HC55185 and the Texas Instruments TCM38C17 Quad Combo

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Reference Design using the HC55185 and the Texas Instruments TCM38C17 Quad Combo

The purpose of this application note is to provide a reference design for the HC55185 and Texas Instruments TCM38C17 Quad Combo.

The network requirements of many countries require the analog subscriber line circuit (SLIC) to terminate the subscriber line with an impedance for voiceband frequencies which is complex, rather than resistive (e.g. 600Ω). The HC55185 accomplishes this impedance matching with a single network connected between the VTX pin and the -IN pin.

The TCM38C17 Quad Combo has a programmable receive output amplifier to adjust the output gain into the SLIC. The output amplifier gain is programmed with two simple resistors. Transhybrid balance is achieved via the TCM38C17 GSX amplifier.

Discussed in this application note are the following:

- 2-wire 600Ω impedance matching.
- Receive gain (4-wire to 2-wire) and transmit gain (2-wire to 4-wire) calculations.
- Transhybrid balance calculations.
- Reference design for 600Ω 2-wire load.
- Reference design for China complex 2-wire load.

Impedance Matching

Impedance matching of the HC55185 to the subscriber load is important for optimization of 2 wire return loss, which in turn cuts down on echoes in the end to end voice communication path. Impedance matching of the HC55185 is accomplished by making the SLIC's impedance (Z_O, Figure 1) equal to the desired terminating impedance Z_L, minus the value of the protection resistors (R_P). The formula to calculate the proper R_S for matching the 2-wire impedance is shown in Equation 1.

$$R_S = 133.3 \bullet (Z_L - 2R_P) \quad \text{(EQ. 1)}$$

Equation 1 can be used to match the impedance of the SLIC and the protection resistors (Z_{TR}) to any known line impedance (Z_L). Figure 1 shows the calculations of R_S to match a resistive and 2 complex loads.

EXAMPLE 1:

Calculate R_S to make Z_{TR} = 600Ω in series with 2.16µF. R_P = 49Ω.

$$R_S = 133.3\left(600 + \frac{1}{j\omega 2.16X10^{-6}} - (2)(49)\right) \quad \text{(EQ. 2)}$$

R_S = 66.9kΩ in series with 16.2nF.

Note: Some impedance models, with a series capacitor, will cause the op-amp feedback to behave as an open circuit DC. A resistor with a value of about 10 times the reactance of the R_S capacitor (2.16µF/133.3 = 16.2nF) at the low frequency of interest (200Hz for example) can be placed in parallel with the capacitor in order to solve the problem (491kΩ for a 16.2nF capacitor).

EXAMPLE 2:

Calculate R_S to make Z_{TR} = 200 + 680//0.1µF R_P = 49Ω.

$$Z_T = 133.3\left(200 + \frac{680}{1 + j\omega 680(0.1)X10^{-6}} - (2)(49)\right) \quad \text{(EQ. 3)}$$

R_S = 13.6kΩ in series with the parallel combination of 90.6kΩ and 750pF.

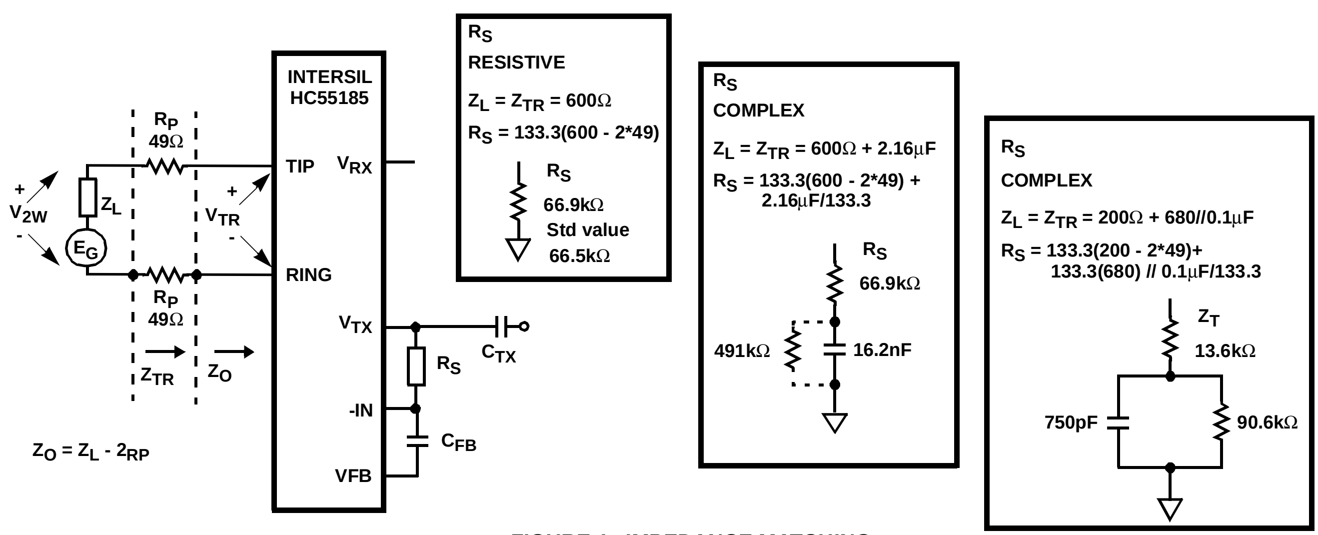

FIGURE 1. IMPEDANCE MATCHING

SLIC in the Active Mode

Figure 2 shows a simplified AC transmission model of the HC55185 and the connection of the TCM38C17 to the SLIC. Circuit analysis of the HC55185 yields the following design equations:

The Sense Amplifier is configured as a 4 input differential amplifier with a gain of 3/4. The voltage at the output of the sense amplifier (V_{SA}) is calculated using superposition. $V_{SA}1$ is the voltage resulting from V1, $V_{SA}2$ is the voltage resulting from V2 and so on (reference Figure 2).

$$V_{SA}1 = -\frac{3}{4}(V_1) \quad \text{(EQ. 4)}$$

$$V_{SA}2 = \frac{3}{4}(V_2) \quad \text{(EQ. 5)}$$

$$V_{SA}3 = -\frac{3}{4}(V_3) \quad \text{(EQ. 6)}$$

$$V_{SA}4 = \frac{3}{4}(V_4) \quad \text{(EQ. 7)}$$

$$V_{SA} = [(V_2 - V_1) + (V_4 - V_3)]\frac{3}{4} = [\Delta V + \Delta V]\frac{3}{4} \quad \text{(EQ. 8)}$$

Where ΔV is equal to $I_M R_{SENSE}$ ($R_{SENSE} = 20\Omega$)

$$V_{SA} = 2(\Delta I_M \times 20)\frac{3}{4} = \Delta I_M 30 \quad \text{(EQ. 9)}$$

The voltage at VTX is equal to:

$$V_{TX} = -V_{SA}\left(\frac{R_S}{8K}\right) = -\left(\frac{R_S}{8K}\right)\Delta I_M 30 \quad \text{(EQ. 10)}$$

V_{TR} is defined in Figure 2, note polarity assigned to $V_{\mathbf{TR}}$:

$$V_{TR} = 2(V_{RX} + V_{TX}) \quad \text{(EQ. 11)}$$

Setting V_{RX} equal to zero, substituting EQ. 10 into EQ. 11 and defining $Z_O = -V_{TR}/\Delta I_M$ will enable the user to determine the require feedback to match the line impedance at V_{2W}.

$$Z_O = \frac{1}{133.33}R_S \quad \text{(EQ. 12)}$$

Z_O is the source impedance of the device and is defined as $Z_O = Z_L - 2R_p$. ZL is the line impedance. R_S is defined as:

$$R_S = 133.33(Z_L - 2R_P) \quad \text{(EQ. 13)}$$

<u>Node Equation</u> at HC55185 V_{RX} input

$$I_X = \frac{V_{RX}}{R} + \frac{V_{TX}}{R} \quad \text{(EQ. 14)}$$

Substitute Equation 10 into Equation 14

$$I_X = \frac{V_{RX}}{R} - \left(\frac{R_S \Delta I_M 30}{R8K}\right) \quad \text{(EQ. 15)}$$

<u>Loop Equation</u> at HC55185 feed amplifier and load

$$I_X R - V_{TR} + I_X R = 0 \quad \text{(EQ. 16)}$$

Substitute Equation 15 into Equation 16

$$V_{TR} = 2V_{RX} - \left(\frac{R_S \Delta I_M 60}{8K}\right) \quad \text{(EQ. 17)}$$

Substitute Equation 12 for R_S and $-V_{2W}/Z_L$ for ΔI_M into Equation 17.

$$V_{TR} = 2V_{RX} + \frac{Z_O V_{2W}}{Z_L} \quad \text{(EQ. 18)}$$

<u>Loop Equation</u> at Tip/Ring interface

$$V_{2W} - I_M 2R_P + V_{TR} = 0 \quad \text{(EQ. 19)}$$

Substitute Equation 18 into Equation 19 and combine terms.

$$V_{2W}\left[\frac{Z_L + Z_O + 2R_P}{Z_L}\right] = -2V_{RX} \quad \text{(EQ. 20)}$$

where:

V_{RX} = The input voltage at the V_{RX} pin.

V_{SA} = An internal node voltage that is a function of the loop current and the output of the Sense Amplifier.

I_X = Internal current in the SLIC that is the difference between the input receive current and the feedback current.

I_M = The AC metallic current.

R_P = A protection resistor (typical 49.9Ω).

R_S = An external resistor/network for matching the line impedance.

V_{TR} = The tip to ring voltage at the output pins of the SLIC.

V_{2W} = The tip to ring voltage including the voltage across the protection resistors.

