HC5503PRC SLIC and the Texas Instruments TP3057A Combined PCM CODEC and Filter

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Reference Design using the HC5503PRC SLIC and the Texas Instruments TP3057A Combined PCM CODEC and Filter

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). This requires that the physical resistance that is situated between the SLIC and the subscriber line, comprised of protection and/or sensing resistors, and the output resistance of the SLIC itself, be adapted to present an impedance to the subscriber line that varies with frequency. This is accomplished using feedback around the SLIC.

The purpose of this application note is to show a means of accomplishing this task for the HC5503PRC and Texas Instruments TP3057A Combo.

Discussed in this application note is the following:

- 2-wire 600Ω impedance matching
- 2-wire complex 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 HC5503PRC 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. It is also important for maintaining voice signal levels on long loops. Consider the equivalent circuit shown in Figure 1.

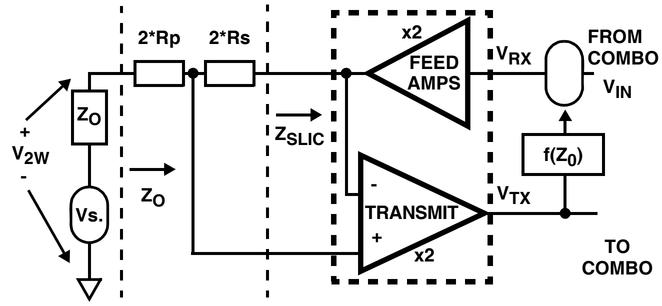

FIGURE 1. IMPEDANCE MATCHING BLOCK DIAGRAM

The circuitry inside the dotted box is representative of the SLIC feed and transmit amplifiers. The feed and transmit amplifiers pass the voice signals in the receive and transmit directions respectively. Without the feedback block $f(Z_0)$, the termination resistance at V_{2W} would equal the two protection resistors (R_P) and the two sense resistors (R_S), as the feed amplifiers present a very low output impedance to the subscriber line. The desired termination impedance at V_{2W} is Z_0. The feedback block $f(Z_0)$ matches the SLICs output impedance (Z_{SLIC}) plus the two protection resistors (R_P) and the two sense resistors (R_S) to the load (Z_0).

Impedance matching of the HC5503PRC is accomplished by making the SLIC's impedance (Z_{SLIC}, Figure 2) equal to the desired terminating impedance Z_0, minus the value of the protection and sense resistors. The desired impedance at the input to the SLIC is given in Equation 1.

