP} \approx - (30 \Omega \cdot I_L)$$

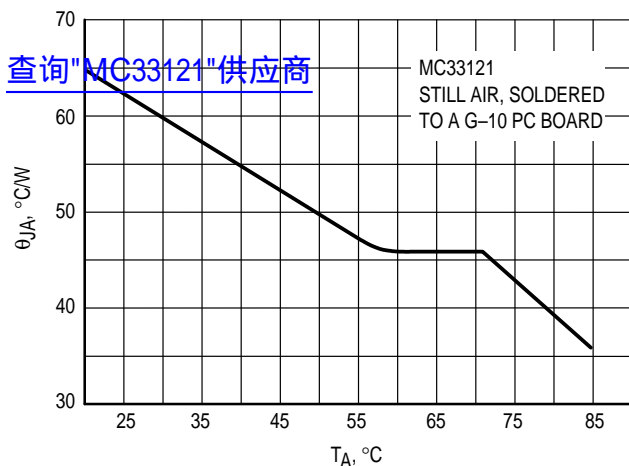
$$V_{EN} \approx |V_{EE}| + 2.1 \text{ V} + (I_L \cdot 35 \Omega)$$

Refer to Figure 23. The junction temperature is then calculated from:

$$T_J = T_A + (P_D \cdot \theta_{JA}) \quad (21)$$

where  $T_A$  is the ambient air temperature at the IC package, and  $\theta_{JA}$  is the junction-to-ambient thermal resistance shown in Figure 39. The highest junction temperature will occur at maximum  $V_{EE}$  and  $V_{DD}$ , maximum loop current, and maximum ambient temperature.

If the above calculations indicate the junction temperature will exceed the maximum specified, then it is necessary to reduce the maximum loop current, ambient temperature, and/or  $V_{EE}$  supply voltage. Air flow should not be restricted near the IC by tall components or other objects since even a small amount of air flow can substantially reduce junction temperature. For example, typically an air flow of 300 LFPM (3.5 mph) can reduce the effective  $\theta_{JA}$  by 14 to 20% from that which occurs in still air. Additionally, providing as much copper area as possible at the IC pins will assist in drawing away heat from within the IC package. For additional information on this subject, refer to the "Thermal Considerations" section of *Motorola MECL System Design Handbook*, and the "System Design Considerations" section of *Motorola MECL Device Data*.



**Figure 39. Thermal Resistance (Junction-to-Ambient)**

### Selecting the Transistors

The specifications for the two loop current pass transistors involve their current gain, voltage rating, and power dissipation capabilities at the highest ambient temperatures. Power dissipation during both normal operation and faults must be considered when determining worst case situations. Generally, more power is dissipated during a fault condition than during normal operation.

The transistors' minimum beta is recommended to be 40 at the loop currents involved in the application. A lower beta could degrade gain and balance performance. Maximum beta should be less than 500 to prevent possible oscillations. Darlington type transistors should not be used. The voltage rating should be consistent with the maximum  $V_{EE}$ , expected transients, and the protection scheme used.

Referring to Figure 27, during normal operation the loop current and the voltage across the transistors are both at a maximum when the load impedance ( $R_L$ ) is at a minimum. The loop current is determined by RRF and the graphs of Figures 5 – 7. The voltage across each transistor is determined from the following:

$$V_T = \frac{|V_{EE}| - 2.1 - [(65 + 2RP + R_L) \cdot I_L]}{2} \quad (22)$$

The power in each transistor is then ( $V_T \cdot I_L$ ). The voltage across the two transistors will always be nearly equal during normal operation, resulting in equal power dissipation. The graph of Figure 24 indicates the power dissipated in each transistor where  $RP = 100 \Omega$ .

During a fault condition, depicted in Figure 38, if the leakage resistance from Tip to  $V_{EE}$  or from Ring to  $V_{CC}$  is less than that shown in Figures 12 – 14 (when off-hook), the MC33121 will power down the transistors to protect them from overheating. Should the leakage resistance be slightly higher than that shown in the graphs, however, and the fault detection has not been activated, the power in one transistor (in a single fault, both transistors in a double fault) will be higher than normal. The power will depend on  $V_{EE}$ ,  $R_L$ ,  $RP$  and the leakage resistance. Table 3 is a guide of the power in the transistor dissipating the higher power level.

The power (in watts) in the two right columns indicates the power dissipated by that transistor if it is carrying the maximum fault current. The system designer should attempt to predict possible fault conditions for the system, and then

measure the conditions on the transistors during the worse case fault(s).

**Table 3. Transistor Power During a Fault**

$V_{EE}$	$R_L$	PNP	NPN
-42	150	0.835	0.615
-24	150	0.257	0.176
-42	600	0.601	0.185
-24	600	0.109	0.057

For most applications involving a maximum loop current of 30 – 40 mA, and a maximum  $T_A$  of + 85°C, and where faults may occur, the MJD243 and MJD253 DPAK transistors are recommended. When mounted as described in their data sheet, they will handle both the normal loop current as well as most fault conditions. If faults are not expected to occur in a particular application, then smaller package transistors, such as MPS6717 and MPS6729, may be used. Each application must be evaluated individually when selecting the transistors.

Other possible transistors which can be considered:

PNP	NPN
MJD253-1	MJD243-1
MJE253	MJE243
MJD32	MJD31
MJD42	MJD41
MJD350	MJD340
TIP30A,B,C	TIP29A,B,C

### Longitudinal Current Capability

The maximum longitudinal current which can be handled without distortion is a function of loop current, battery feed resistance, the longitudinal impedance, and the components on ST2.

Since the pass transistors cannot pass current in the reverse direction, the DC loop current provides one upper boundary for the peak longitudinal current plus peak speech signal current. The battery feed resistance determines, in effect, the DC voltage across the transistors, which is a measure of the headroom available for the circuit to handle the peak longitudinal voltage plus peak speech signal voltage. The longitudinal impedance, determined by the  $R_S$  resistors (equation 4), determines the longitudinal current for a given longitudinal voltage.

While analysis of the above items may yield one value of maximum longitudinal current, a different limit (which may be higher or lower) is imposed by the capacitor  $C_T$ , and any pulldown resistance  $R_T$ , on Pin 12 (ST2). This is due to the fact that the sense currents at TSI and RSI will be alternately mismatched as Tip and Ring move up and down together in the presence of longitudinal signals. When the longitudinals are strong, the internal fault detect circuit is activated with each 1/2 cycle, which attempts to switch ST2 low (see the section on Fault Detection). The speed at which ST2 can switch low is a function of both the external capacitor,  $C_T$  and any pulldown resistance,  $R_T$ .

The graphs of Figures 25 and 26 indicate the maximum longitudinal current which can be handled (in Tip and in Ring) without distortion or causing ST2 to switch low.