Z_L = The line impedance.

Z_O = The source impedance of the device.

Receive Gain (V_{IN} to V_{2W})

4-wire to 2-wire gain is equal to the V_{2W} divided by the input voltage V_{IN}, reference Figure 2. The gain through the TCM38C17 is programmed to be 1.0 using Equation 21.

$$G_{(PCMIN-PWRO+\ )} = \frac{R_1 + R_2}{4\left(R_2 + \frac{R_1}{4}\right)} \quad \text{(EQ. 21)}$$

The input and output gain adjustments are discussed in detail in PCM CODEC / Filter Combo Family: Device Design-in and Application Data [1]. The maximum output (Gain =1) can be obtained by maximizing R_1 and minimizing R_2 (Figure 2). This can be done by letting R_1 = infinity and R_2 = 0, as shown in Figure 3.

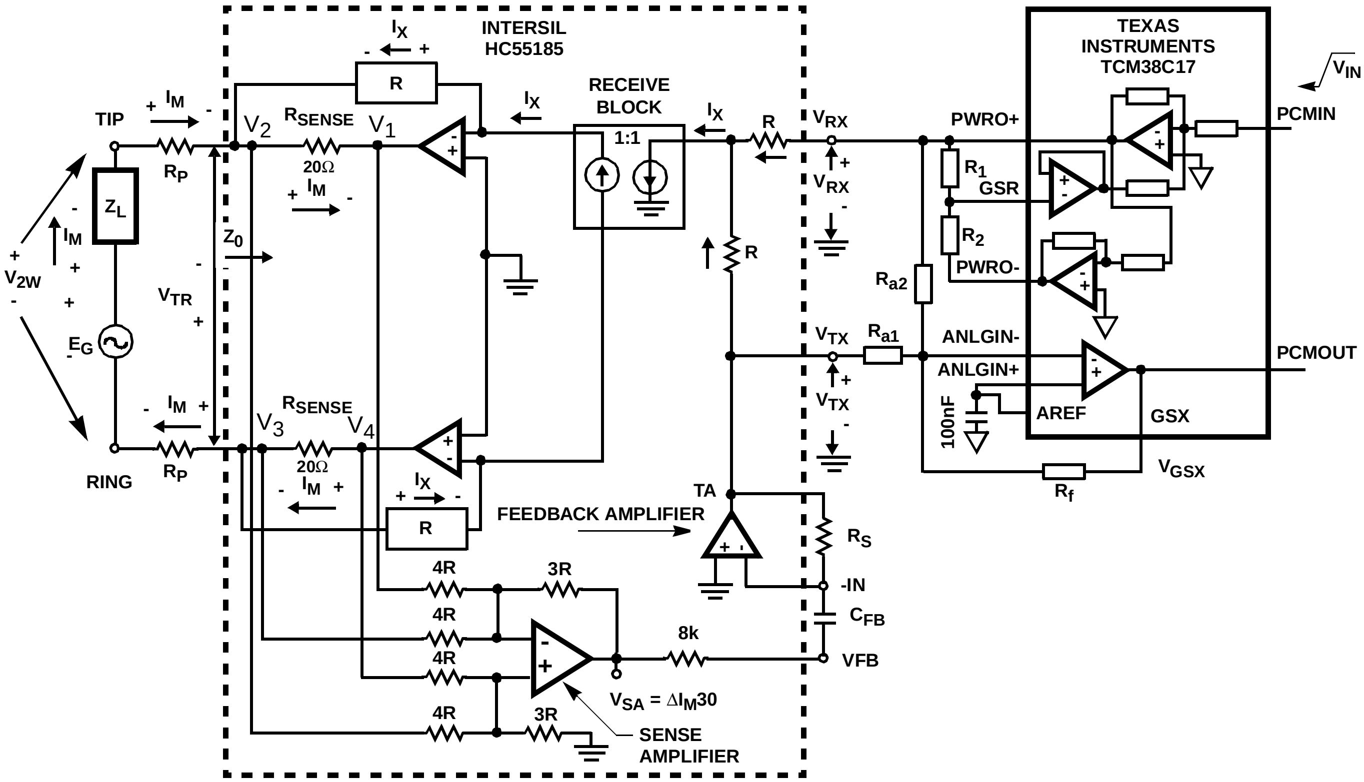

FIGURE 2. HC55185 SIMPLIFIED AC TRANSMISSION CIRCUIT AND TCM38C17

The receive gain is calculated using Equation 20 and the relationship $R_S = 133.33(Z_L\text{-}2R_P)$.

Equation 22 expresses the receive gain (V_{IN} to V_{2W}) in terms of network impedances, where $V_{IN} = V_{PCMIN} = V_{PWRO+} = V_{RX}$.

$$G_{4\text{-}2} = \frac{V_{2W}}{V_{IN}} = -2\frac{Z_L}{Z_L + Z_O + 2_{RP}} \qquad \text{(EQ. 22)}$$

Notice that the phase of the 4-wire to 2-wire signal is 180^o out of phase with the input signal.