$$Z_{SLIC} = Z_0 - 2 \times R_P - 2 \times R_S \quad \text{(EQ. 1)}$$

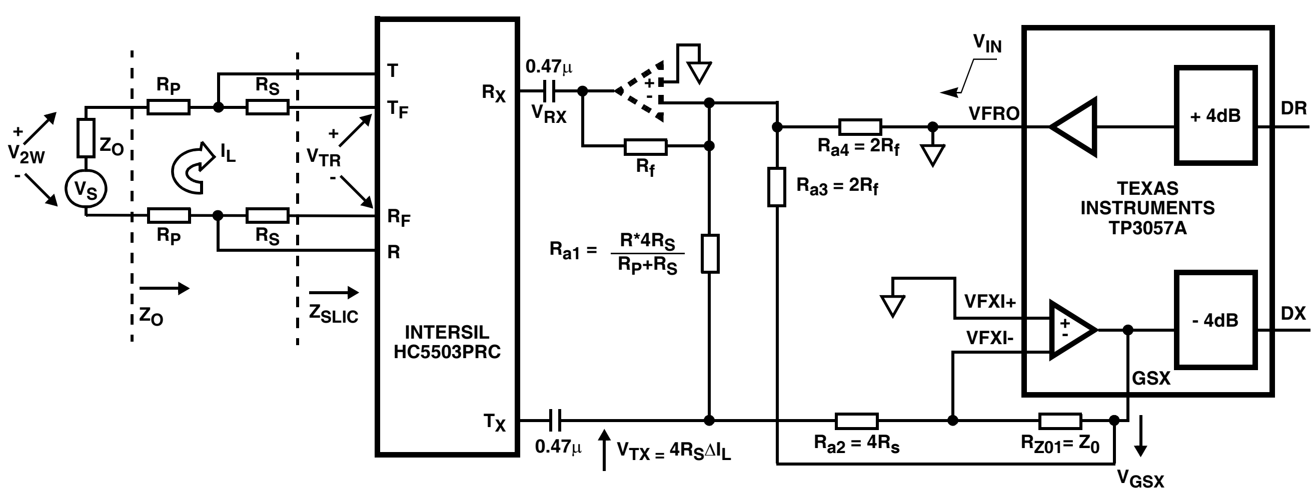

FIGURE 2. IMPEDANCE MATCHING

The AC loop current required to satisfy this condition is given in Equation 2.

$$\Delta I_L = \frac{V_{TR}}{(Z_0 - 2 \times R_P - 2 \times R_S)} \text{ at matching} \quad \text{(EQ. 2)}$$

The current calculated in Equation 2 is used as feedback to match the impedance of the SLIC and both protection and sense resistors to the load Z_0.

The output voltage of the SLIC (V_{TX}) is defined by design and given in Equation 3.

$$V_{TX} = 4R_S\Delta I_L \quad \text{(EQ. 3)}$$

Substituting for ΔI_L from Equation 2 into Equation 3 results in the voltage at the V_{TX} output that will be used to generate the required feedback.

$$V_{TX} = \frac{4R_S \times V_{TR}}{(Z_0 - 2 \times R_P - 2 \times R_S)} \quad \text{(EQ. 4)}$$

By design, V_{TR} is equal to 2 times the voltage at the receive input (R_X) Figure 2.

$$V_{TR} = 2 \times V_{RX} \quad \text{(EQ. 5)}$$

Substituting Equation 5 into Equation 4.

$$V_{TX} = \frac{4R_S \times 2 \times V_{RX}}{(Z_0 - 2 \times R_P - 2 \times R_S)} \quad \text{(EQ. 6)}$$

Solving Equation 6 for the voltage at V_{RX} as a function of V_{TX} (when matching the Z_{SLIC}, the two protection resistors (R_P) and the two sense resistors (R_S) to the load Z_O) is given in Equation 7.

$$\frac{V_{RX}}{V_{TX}} = \frac{(Z_0 - 2 \times R_p - 2 \times R_S)}{8 \times R_S} \quad \text{(EQ. 7)}$$

Equation 7 is the gain of the feedback circuit (output/input = V_{RX}/V_{TX}) used to match the impedance of the SLIC and both protection and sense resistors. Note: In Equation 7 it seemed logical to simplify the numerator by trying to combine Z_0 and the two subsequent terms together. In practice however, the impedance of the network you want to match (Z_0) cannot easily have 2*R_p and 2*R_s subtracted from it since the sum of these resistors is often larger than the value of the series resistance of the complex network.

Equation 7 is therefore rewritten in Equation 8.

$$\frac{V_{RX}}{V_{TX}} = \frac{Z_0}{8 \times R_S} - \frac{2 \times (R_P + R_S)}{8 \times R_S} \quad \text{(EQ. 8)}$$

Analysis of Equation 8 yields a 2 op amp feedback network. The first term has Z_O and no phase inversion. This requires the path to flow through 2 op amps and makes the matching of different complex loads easy. (i.e., can set Z_O in feedback network equal to the Z_O you want to match). The second term has a phase inversion and requires only one op amp in the feedback path.

Figure 2 shows the circuit required to achieve matching of the SLIC's impedance to the load Z_O. The voltage at V_{RX} is a function of V_{TX}, V_{GSX} ($V_{TX}R_{ZO1}/R_{a2}$) and V_{IN}.