### PC Board Layout Considerations

PC board considerations include thermal, RFI/EMI, transient conditions, interconnection of the four wire side to the

codec/filter, and others. Wirewrapped boards should be avoided — breadboarding should be done on a (at least) reasonably near PC board.

a) Thermal

Power dissipated by the MC33121 and the two transistors must be removed to prevent excessively high junction temperatures. The equations for calculating junction temperatures are mentioned elsewhere in this data sheet. Heat is removed by both air flow and copper foil on the PC board. Since even a small amount of air flow substantially reduces junction temperatures compared to still air, tall components or other objects should not be placed such that they block air flow across the heat generating devices. Increasing, wherever possible, the area of the copper foil at the IC pins will provide additional heat removal capability. A ground plane can generally help here, while at the same time helping to reduce RFI problems.

b) RFI/EMI

While the MC33121 is intended for use at audio frequencies, the internal amplifiers have bandwidths in excess of 1.0 MHz, and can therefore respond to externally induced RFI and EMI. Interference signals can come in on the phone line, or be radiated on to the PC board from nearby radio stations or from high frequency circuitry (digital & microprocessor circuitry) in the vicinity of the line card.

Usually RFI entering from the phone line at Tip and Ring can be removed by the compensation capacitors ( $C_C$ ) provided they are connected to a good quality RF ground (generally the same ground which connects to  $V_{CC}$  on the MC33121). The ground track should be as wide and as direct as possible to minimize lead inductance. Generally better results can be obtained if an RF bleedoff to earth (or chassis)

ground can be provided where the twisted pair phone line comes into the system.

To minimize problems due to noise radiating directly onto the PC board from nearby high frequency circuitry, all components associated with the MC33121 should be physically as close as possible to the IC. The most sensitive pins in this respect are the CP, CN, RSI, TSI, VAG and RXI pins. Keeping the tracks short minimizes their “antenna” effect.

c) Transient Conditions

When transient voltages come in to Tip and Ring, the transient currents, which can be several amperes, must be carried by the ground line ( $V_{CC}$ ) an/or the  $V_{EE}$  line. These tracks, along with the protection and clamping devices, must be designed for these currents at the frequencies involved. If the tracks are narrow, not only may they be destroyed by the high currents, but their inductance can allow the voltage at the IC, and other nearby components, to rise to damaging levels.

The protection circuits shown in Figure 4, and in other figures in this data sheet, are such that the bulk of the transient energy is dissipated by external components (the protection resistors and the clamp diodes). The MC33121 has internal diodes to limit voltage excursions on the pins, and to pass a small amount of the transient current — typically less than 1.0 A peak. The arrangement of the diodes is shown in Figure 40.

d) Interconnection of the four-wire side

The connections on the four-wire side to the codec and other digital circuitry involves keeping digital noise out of the speech paths, and also ensuring that potentially destructive transients on Tip and Ring do not get through to the + 5.0 V system.

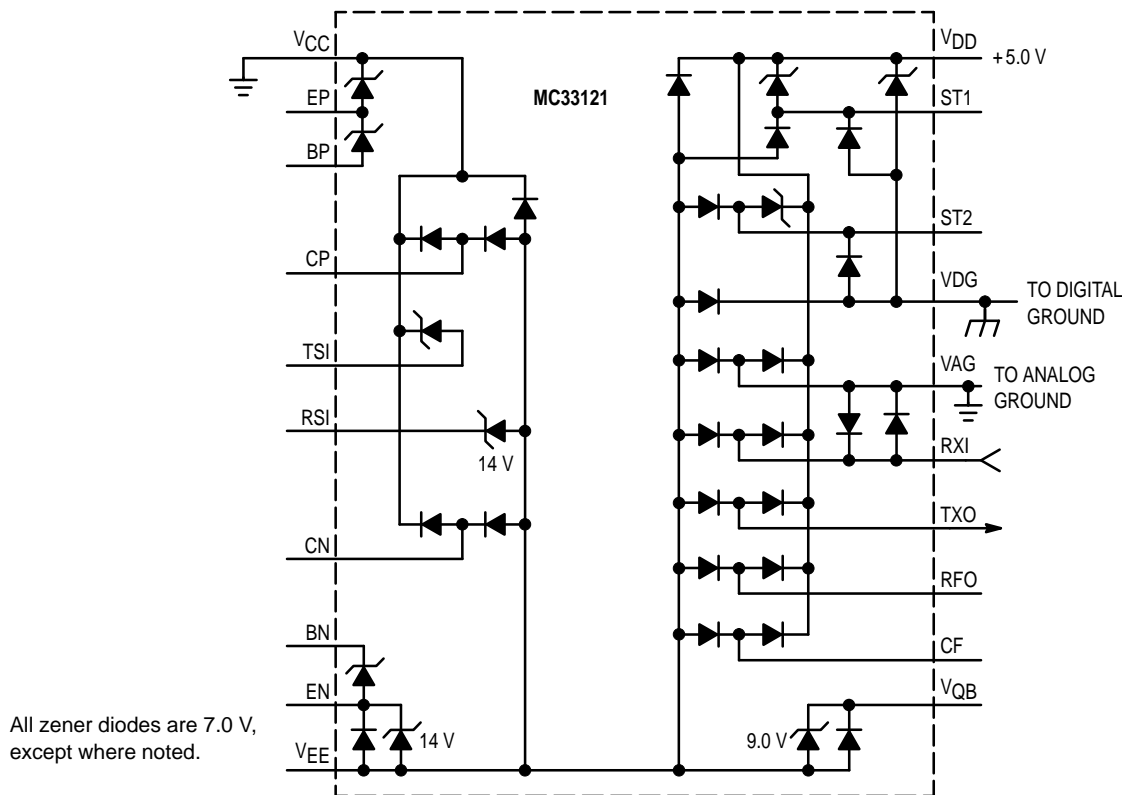


Figure 40. Protection Diodes



Basically, digital connections to ST1 and ST2 should be referenced to the  $V_{DD}$  and  $V_{DG}$  pins, while the transmit and receive analog signals should be referenced to the analog ground ( $V_{AG}$ ).  $V_{CC}$  should be connected to a clean battery ground, and generally should not be connected directly to  $V_{DG}$  and/or  $V_{AG}$  (on the line card) when strong transients are anticipated. Even with a good layout,  $V_{CC}$  can move several volts when a transient hits, possibly damaging components on the +5.0 V line if their grounds have a direct connection at the line card. The MC33121 is designed to allow  $V_{CC}$  to move as much as  $\pm 30$  V with respect to  $V_{DG}$  and  $V_{AG}$  on a transient basis only.  $V_{CC}$  and the other grounds should preferably be connected together at the power supply rather than at the IC. Internally, the MC33121 has clamp diodes on the 4-wire side pins as indicated in Figure 40.