Transmit Gain Across HC55185 (E_G to V_{TX})

The 2-wire to 4-wire gain is equal to V_{TX}/E_G with $V_{RX} = 0$, reference Figure 2.

Loop Equation

$$-E_G + Z_L I_M + 2R_P I_M - V_{TR} = 0 \qquad \text{(EQ. 23)}$$

From Equation 18 with $V_{RX} = 0$

$$V_{TR} = \frac{Z_O V_{2W}}{Z_L} \qquad \text{(EQ. 24)}$$

Substituting Equation 24 into Equation 23 and simplifying.

$$E_G = -V_{2W}\left[\frac{Z_L + 2R_P + Z_O}{Z_L}\right] \qquad \text{(EQ. 25)}$$

Substituting Equation 12 into Equation 10 and defining $\Delta I_M = -V_{2W}/Z_L$ results in Equation 26 for VTX.

$$V_{TX} = \frac{V_{2W}}{2}\left[\frac{Z_L - 2R_P}{Z_L}\right] \qquad \text{(EQ. 26)}$$

Combining Equations 25 and 26 results in Equation 27.

$$G_{2\text{-}4} = \frac{V_{TX}}{E_G} = -\frac{Z_L - 2R_P}{2(Z_L + 2R_P + Z_O)} = -\frac{Z_O}{2(Z_L + 2R_P + Z_O)} \qquad \text{(EQ. 27)}$$

A more useful form of the equation is rewritten in terms of V_{TX}/V_{2W}. A voltage divider equation is written to convert from E_G to V_{2W} as shown in Equation 28.

$$V_{2W} = \left(\frac{Z_O + 2_{RP}}{Z_L + Z_O + 2_{RP}}\right)E_G \qquad \text{(EQ. 28)}$$

Substituting $Z_L = Z_O + 2_{RP}$ and rearranging Equation 28 in terms of E_G results in Equation 29.

$$E_G = 2V_{2W} \qquad \text{(EQ. 29)}$$

Substituting Equation 29 into Equation 27 results in an equation for 2-wire to 4-wire gain that's a function of the synthesized input impedance of the SLIC and the protection resistors.

$$G_{2\text{-}4} = \frac{V_{TX}}{V_{2W}} = -\frac{Z_O}{(Z_L + 2R_P + Z_O)} \qquad \text{(EQ. 30)}$$

Notice that the phase of the 2-wire to 4-wire signal is out-of-phase with the input signal and when the protection resistors are set to zero, the transmit gain is -6dB.

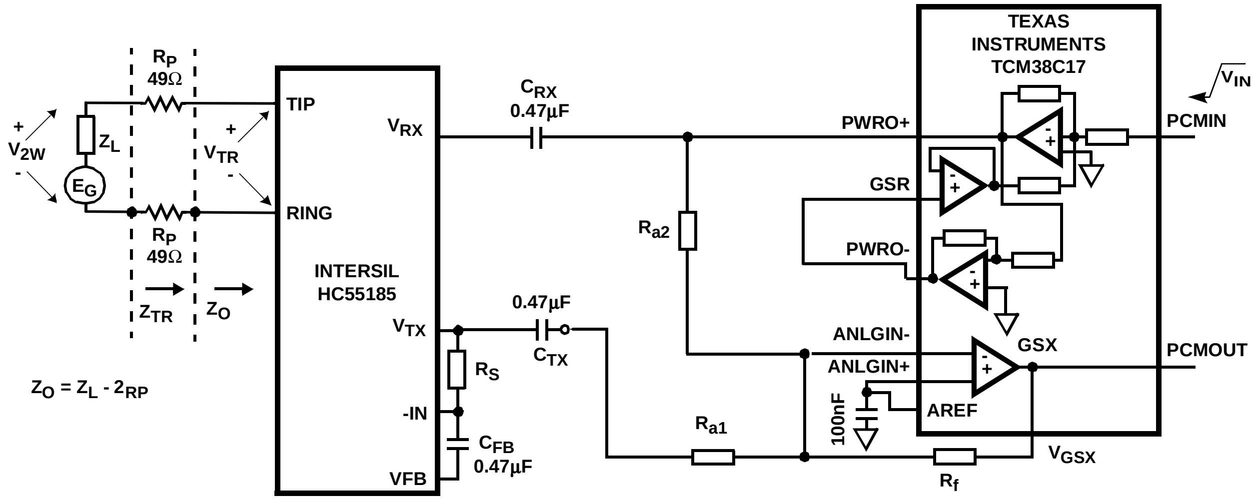

FIGURE 3. RECEIVE GAIN G(4-2), TRANSMIT GAIN (2-4) AND TRANSHYBRID BALANCE

Transmit Gain Across the System (V_{2W} to V_{PCMOUT})

2-wire to 4-wire gain is equal to the V_{PCMOUT} voltage divided by the 2-wire voltage V_{2W}, reference Figure 3.