The voltage at V_{RX} is determined via superposition. The circuit equation for the feedback network is given in Equation 9.

$$V_{RX} = -V_{TX}\frac{R_f}{R_{a1}} + \frac{V_{TX}R_{ZO1}Rf}{R_{a2}R_{a3}} - \frac{V_{IN}R_f}{R_{a4}} \quad \text{(EQ. 9)}$$

For impedance matching of the two wire side, we set V_{IN} equal to zero. This reduces Equation 9 to that shown in Equation 10.

$$V_{RX} = -V_{TX}\frac{R_f}{R_{a1}} + \frac{V_{TX}R_{ZO1}Rf}{R_{a2}R_{a3}} \quad \text{(EQ. 10)}$$

To achieve the desired matching of the circuit to the line impedance Z_O, we set our design Equation 8 equal to our circuit Equation 10. By inspection of the correct phase in Equations 8 and 10, we have Equations 11 and 12.

$$\frac{Z_0}{8 \times R_S} = \frac{R_{ZO1}Rf}{R_{a2}R_{a3}} \quad \text{(EQ. 11)}$$

$$\frac{2 \times (R_P + R_S)}{8 \times R_S} = \frac{R_f}{R_{a1}} \quad \text{(EQ. 12)}$$

Given: $R_f = R$, $R_{a3} = 2R$, $R_{ZO1} = Z_O$ Note: by making $R_{a3} = 2R_f$, the value of R_{a2} becomes $4R_S$ (Equation 13). This results in the 2-wire to 4-wire gain being equal to 1 (Equation 24 and Equation 25)

From Equation 11.

$$R_{a2} = 4R_S \quad \text{(EQ. 13)}$$

From Equation 12.

$$R_{a1} = \frac{R \times 4R_S}{R_P + R_S} \quad \text{(EQ. 14)}$$

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 3. The gains through the CODEC are not considered at this point.

$$A_{4W-2W} = \frac{V_{2W}}{V_{IN}} \qquad \text{(EQ. 15)}$$

The 2-wire voltage V_{2W} is determined by a loop equation and is given in Equation 16.

$$V_{2W} = (2R_P + 2R_S)\Delta I_L + V_{TR} \qquad \text{(EQ. 16)}$$

Combining Equation 5 and Equation 9, gives and expression for V_{TR} in terms of V_{RX}, as shown in Equation 17.

$$V_{TR} = 2V_{RX} = 2\left(-V_{TX}\frac{R_f}{R_{a1}} + \frac{V_{TX}R_{ZO1}Rf}{R_{a2}R_{a3}} - \frac{V_{IN}R_f}{R_{a4}}\right) \qquad \text{(EQ. 17)}$$

The voltage at V_{TR} is therefore a function of V_{TX} and V_{IN}. Note: Contribution from V_{GSX} (middle term in Equation 17) is zero due to the transhybrid circuit, reference section titled “Transhybrid Balance G(4-4)”.

This reduces Equation 17 to Equation 18.

$$V_{TR} = 2V_{RX} = -2\left(V_{TX}\frac{R_f}{R_{a1}} + \frac{V_{IN}R_f}{R_{a4}}\right) \qquad \text{(EQ. 18)}$$

Substituting $4R_S\Delta I_L$ (Equation 3) for V_{TX} in Equation 18 and combining this with Equation 16, results in an equation for V_{2W} in terms of: ΔI_L, the external resistors and the input voltage V_{IN} (Equation 19).

$$V_{2W} = (2R_P + 2R_S)\Delta I_L - 8R_S\Delta I_L\frac{R_f}{R_{a1}} - 2\frac{V_{IN}R_f}{R_{a4}} \qquad \text{(EQ. 19)}$$

Ohms law defines ΔI_L as being equal to $-V_{2W}/Z_O$. Substituting $-V_{2W}/Z_O$ for ΔI_L in Equation 19 gives Equation 20.

$$V_{2W} = -(2R_P + 2R_S)\frac{V_{2W}}{Z_O} + 8R_S\frac{V_{2W}}{Z_O}\frac{R_f}{R_{a1}} - 2\frac{V_{IN}R_f}{R_{a4}} \qquad \text{(EQ. 20)}$$

Equation 20 can be rearranged to solve for the 4-wire to 2-wire gain $V2W/V_{IN}$, as shown in Equation 21.

$$A_{4W-2W} = \frac{V_{2W}}{V_{IN}} = -\left(\frac{2R_f}{R_{a4}}\right) \times \frac{R_{a1}Z_O}{R_{a1}(2R_P + 2R_S) + R_{a1}Z_O - 8R_SR_f} \qquad \text{(EQ. 21)}$$

Given: R_f=100kΩ, R_{a4}=200kΩ, R_{a1}=267kΩ, Z_O=600Ω, R_S=100Ω, R_P=50Ω.