If the codec has a single ground pin, as in Figure 41, it will be the reference for both the digital and analog signals, and must be connected to both  $V_{AG}$  and  $V_{DG}$  on the MC33121. If the codec has separate digital and analog grounds, as in Figure 42 (the MC145503 internally generates the analog ground), then each ground should be connected to the

appropriate ground on the MC33121.

e) Other

A 0.1  $\mu\text{F}$  capacitor should be provided across  $V_{CC}$  to  $V_{EE}$  on the MC33121 to help keep idle channel noise to a minimum.

The  $C_{QB}$  capacitor (on the  $V_{QB}$  pin) forms a pole with an internal 7.5 k $\Omega$  resistor to filter noise from the  $V_{EE}$  pin, providing an internal quiet battery supply for the speech amplifiers. Power supply rejection will depend on the value and quality of this capacitor at the frequencies of concern. Tantalum capacitors generally have better high frequency characteristics than electrolytics. See Figure 17 and 18 for ripple rejection characteristics (the four-wire data was measured at pin 11 (TXO)). Figure 16 indicates ripple rejection from the +5.0 V supply ( $V_{DD}$ ).

In general, pc board tracks carrying analog signals (on the four-wire side and Tip/Ring) should not be routed through the digital section where they could pick up digital noise. Any tracks longer than a few inches should be considered an antenna and should be checked for potential noise or RFI pick-up which could affect the circuit operation.

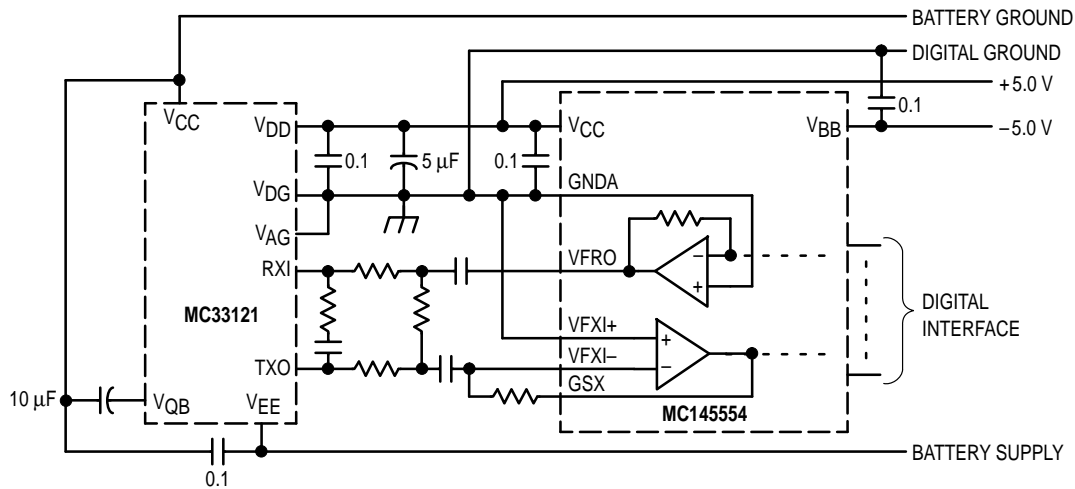


Figure 41. Connection to a CODEC With a Single Ground

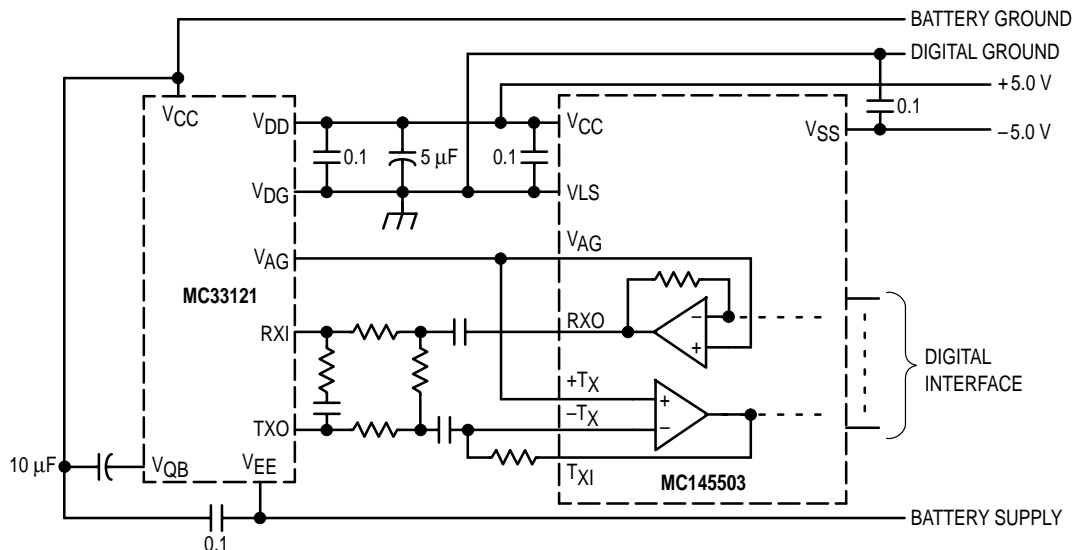


Figure 42. Connection to a CODEC With Separate Grounds

### Alternate Circuit Configurations

#### Alternate Current Limit

Replacing the RRF resistor with the circuit in Figure 43 will change the DC loop current characteristics in two ways from the graphs of Figures 5 – 7; a) the maximum loop current on a short line can be reduced while increasing the current on a long line, and b) the temperature dependence of the maximum current is reduced to the TC of the external reference diode.

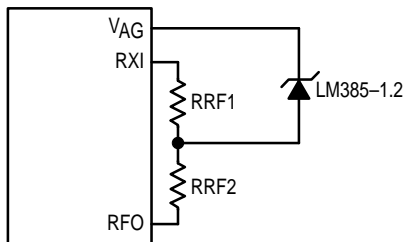


Figure 43. Alternate Current Limit Circuit

The LM385-1.2 is a precision temperature stable zener diode. As the load impedance at Tip and Ring is reduced, the voltage at RFO goes increasingly negative. When the zener diode is turned on, the current into RXI is then clamped at a value determined by RRF1 and the zener diode. To calculate the two resistors, use the following procedure:

RRF1 must be  $> 0.7 \cdot (RRF1 + RRF2)$ ;

Determine RRF1 to set the current limit on a short line by using the following equation:

$$RRF1 = \frac{102 \cdot 1.23 \text{ V}}{I_{L(\max)} - 3.0 \text{ mA}} \quad (23)$$

Then using Equation 1 calculate RRF for the long line current. RRF2 is then determined by;

$$RRF2 = RRF - RRF1 \quad (24)$$

Figure 44 illustrates one example using the above circuit. Comparing this graph to the 5100 Ω curve of Figure 7 shows a substantial decrease in the current limit (at  $R_L = 0$ ), resulting in reduced power consumption and dissipation. Use of this circuit does not affect the hookswitch or fault thresholds.