$$G_{2-4} = \frac{V_{PCMOUT}}{V_{2W}} \quad \text{(EQ. 31)}$$

V_{PCMOUT} is only a function of V_{TX} and the feedback resistors R_{a1} and R_f Equation 32. This is because V_{IN} is considered ground for this analysis, thereby effectively grounding the V_{PWRO+} output.

$$V_{PCMOUT} = -V_{TX}\frac{R_f}{R_{a1}} \quad \text{(EQ. 32)}$$

An equation for the system transmit gain is achieved by substituting Equation 30 into Equation 32.

$$G_{2-4} = \frac{V_{PCMOUT}}{V_{2W}} = -\frac{V_{TX}}{V_{2W}}\frac{R_f}{R_{a1}} = \frac{Z_O}{(Z_L + 2R_P + Z_O)}\frac{R_f}{R_{a1}} \quad \text{(EQ. 33)}$$

To achieve aTransmit Gain of one (V_{PCMOUT}/V_{2W}), make $R_f = (Z_L+2R_P+Z_0)$ and $R_{a1} = Z_0$. Actual values of R_{a1} and R_f were multiplied by 100 to reduce loading effects on the GSX opamp.

Transhybrid Balance G(4-4)

Transhybrid balance is a measure of how well the input signal is canceled (that being received by the SLIC) from the transmit signal (that being transmitted from the SLIC to the CODEC). Without this function, voice communication would be difficult because of the echo.

The signals at V_{PWRO+} and V_{TX} (Figure 3) are opposite in phase. Transhybrid balance is achieved by summing two signals that are equal in magnitude and opposite in phase into the GSX amplifier.

Transhybrid balance is achieved by summing the PWRO+ signal with the output signal from the HC55185 when proper gain adjustments are made to match V_{PWRO+} and V_{TX} magnitudes.

For discussion purpose, the GSX amplifier is redrawn with the external resistors in Figure 4.

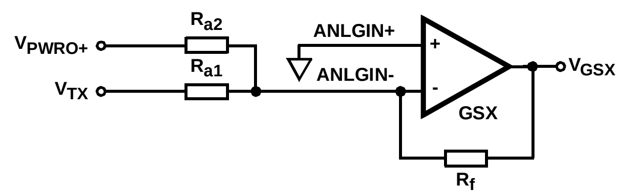

FIGURE 4. TRANSHYBRID BALANCE CIRCUIT

The gain through the GSX amplifier from V_{PWRO+} is set by resistors R_{a2} and R_f. The gain through the GSX amplifier from V_{TX} is set by resistors R_{a1} and R_f.

Transhybrid balance is achieved by adjusting the magnitude from both V_{PWRO+} and V_{TX} so their equal to each other.

Reference Design of the HC55185 and the TCM38C17 With a 600Ω Load

The design criteria is as follows:

- 4-wire to 2-wire gain (VP_{CMIN} to V_{2W}) equal 0dB
- 2-wire to 4-wire gain (V_{2W} to VP_{CMOUT}) equal 0dB
- Two Wire Return Loss greater than -30dB (200Hz to 4kHz)
- Rp = 49.9Ω

Figure 5 gives the reference design using the Intersil HC55185 and the Texas Instruments TCM38C17 Quad Combo. Also shown in Figure 5 are the voltage levels at specific points in the circuit.

Impedance Matching

The 2-wire impedance is matched to the line impedance Z_0 using Equation 1, repeated here in Equation 34.

$$R_S = 133.3 \bullet (Z_L - 2R_P) \quad \text{(EQ. 34)}$$

For a line impedance of 600Ω, R_S equals:

$$R_S = 133.3 \bullet (600 - 98) = 66.9k\Omega \quad \text{(EQ. 35}$$

The closest standard value for R_S is 66.5kΩ.

Transhybrid Balance (Z_L = 600Ω)

The internal GSX amplifier of the TCM38C17 is used to perform the transhybrid balance function.Transhybrid balance is achieved by summing two signals that are equal in magnitude and opposite in phase into the GSX amplifier. From Equation 33, repeated here in Equation 36, aTransmit Gain (V_{PCMOUT}/ V_{2W}) of one is achieved if we make $R_f = (Z_L + 2R_P + Z_0)$ and $R_{a1} = Z_0$.

$$G_{2-4} = \frac{V_{PCMOUT}}{V_{2W}} = -\frac{V_{TX}}{V_{2W}}\frac{R_f}{R_{a1}} = \frac{Z_O}{(Z_L + 2R_P + Z_O)}\frac{R_f}{R_{a1}} \quad \text{(EQ. 36)}$$

$$R_f = (Z_L + 2R_P + Z_O) = (600 + 98 + 502)(100) = 120k\Omega \quad \text{(EQ. 37)}$$

$$R_{a1} = (Z_O) = 502\Omega(100) = 50.2k\Omega \quad \text{(EQ. 38)}$$

Actual values of R_{a1} and R_f were multiplied by 100 to reduce loading effects on the GSX op-amp.