Note: By making Ra4 equal to 2Rf the 4-wire to 2-wire gain becomes -1.

Transmit Gain across HC5503PRC (V_{2W} to V_{TX})

The output voltage of the SLIC (V_{TX}) was defined in Equation 3 as being equal to $4R_S\Delta I_L$. ΔI_L is equal to twice the input voltage ($2V_{RX}$) divided by the total loop resistance as shown in Figure 4. If the load impedance is 600Ω, then the gain across the HC5503PRC is 2/3 the input voltage V_{RX}. Likewise, if the load impedance is 811Ω, (next example with a complex load) then the gain across the HC5503PRC is 400/811 times the input voltage V_{RX}.

Transmit Gain (V_{2W} to V_{GSX})

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

$$A_{2W-4W} = \frac{V_{GSX}}{V_{2W}} \qquad \text{(EQ. 22)}$$

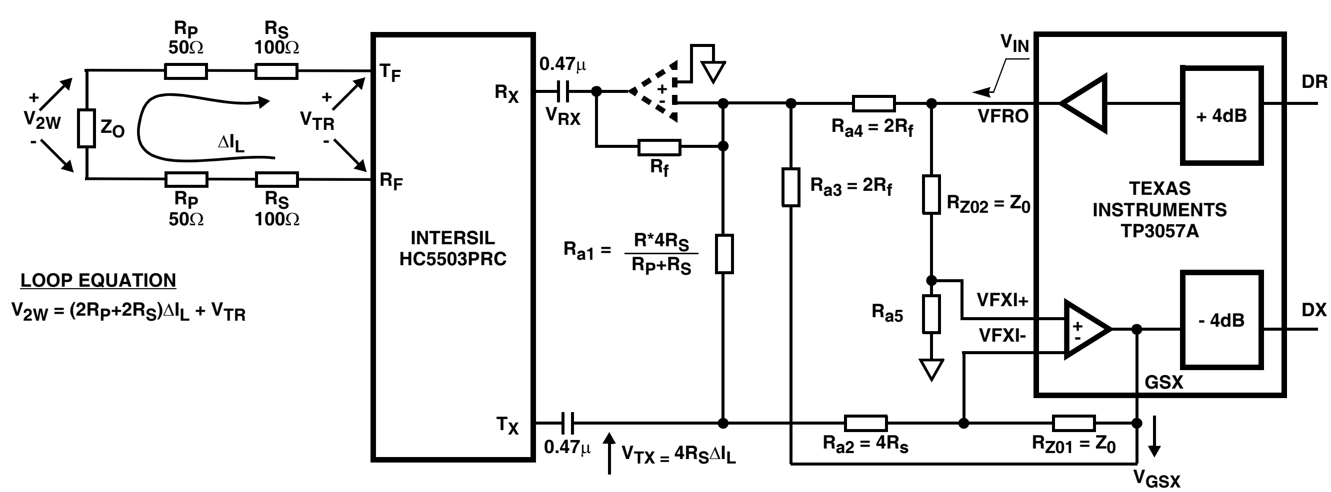

FIGURE 3. RECEIVE GAIN G(4-2), TRANSMIT GAIN (2-4) AND TRANSHYBRID BALANCE (FEEDBACK CIRCUIT ONLY)

$$A_{2-4} = \frac{V_{2W}}{V_{TX}}$$

$$V_{TX} = 4R_S\Delta I_L$$

$$\Delta I_L = \frac{2V_{RX}}{R_{TOTAL}}$$

$$\Delta I_L = \frac{V_{RX}}{600\Omega}$$

$$V_{TX} = 4R_S\Delta I_L = 400\frac{V_{RX}}{600\Omega}$$

$$V_{TX} = \frac{2}{3}V_{RX}$$

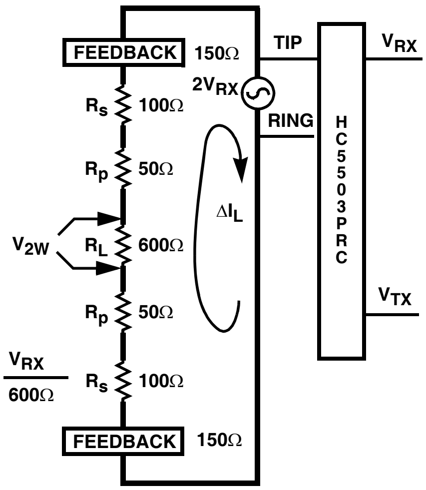

FIGURE 4. TRANSMIT GAIN ACROSS HC5503PRC (V_{2W} to V_{TX})