#### b) Protection Scheme

The protection circuit shown in Figure 45 has the advantage of drawing  $\approx 90\%$  of the transient current from ground ( $V_{CC}$ ) on a negative transient, rather than from the  $V_{EE}$  line as the circuit of Figure 4 does. The majority of the transient current flows through the RP resistors and the Mosorbs while

a small amount ( $\approx 10\%$ ) flows through the sense resistors and the CP, CN, RSI pins. On a positive transient, all the current is directed to ground. The diode in the NPN's collector prevents reverse current through the base-collector junction of the transistor during a negative transient.

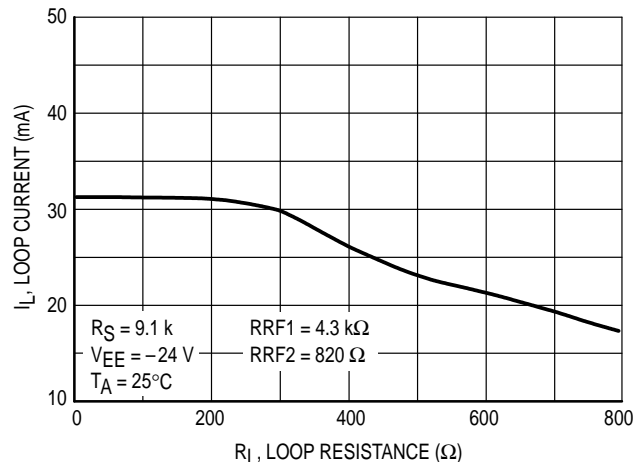
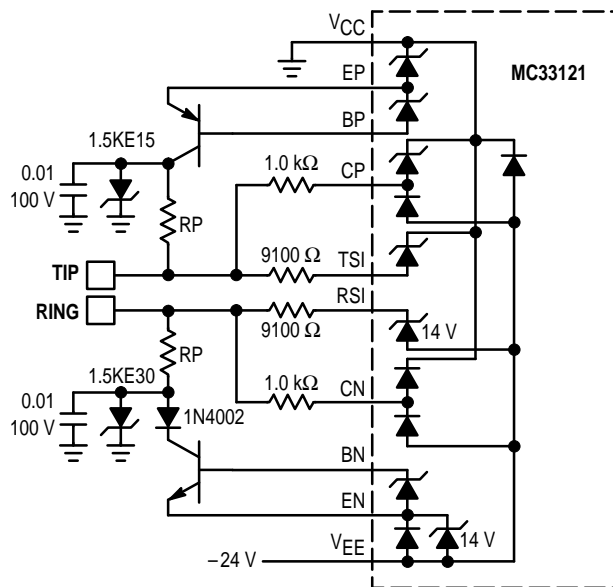


Figure 44. Loop Current versus Loop Resistance  
Alternate Loop Current Limit Configuration



All zener diodes are 7.0 V except as noted.