Closest standard value for R_f is 121.0kΩ
Closest standard value for R_{a1} is 49.9kΩ

The TCM38C17 receive gain is programmed to 1.0 by maximizing R_1 and minimizing R_2 resistor values (Figure 2).

The gain from PWRO+/V_{RX} through the SLIC at V_{TX} is 0.418 (Eq. 10 in the Intersil HC55185 data sheet, repeated here in Equation 39).

$$G_{4\text{-}4} = \frac{V_{TX}}{V_{RX}} = -\frac{Z_O}{(Z_L + 2R_P + Z_O)} \quad \text{(EQ. 39)}$$

To achieve transhybrid balance from the PWRO+ pin to PCMOUT set $R_{a2} = R_{a1}$ / 0.418.

$$R_{a2} = \left(\frac{49.9k\Omega}{0.418}\right) = 119.37k\Omega \quad \text{(EQ. 40)}$$

Closest standard value for R_{a2} = 118kΩ.

Specific Implementation for China

The design criteria for a China specific solution are as follows:

- Desired line circuit impedance is 200 + 680//0.1µF
- Receive gain (V_{2W}/V_{PCMIN}) is -3.5dB
- Transmit gain (V_{PCMOUT}/V_{2W}) is 0dB
- 0dBm0 is defined as 1mW into the complex impedance at 1020Hz
- R_p = 49.9Ω

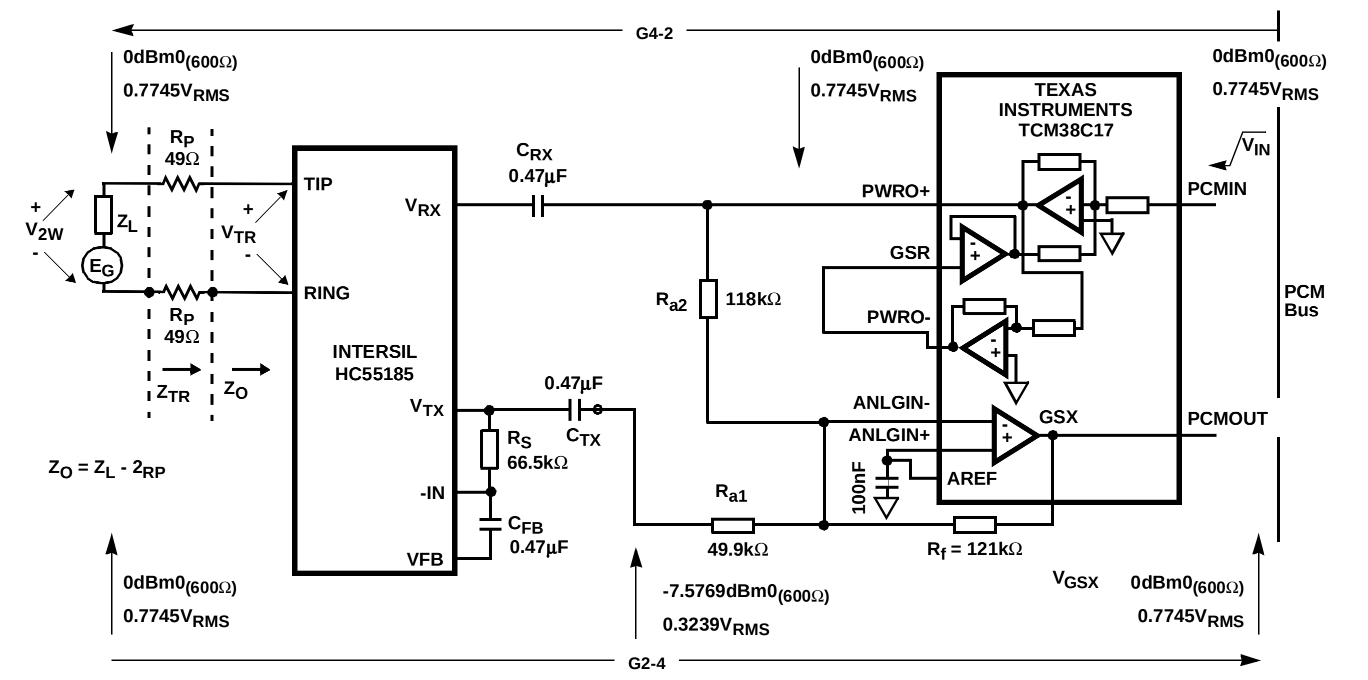

FIGURE 5. REFERENCE DESIGN OF THE HC55185 AND THE TCM38C17 WITH A 600Ω LOAD IMPEDANCE

Figure 6 gives the reference design using the Intersil HC55185 and theTexas Instruments TCM38C17 Quad Combo. Also shown in Figure 6 are the voltage levels at specific points in the circuit. These voltages will be used to adjust the gains of the network.