V_{GSX} is only a function of V_{TX} and the feedback resistors R_{a2} and R_{ZO1} Equation 23. This is because V_{IN} is considered ground for this analysis, thereby effectively grounding the positive terminal of the GSX OpAmp.

$$V_{GSX} = -V_{TX}\frac{R_{ZO1}}{R_{a2}} \quad \text{(EQ. 23)}$$

Substituting Equation 3 for V_{TX} and ΔI_L for $-V_{2W}/Z_O$ into Equation 23, V_{GSX} equals:

$$V_{GSX} = 4R_S\frac{V_{2W}}{Z_O}\left(\frac{R_{ZO1}}{R_{a2}}\right) \quad \text{(EQ. 24)}$$

Z_O is equal to R_{ZO1} (actual values of R_{ZO1} and Ra2 were multiplied by 1000 to reduce loading effects on the opamps). Simplifying Equation 24 and assuming R_{a2}=4RS from Equation 13 results in Equation 25.

$$A_{2W-4W} = \frac{V_{GSX}}{V_{2W}} = \left(\frac{4R_S}{4R_S}\right) = 1 \quad \text{(EQ. 25)}$$

The transmit gain 2-wire to 4-wire is equal to one.

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_{IN} and V_{TX} (Figure 3) are in phase. If V_{IN} and V_{TX} are summed together with the correct magnitudes at the input to the Combo transmit GSX OpAmp, they will cancel out and not be present at the V_{GSX} output.

The circuit in Figure 5 has been set up so that the SLIC matches the load impedance and that both G(4-2) and G(2-4) are adjusted to be 1.0 and flat over frequency.

The GSX OpAmp in the CODEC is configured as a differential amplifier with its output defined in Equation 26.

$$V_{GSX} = V_{IN}\frac{R_{a5}}{R_{a5}+R_{Z02}}\left(\frac{R_{a2}+R_{Z01}}{R_{a2}}\right) - V_{TX}\frac{R_{Z01}}{R_{a2}} \quad \text{(EQ. 26)}$$

The values of R_{a2}, $R_{a5,}$ R_{ZO1} and R_{ZO2} should be scaled by 1000 to minimize the effects of parallel resistance on the gain adjustment resistor R2 (Figure 5). Resistors R1 and R2 adjust the gain of the input signal from the TP3057A to account for the +4dB gain in the receive path. Scaling of a complex load is shown in EQ 27.

$$R_{ZO1} \text{or} R_{ZO2} = 100(\text{Resistive}) + \frac{\text{Reactive}}{100} \quad \text{(EQ. 27)}$$

Note: When matching a complex impedance some impedance models (900+2.15µF, K=100) will cause the OpAmp feedback to be open at DC currents, bringing the OpAmp to an output rail. A resistor with a value of about 10 times the reactance of the capacitor (21.6nF) at the low frequency of interest (200Hz for example) can be placed in parallel with the capacitor in order to solve the problem (368kΩ for a 21.6nF capacitor).

Reference Design of the HC5503PRC and the TP3057A with a 600Ω Load Impedance

The design criteria is as follows:

- 4-wire to 2-wire gain (DR to V_{2W}) equal 0dB
- 2-wire to 4-wire gain (V_{2W} to D_X) equal 0dB
- Two Wire Return Loss greater than -30dB (200Hz to 4kHz)

R_p = 50, R_s = 100.

Figure 5 gives the reference design using the Intersil HC5503PRC SLIC and the Texas Instruments TP3057A combined PCM CODEC and filter. Also shown in Figure 5 are the voltage levels at specific points in the circuit. These voltages will be used to adjust the gains of the network.