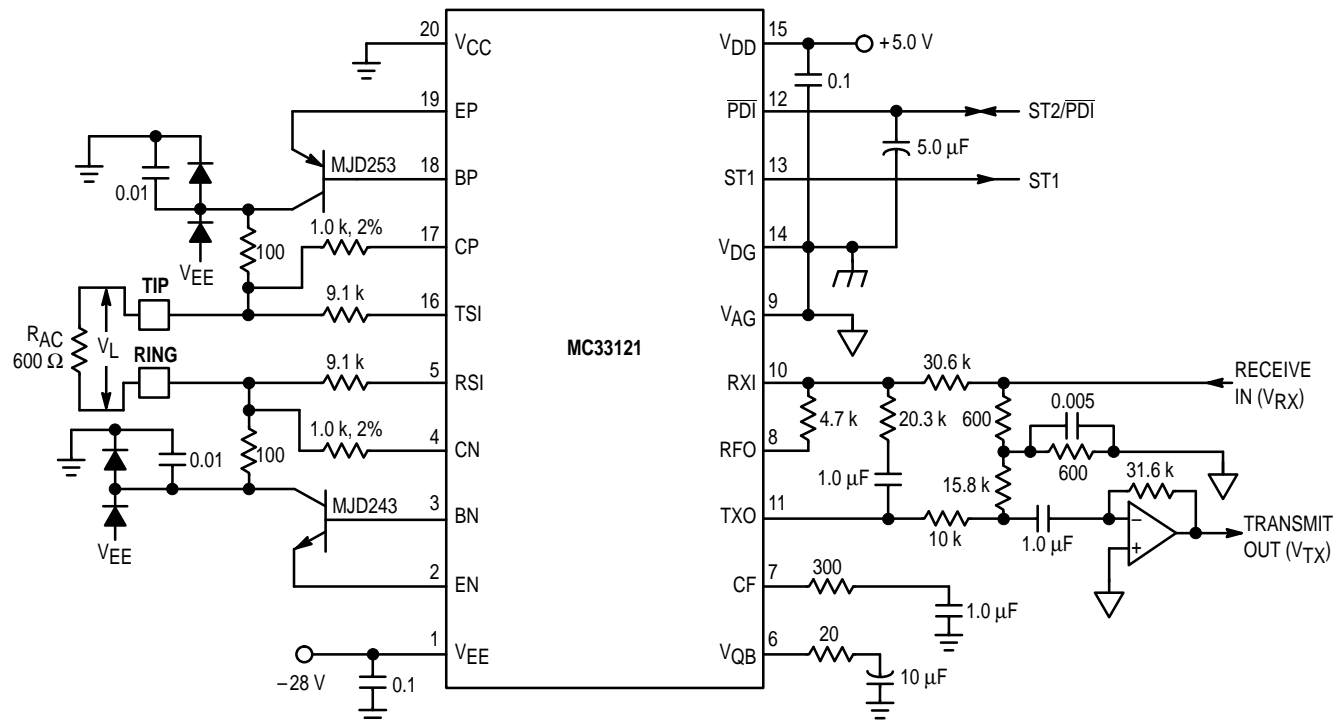
Figure 45. Alternate Protection Scheme

## CIRCUIT PERFORMANCE

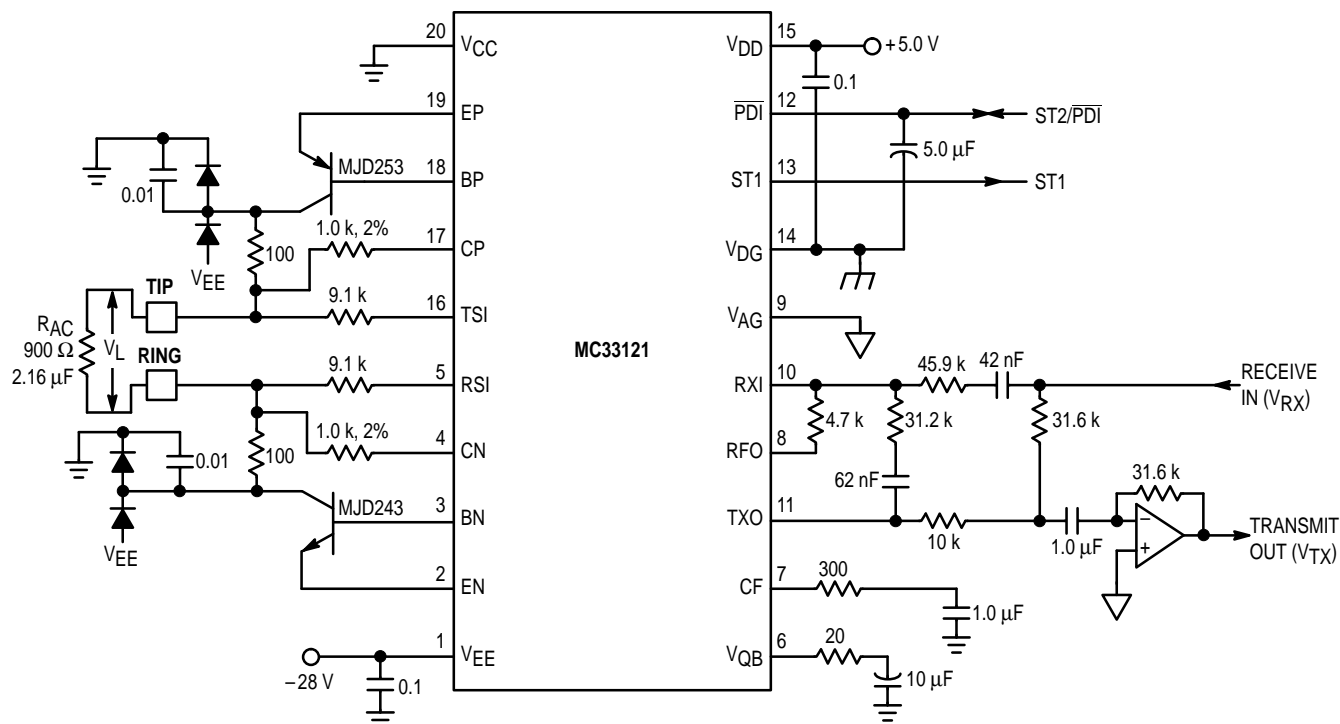
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The following three circuits are presented as typical application examples, and the accompanying graphs indicate their measured performance. The first circuit (Figure 46) has a 600  $\Omega$  pure resistance as the AC load. The second circuit (Figures 47) has as an AC load a 900  $\Omega$  resistor in series with a 2.16  $\mu\text{F}$  capacitor. The third circuit (Figure 48) has as an

AC load, a complex network composed of an 820  $\Omega$  resistor in parallel with 0.115  $\mu\text{F}$ , and those in series with a 220  $\Omega$  resistor. In the graphs of Figures 49 – 51,  $R_L$  = Return Loss, THR = Transhybrid Rejection, GTX = Transmit Gain, GRX = Receive Gain.



**Figure 46. 600  $\Omega$  System**



**Figure 47. 900  $\Omega$  and 2.16  $\mu\text{F}$  System**

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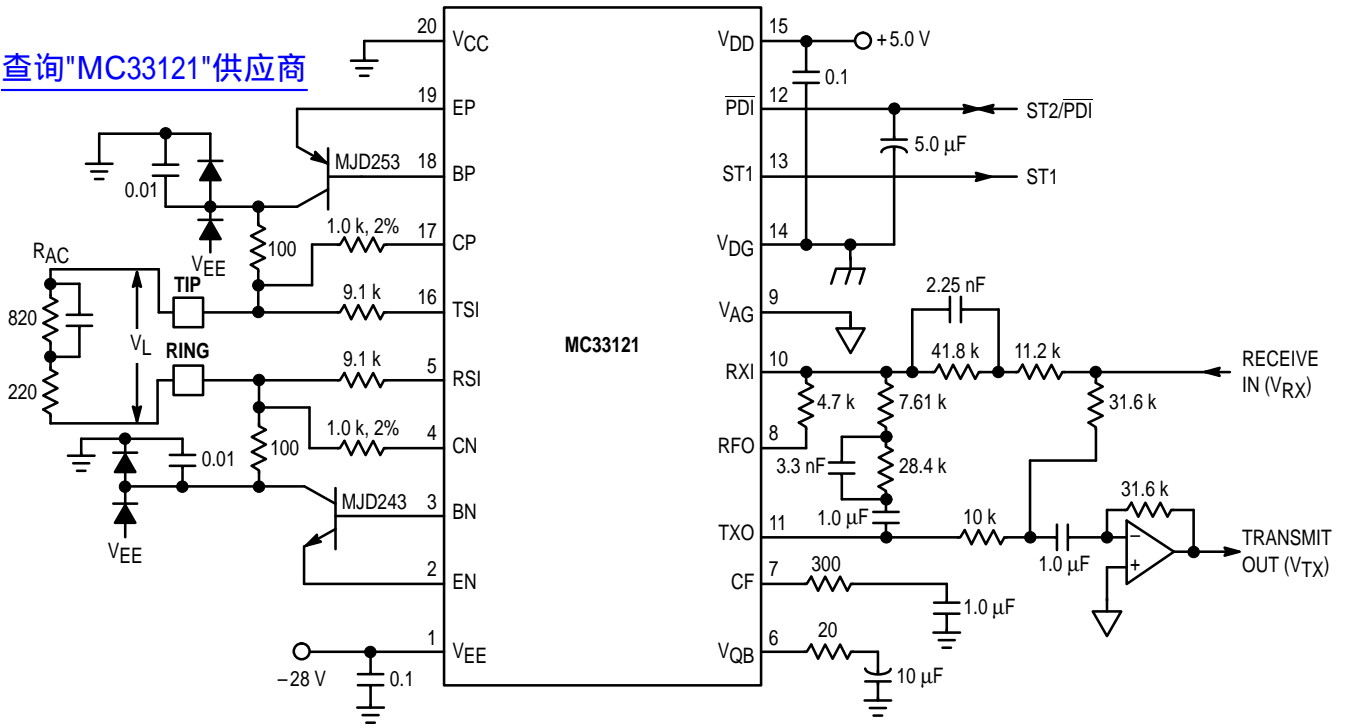