Adjustment to Get -3.5dBm0 at the Load Referenced to 600Ω

The voltage equivalent to 0dBm0 into 811Ω ($0dBm0_{(811\Omega)}$) is calculated using Equation 41 (811Ω is the impedance of complex China load at 1020Hz).

$$0dBm_{(811\Omega)} = 10\log\frac{V^2}{811(0.001)} = 0.90055V_{RMS} \quad \text{(EQ. 41)}$$

The gain referenced back to $0dBm0_{(600\Omega)}$ is equal to:

$$GAIN = 20\log\frac{0.90055V_{RMS}}{0.7745V_{RMS}} = 1.309dB \quad \text{(EQ. 42)}$$

The adjustment to get -3.5dBm0 at the load referenced to 600Ω is:

$$Adjustment = -3.5dBm0 + 1.309dBm0 = -2.19dB \quad \text{(EQ. 43)}$$

The voltage at the load (referenced to 600Ω) is given in Equation 44:

$$-2.19dBm_{(600\Omega)} = 10\log\frac{V^2}{600(0.001)} = 0.60196V_{RMS} \quad \text{(EQ. 44)}$$

Impedance Matching

The 2-wire impedance is matched to the line impedance Z_L using Equation 1, repeated here in Equation 45.

$$R_S = 133.3 \bullet (Z_L - 2R_P) \quad \text{(EQ. 45)}$$

For a line impedance of 200 + 680//0.1µF, R_S equals:

$$R_S = 133.3 \bullet \left(200 + \frac{680}{1 + j\omega 680(0.1)X10^{-6}} - (2)(49.9)\right) \quad \text{(EQ. 46)}$$

$$R_S = 133.3 \bullet (102\Omega) + \left[133.3 \bullet 680\Omega \parallel \frac{0.1\mu F}{133.3}\right] \quad \text{(EQ. 47)}$$

R_S = 13.6kΩ in series with the parallel combination of 90.6kΩ and 750pF (closest standard values are: R_S = 13.7kΩ, R_P = 90.9kΩ and C_P = 680pF).

To achieve a 4-wire to 2-wire gain (V_{PCMIN} to V_{2W}) that is equivalent to 0dBm(600Ω) at the complex load, the gain through the TCM38C17 (V_{PCMIN} to V_{PWRO+}) must equal -2.19dBm (0.60196V_{RMS}). The gain through the TCM38C17 will then equal -2.19dBm (0.60196V_{RMS}) divided by the input voltage 0dBm (0.7745V_{RMS}). This gain is equal to 0.777.

The gain through the TCM38C17 (V_{PCMIN} to V_{PWRO+}) is given in Equation 21 and repeated here in Equation 48.

$$G_{(PCMIN-PWRO)} = \frac{R_1 + R_2}{4\left(R_2 + \frac{R_1}{4}\right)} \quad \text{(EQ. 48)}$$

Setting the gain equal to 0.777 we can now determine the value of the gain setting resistors R_1 and R_2. Selecting the value of R_1 to be 49.9kΩ, R_2 is calculated to 5.27kΩ. (Note: the value of R_1 + R_2 should be greater than 10kΩ but less than 100kΩ.)

$$0.777 = \frac{R_1 + R_2}{4\left(R_2 + \frac{R_1}{4}\right)} \quad \text{(EQ. 49)}$$