Impedance Matching

For impedance matching of the 2-wire side we set the input voltage at DR equal to zero. This effectively grounds the VFXI+ input of the GSX amplifier. To achieve a 2-wire to 4-wire gain (V_{2W} to DX) of 0dB we need to increase the gain of the GSX amplifier to overcome the -4dB loss in the TP3057A. The required gain is found by using Equation 24, repeated here for convenience in Equation 28.

$$V_{GSX} = 4R_S\frac{V_{2W}}{Z_O}\left(\frac{R_{ZO1}}{R_{a2}}\right) \quad \text{(EQ. 28)}$$

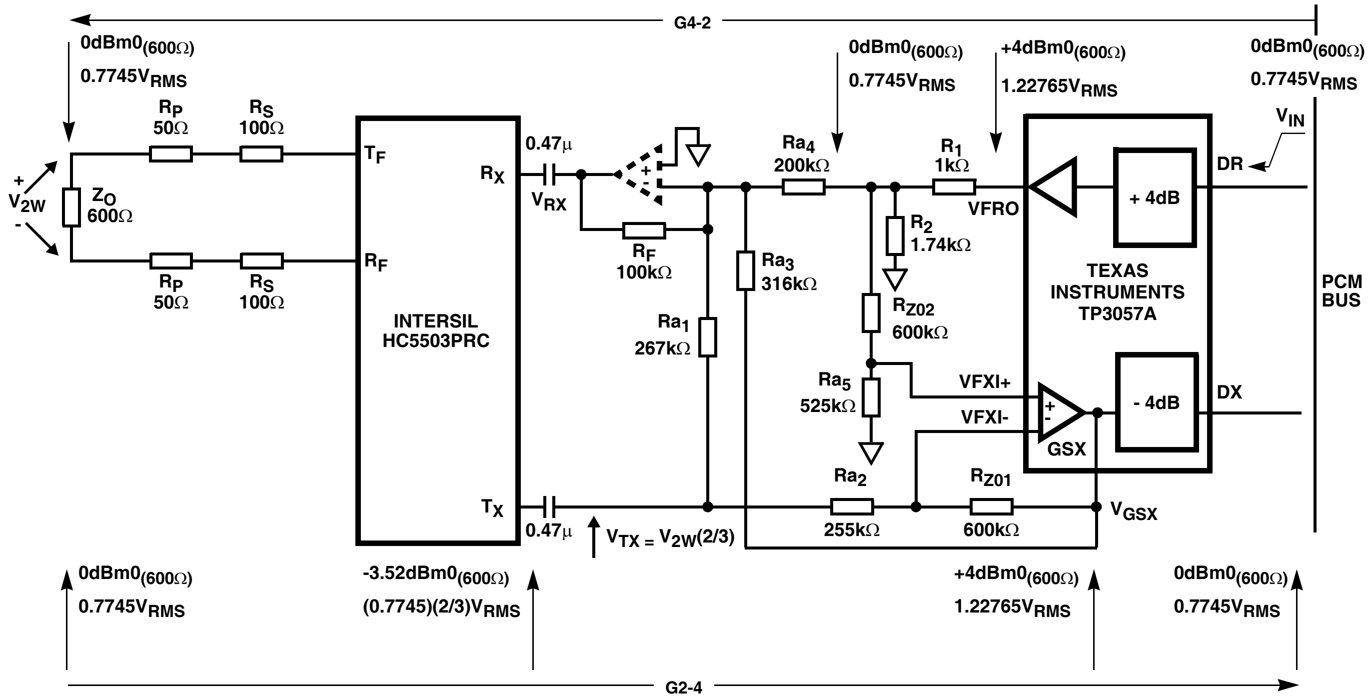

FIGURE 5. REFERENCE DESIGN OF THE HC5503PRC AND THE TP3057A WITH A 600Ω LOAD IMPEDANCE

Substituting the required voltage levels (Figure 5) for V_{GSX} (1.2276) and V_{2W} (0.7745) and rearranging to solve for R_{a2} results in Equation 29. Where: V_{GSX} / V_{2W} =1.585, and Z_0=R_{Z01}

$$R_{a2} = \frac{400}{1.585} = 252.3 \quad \text{(EQ. 29)}$$

The value of R_{a2} needs to be scaled by 1000 to minimize the effects of the parallel resistance R_{Z02} and R_{a5} on the gain adjustment resistor R2.