Figure 48. 220  $\Omega$  and 820  $\Omega$ //0.115  $\mu$ F System

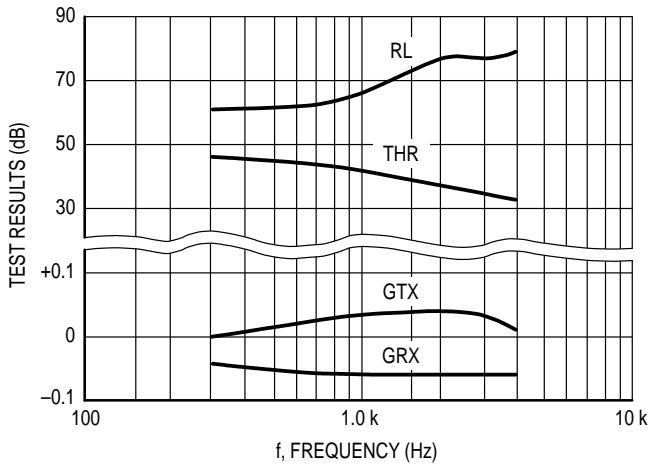


Figure 49. Circuit Performance, 600  $\Omega$  System

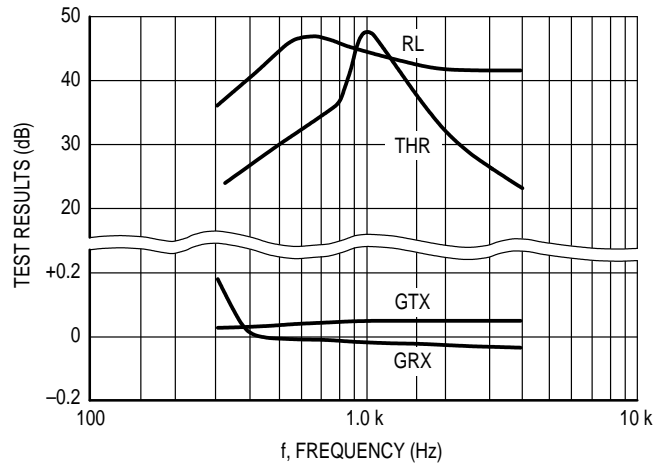
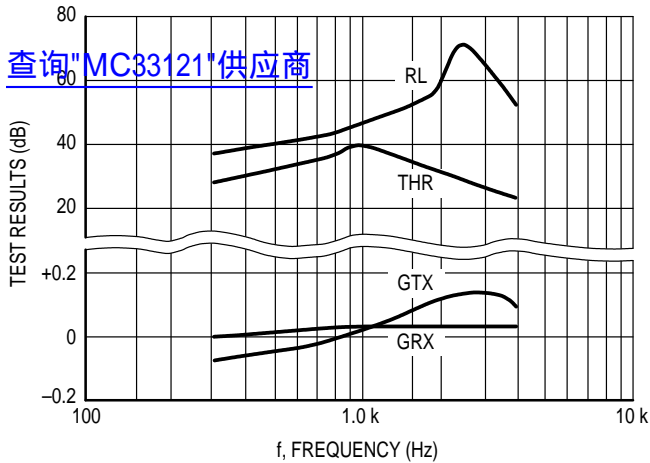
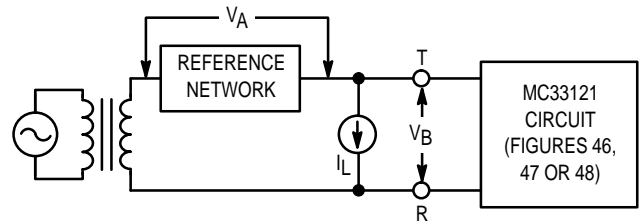


Figure 50. Circuit Performance 900  $\Omega$  and 2.16  $\mu$ F System



**Figure 51. Circuit Performance**  
820 Ω//0.115 μF and 220 Ω System



Reference Network =  $R_{AC}$  of Figures 46 to 48.  
Return Loss =  $20 \log \frac{|V_A + V_B|}{|V_A - V_B|}$

**Figure 52. Return Loss Test Circuit**  
for Figures 46 to 51

## GLOSSARY

**ATTENUATION** — A decrease in magnitude of a communication signal, usually expressed in dB.

**BALANCE NETWORK** — That part of the SLIC circuit which provides transhybrid rejection.

**BANDWIDTH** — The range of information carrying frequencies of a communication system.

**BATTERY** — The voltage which provides the loop current, and in some cases powers the SLIC circuit. The name derives from the fact that COs have always used batteries, in conjunction with AC power, to provide this voltage.

**BATTERY FEED RESISTANCE** — The equivalent Thevenin DC resistance of the SLIC circuit for supplying loop current. Traditionally it is 400 Ω.

**C-MESSAGE FILTER** — A frequency weighting which evaluates the effects of noise on a typical subscriber's system.

**CENTRAL OFFICE** — Abbreviated CO, it is a main telephone office, usually within a few miles of its subscribers, that houses switching gear for interconnection within its exchange area, and to the rest of the telephone system. A typical CO can handle up to 10,000 subscriber numbers.

**CODEC** — Coder/Decoder — Interfacing between the SLIC and the digital switch, it converts the SLIC's transmit signal to digital, and converts the digital receive signal to analog.

**dB** — A power or voltage measurement unit, referred to another power or voltage. It is generally computed as:

$$10 \cdot \log (P_1/P_2) \text{ for power measurements, and} \\ 20 \cdot \log (V_1/V_2) \text{ for voltage measurements.}$$

**dBm** — An indication of signal power. 1.0 mW across 600 Ω, or 0.775 V<sub>rms</sub>, is defined as 0 dBm. Any other voltage level

is converted to dBm by:

$$\text{dBm} = 20 \cdot \log (V_{\text{rms}}/0.775), \text{ or} \\ \text{dBm} = [20 \cdot \log (V_{\text{rms}})] + 2.22.$$

**dBmp** — Indicates dBm measurement using a psophometric weighting filter.

**dB<sub>rn</sub>** — Indicates a dBm measurement relative to 1.0 pW power level into 600 Ω. Generally used for noise measurements, 0 dB<sub>rn</sub> = -90 dBm.

**dB<sub>rnC</sub>** — Indicates a dB<sub>rn</sub> measurement using a C-message weighting filter.

**DTMF** — Dual Tone Multifrequency. It is the "tone dialing" system based on outputting two non-harmonic related frequencies simultaneously to identify the number dialed. Eight frequencies have been assigned to the four rows and four columns of a keypad.