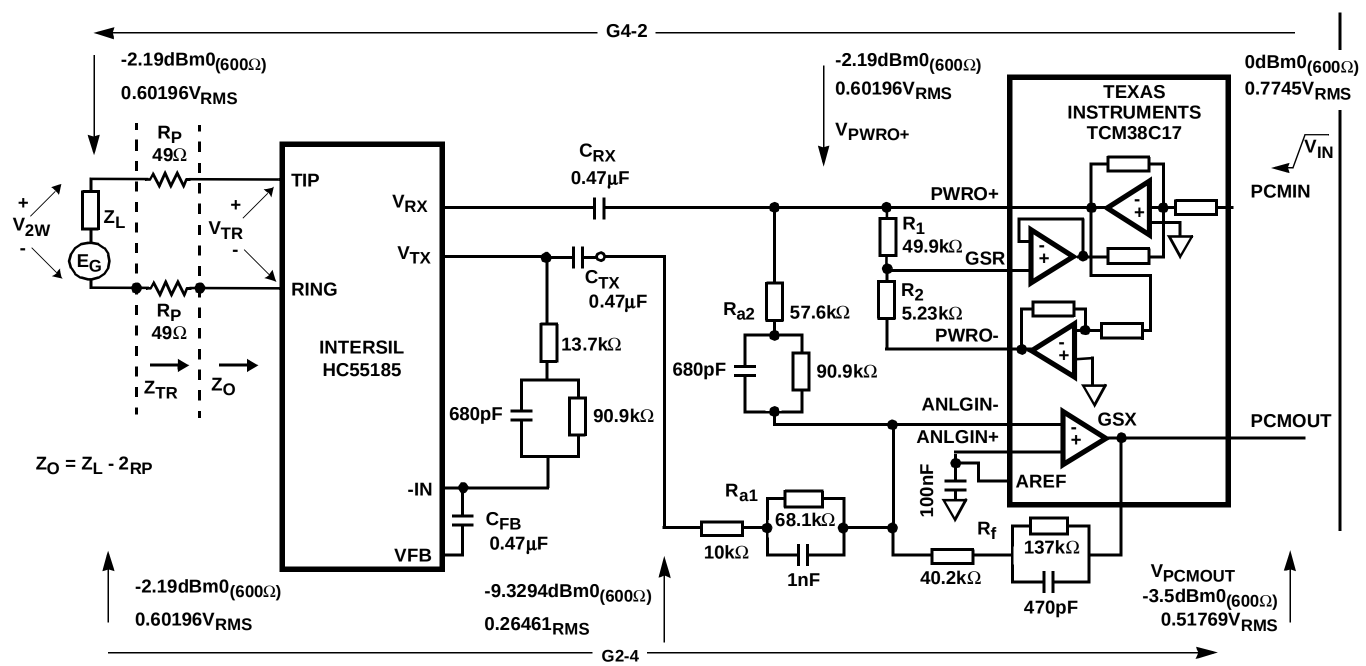

FIGURE 6. REFERENCE DESIGN OF THE HC55185 AND THE TCM38C17 WITH CHINA COMPLEX LOAD IMPEDANCE

$$R_2 = R_1\left(\frac{0.222}{2.108}\right) = 49.9k\Omega(0.105) = 5.27k\Omega \quad \text{(EQ. 50)}$$

The closest standard value for R_2 is 5.23kΩ.

Transhybrid Balance (Z_L= 200 + 680//0.1μF)

The internal GSX amplifier of the TCM38C17 is used to perform the transhybrid balance function.Transhybrid balance is achieved by summing two signals that are equal in magnitude and opposite in phase into the GSX amplifier. From Equation 33, repeated here in Equation 51, aTransmit Gain (V_{PCMOUT}/ V_{2W}) of one is achieved if we make $R_f = (Z_L+2R_P+Z_0)$ and $R_{a1} = Z_0$.

$$G_{2-4} = \frac{V_{PCMOUT}}{V_{2W}} = -\frac{V_{TX}}{V_{2W}}\frac{R_f}{R_{a1}} = \frac{Z_O}{(Z_L+2R_P+Z_O)}\frac{R_f}{R_{a1}} \quad \text{(EQ. 51)}$$

Note: $2R_P + Z_O = Z_L$

$$R_f = (Z_L + 2R_P + Z_O) = (2Z_L) = 2(200 + 680 \parallel 0.1\mu F) \quad \text{(EQ. 52)}$$

$$R_f = (400\Omega + 1.36k\Omega \parallel 50nF)100 = 40k\Omega + 136k\Omega \parallel 500pF \quad \text{(EQ. 53)}$$

$$R_{a1} = (Z_O) = (Z_L - 2R_P) = (100.2 + 680 \parallel 0.1\mu F) \quad \text{(EQ. 54)}$$

$$R_{a1} = (100.2\Omega + 680\Omega \parallel 0.1\mu F)100 = 10k\Omega + 68k\Omega \parallel 1nF \quad \text{(EQ. 55)}$$

Actual values of R_{a1} and R_f were multiplied by 100 to reduce loading effects on the GSX op-amp.

Closest standard values for R_f are: R_S = 40.2kΩ, R_P = 137kΩ and C_P = 470pF.

Closest standard values for R_{a1} are: R_S = 10kΩ, R_P = 68.1kΩ and C_P = 1nF.

The TCM38C17 receive gain is programmed to 1.0 by maximizing R_1 and minimizing R_2 resistor values (Figure 2).

The gain from PWRO+/V_{RX} through the SLIC at V_{TX} is 0.442 @ 1kHz (Z_L = 813, Z_O = 719) (Eq. 10 in the Intersil HC55185 data sheet, repeated here in Equation 56).

$$G_{4\text{-}4} = \frac{V_{TX}}{V_{RX}} = -\frac{Z_O}{(Z_L + 2R_P + Z_O)} \quad \text{(EQ. 56)}$$

To achieve transhybrid balance from the PWRO+ pin to PCMOUT set $R_{a2} = R_{a1}$ / 0.442.

$$R_{a2} = \left(\frac{10k\Omega + 68k\Omega \parallel 1nF}{0.442}\right) = (22.62k\Omega + 153.8k\Omega \parallel 442pF) \quad \text{(EQ. 57)}$$

Closest standard values for R_{a2} are: R_S = 22.6kΩ, R_P = 154kΩ and C_P = 470pF.

Reference

[1] Website
www.ti.com/sc/docs/psheets/abstract/apps/slwa006.htm

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