The nearest standard value for R_{a2} is 255kΩ.

R_{a3} needs to increase by (1.585) to maintain the same feedback for impedance matching Equation 30.

$$R_{a3} = (200\text{k}\Omega)(1.585) = 317\text{k}\Omega \quad \text{(EQ. 30)}$$

The closest standard value is for R_{a3} is 316kΩ.

To achieve a 4-wire to 2-wire gain (DR to V_{2W}) equal to 0dB we need to decrease the input to the feedback circuit from the VFRO pin to account for the +4dB increase in the TP3057A. A simple voltage divider will decrease the 1.2276 volt input down to the required 0.7745 volts Equation 31.

$$0.7745 = \frac{R_2}{R_2 + R_1} 1.2276 \quad \text{(EQ. 31)}$$

Rearranging to solve for R_2 results in Equation 32.

$$R_2 = R_1(1.709) \quad \text{(EQ. 32)}$$

If R_1 equals 1kΩ then R_2 equals 1.709kΩ. The closest standard value for R_2 is 1.74kΩ.

Transhybrid Balance (Z_L = 600Ω)

The internal GSX amplifier of the TP3057A is used to perform the transhybrid balance function. For discussion purpose, the GSX amplifier is redrawn with the external resistors in Figure 6. The transfer function of the amplifier is given in Equation 33 and Equation 34.

$$V_{OUT} = V2\frac{R_{a5}}{R_{a5} + R_{Z02}}\left(\frac{R_{Z01} + R_{a2}}{R_{a2}}\right) - V1\frac{R_{Z01}}{R_{a2}} \quad \text{(EQ. 33)}$$

$$V_{OUT} = V2\frac{R_{a5}}{R_{a5} + 600\text{K}\Omega}\left(\frac{600\text{K}\Omega + 255\text{K}\Omega}{255\text{K}\Omega}\right) - V1\frac{600\text{K}\Omega}{255\text{K}\Omega} \quad \text{(EQ. 34)}$$

V1 is equal to $(0.7745V_{RMS})(2/3)$. V2 is equal to $0.7745V_{RMS}$ $(0dBm0_{(600\Omega)})$. V_{OUT} is equal to zero. The results of rearranging Equation 34 to solve for Ra5 and substituting in the values for V_1 and V_2 are shown in Equation 35.

$$R_{a5} = \frac{941.17\text{k}\Omega}{(3.352 - 1.568)} = 527.47\text{k}\Omega \quad \text{(EQ. 35)}$$

Closest standard value for Ra5 is 525kΩ.

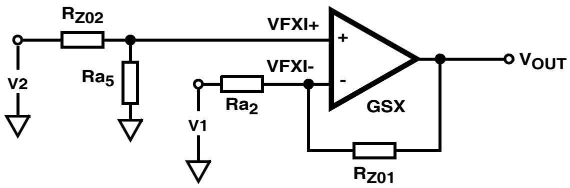

FIGURE 6. TRANSHYBRID BALANCE CIRCUIT

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_{DR}) is -3.5dB
- Transmit gain (V_{DX}/V_{2W}) is 0dB
- 0dBm0 is defined as 1mW into the complex impedance at 1020Hz

R_p = 50, R_S = 100.

Figure 7 gives the reference design using the Intersil HC5503PRC SLIC and the Texas Instruments TP3057A combined PCM CODEC and filter. Also shown in Figure 7 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 36. China complex load @ 1020Hz is equal to 811Ω .

$$0\text{dBm}_{(811\Omega)} = 10\log\frac{V^2}{811(0.001)} = 0.90055V_{RMS} \quad \text{(EQ. 36)}$$

The gain referenced back to 0dBm0$_{(600\Omega)}$ is equal to:

$$\text{GAIN} = 20\log\frac{0.90055V_{RMS}}{0.7745V_{RMS}} = 1.309\text{dB} \quad \text{(EQ. 37)}$$

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

$$\text{Adjustment} = -3.5\text{dBm0} + 1.309\text{dBm0} = -2.19\text{dB} \quad \text{(EQ. 38)}$$

The voltage at the load (referenced to 600Ω) is given in Equation 39.