**FAULT** — An incorrect condition where Tip is accidentally connected to the battery voltage, or Ring is connected to ground, or both. The most common fault is Ring to ground.

**FOUR WIRE CIRCUIT** — The portion of a telephone, or central office, which operates on two pairs of wires. One pair is for the transmit path, and one pair is for the receive path.

**FULL DUPLEX** — A transmission system which permits communication in both directions simultaneously. The standard handset telephone system is full duplex.

**GAIN** — The change in signal amplitude (increase or decrease) after passing through an amplifier, or other circuit stage. Usually expressed in dB, an increase is a positive number, and a decrease is a negative number.

**HALF DUPLEX** — A transmission system which permits communication in one direction at a time. CB radios, with push-to-talk switches, and voice activated speaker-phones, are half duplex.

**HOOKSWITCH** — A switch, within the telephone, which connects the telephone circuit to the subscriber loop. The name derives from old telephones where the switch was activated by lifting the receiver off and onto a hook on the side of the phone.

**HYBRID** — Another name for a two-to-four wire converter.

**IDLE CHANNEL NOISE** — Residual background noise when transmit and receive signals are absent.

**LINE CARD** — The PC board and circuitry in the CO or PBX which connects to the subscriber's phone line. A line card may hold circuitry for one subscriber, or a number of subscribers.

**LONGITUDINAL BALANCE** — The ability of the SLIC to reject longitudinal signals on Tip and Ring.

**LONGITUDINAL SIGNALS** — Common mode signals.

**LOOP** — The loop formed by the two subscriber wires (Tip and Ring) connected to the telephone at one end, and the central office (or PBX) at the other end. Generally, it is a floating system not referred to ground, or AC power.

**LOOP CURRENT** — The DC current which flows through the subscriber loop. It is typically provided by the central office or PBX, and ranges from 20 to 120 mA.

**OFF-HOOK** — The condition when the telephone is connected to the phone system, permitting the loop current to flow. The central office detects the DC current as an indication that the phone is busy.

**ON-HOOK** — The condition when the telephone is disconnected from the phone system, and no DC loop current flows. The central office regards an on-hook phone as available for ringing.

**PABX** — Private Automatic Branch Exchange. In effect, a miniature central office, it is a customer owned switching system servicing the phones within a facility, such as an office building. A portion of the PABX connects to the Bell (or other local) telephone system.

**PROTECTION, PRIMARY** — Usually consisting of carbon blocks or gas discharge tubes, it absorbs the bulk of a lightning induced transient by clamping the voltages to less than  $\pm 1500$  V.

**PROTECTION, SECONDARY** — Usually located on the line card, it protects the SLIC and associated circuits from transient surges. Typically, it must be capable of clamping a  $\pm 1.5$  kV surge of 1.0 ms duration.

**PULSE DIALING** — A dialing system whereby the loop current is interrupted a number of times in quick succession. The number of interruptions corresponds to the number dialed, and the interruption rate is typically 10 per second. The old rotary phones, and many new pushbutton phones, use pulse dialing.

**RECEIVE PATH** — Within the CO or PBX it is the speech path from the internal switching system towards the phone line (Tip & Ring).

**REN** — Ringer Equivalence Number. An indication of the impedance or loading factor of a telephone bell or ringer circuit. An REN of 1.0 equals  $\approx 8.0$  k $\Omega$ . The Bell system typically permits a maximum of 5.0 REN (1.6 k $\Omega$ ) on an individual subscriber line. A minimum REN of 0.2 (40 k $\Omega$ ) is required by the Bell system.

**RETURN LOSS** — Expressed in dB, it is a measure of how well the SLIC's AC impedance matches the line's AC characteristic impedance. With a perfect match, there is no reflected signal, and therefore infinite return loss. It is calculated from:

$$RL = 20 \cdot \log \frac{(Z_{Line} + Z_{CKT})}{(Z_{Line} - Z_{CKT})}$$

**RING** — One of the two wires connecting the central office to a telephone. The name derives from the ring portion of the plugs used by operators (in older equipment) to make the connection. Ring is traditionally negative with respect to Tip.

**SLIC** — Subscriber Line Interface Circuit. It is the circuitry within the CO or PBX which connects to the user's phone line.

**SUBSCRIBER** — The customer at the telephone end of the line.

**SUBSCRIBER LINE** — The system consisting of the user's telephone, the interconnecting wires, and the central office equipment dedicated to that subscriber (also referred to as a loop).

**TIP** — One of the two wires connecting the central office to a telephone. The name derives from the tip of the plugs used by operators (in older equipment) to make the connection. Tip is traditionally positive with respect to Ring.

**TRANSYBRID REJECTION** — The rejection (in dB) of the reflected signal in the transmit path resulting from a receive signal applied to the SLIC.

**TRANSMIT PATH** — Within the CO or PBX it is the speech path from the phone line (Tip & Ring) towards the internal switching system.

**TWO WIRE CIRCUIT** — Refers to the two wires connecting the central office to the subscriber's telephone. Commonly referred to as Tip and Ring, the two wires carry both transmit and receive signals in a differential manner.

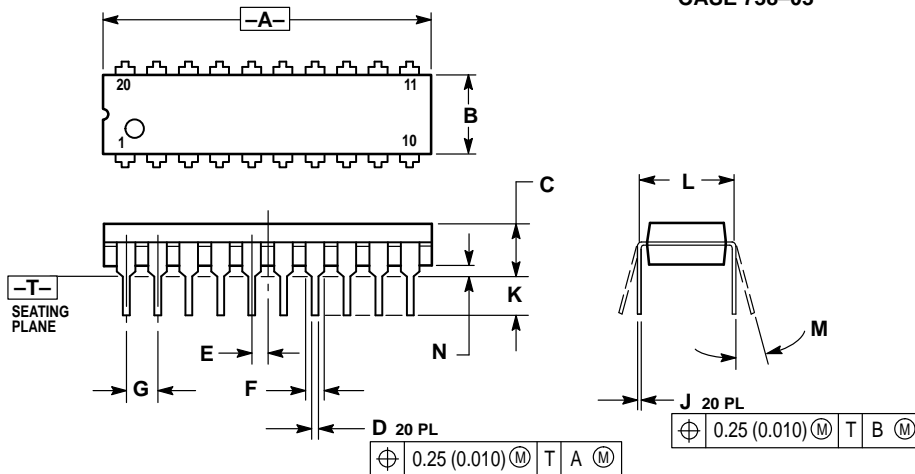
**TWO-TO-FOUR WIRE CONVERTER** — A circuit which has four wires (on one side) — two (signal & ground) for the outgoing signal, and two for the incoming signal. The outgoing signal is sent out differentially on the two wire side (the other side), and incoming differential signals received on the two wire side are directed to the four wire side. Additional circuit within cancels the reflected outgoing signal to keep it separate from the incoming signal.

**VOICEBAND** — That portion of the audio frequency range used for transmission across the telephone system. Typically, it is 300 to 3400 Hz.

## PACKAGE DIMENSIONS

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### P SUFFIX PLASTIC PACKAGE CASE 738-03

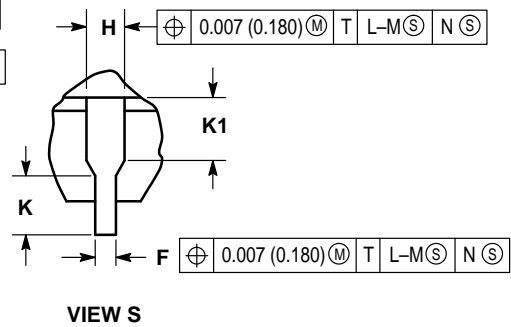
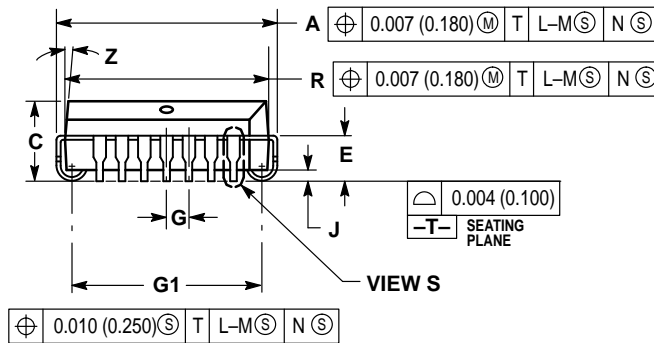
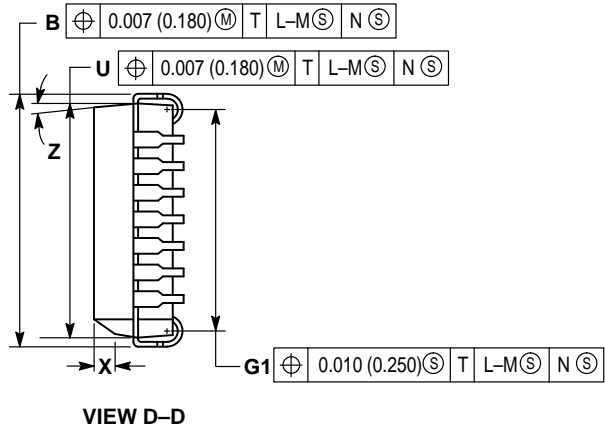
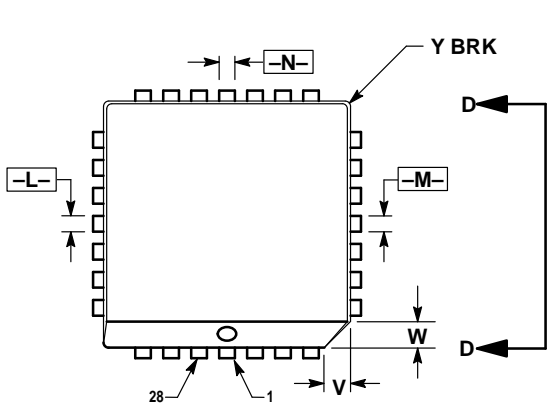


#### NOTES:

1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.
3. DIMENSION L TO CENTER OF LEAD WHEN FORMED PARALLEL.
4. DIMENSION B DOES NOT INCLUDE MOLD FLASH.

DIM	INCHES		MILLIMETERS	
	MIN	MAX	MIN	MAX
A	1.010	1.070	25.66	27.17
B	0.240	0.260	6.10	6.60
C	0.150	0.180	3.81	4.57
D	0.015	0.022	0.39	0.55
E	0.050 BSC		1.27 BSC	
F	0.050	0.070	1.27	1.77
G	0.100 BSC		2.54 BSC	
J	0.008	0.015	0.21	0.38
K	0.110	0.140	2.80	3.55
L	0.300 BSC		7.62 BSC	
M	0°	15°	0°	15°
N	0.020	0.040	0.51	1.01

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NOTES:

- DATUMS  $-L-$ ,  $-M-$ , AND  $-N-$  DETERMINED WHERE TOP OF LEAD SHOULDER EXITS PLASTIC BODY AT MOLD PARTING LINE.
- DIMENSION G1, TRUE POSITION TO BE MEASURED AT DATUM  $-T-$ , SEATING PLANE.
- DIMENSIONS R AND U DO NOT INCLUDE MOLD FLASH. ALLOWABLE MOLD FLASH IS 0.010 (0.250) PER SIDE.
- DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
- CONTROLLING DIMENSION: INCH.
- THE PACKAGE TOP MAY BE SMALLER THAN THE PACKAGE BOTTOM BY UP TO 0.012 (0.300). DIMENSIONS R AND U ARE DETERMINED AT THE OUTERMOST EXTREMES OF THE PLASTIC BODY EXCLUSIVE OF MOLD FLASH, THE BAR BURRS, GATE BURRS AND INTERLEAD FLASH, BUT INCLUDING ANY MISMATCH BETWEEN THE TOP AND BOTTOM OF THE PLASTIC BODY.
- DIMENSION H DOES NOT INCLUDE DAMBAR PROTRUSION OR INTRUSION. THE DAMBAR PROTRUSION(S) SHALL NOT CAUSE THE H DIMENSION TO BE GREATER THAN 0.037 (0.940). THE DAMBAR INTRUSION(S) SHALL NOT CAUSE THE H DIMENSION TO BE SMALLER THAN 0.025 (0.635).

DIM	INCHES		MILLIMETERS	
	MIN	MAX	MIN	MAX
A	0.485	0.495	12.32	12.57
B	0.485	0.495	12.32	12.57
C	0.165	0.180	4.20	4.57
E	0.090	0.110	2.29	2.79
F	0.013	0.019	0.33	0.48
G	0.050 BSC		1.27 BSC	
H	0.026	0.032	0.66	0.81
J	0.020	—	0.51	—
K	0.025	—	0.64	—
R	0.450	0.456	11.43	11.58
U	0.450	0.456	11.43	11.58
V	0.042	0.048	1.07	1.21
W	0.042	0.048	1.07	1.21
X	0.042	0.056	1.07	1.42
Y	—	0.020	—	0.50
Z	2°	10°	2°	10°
G1	0.410	0.430	10.42	10.92
K1	0.040	—	1.02	—

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