$$-2.19\text{dBm}_{(600\Omega)} = 10\log\frac{V^2}{600(0.001)} = 0.60196V_{RMS} \quad \text{(EQ. 39)}$$

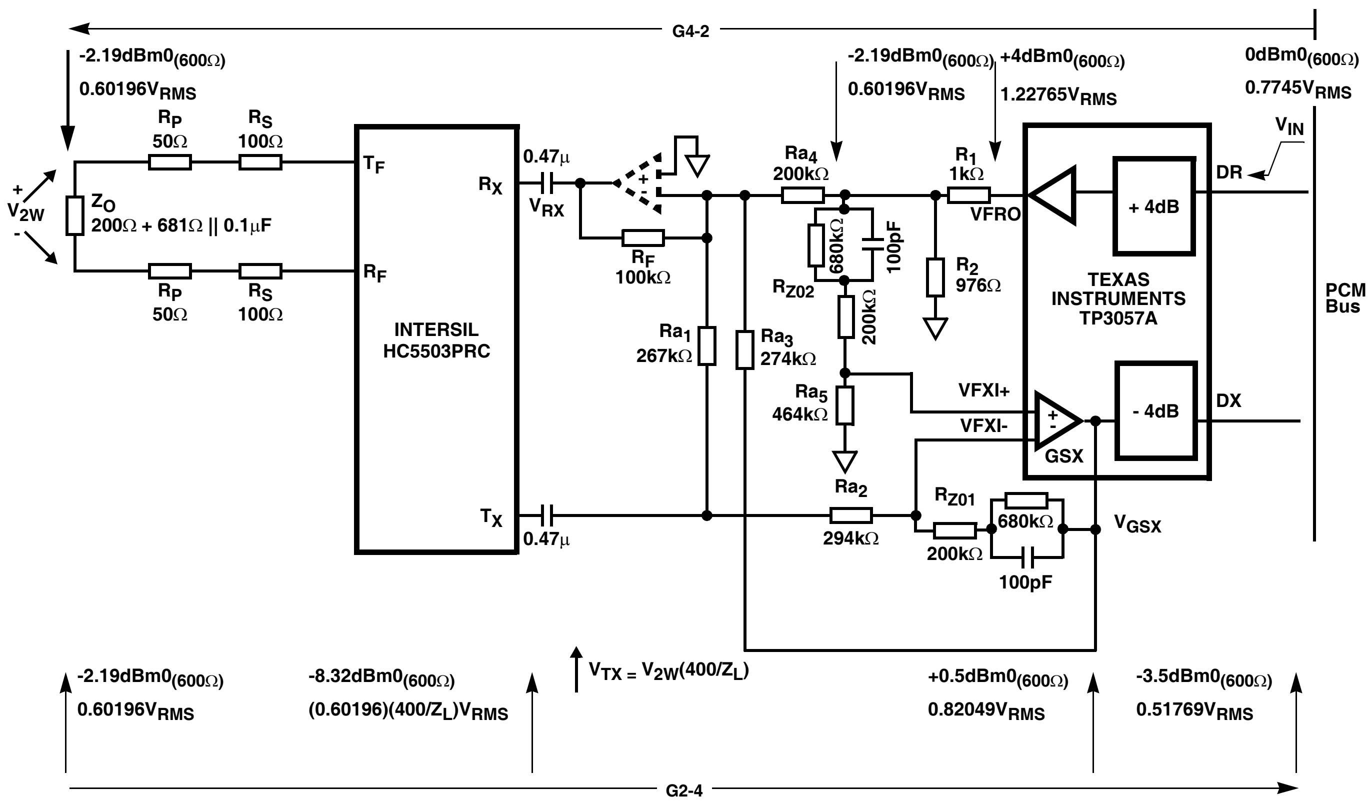

FIGURE 7. REFERENCE DESIGN OF THE HC5503PRC AND THE TP3057A WITH CHINA COMPLEX LOAD IMPEDANCE

Impedance Matching

For impedance matching of the 2-wire side we set the input voltage at DR equal to zero. This effectively grounds the VFXI+ input of the GSX amplifier. To achieve a 2-wire to 4-wire gain (V_{2W} to DX) of 0dB we need to increase the gain of the GSX amplifier to overcome the -4dB loss in the TP3057A. The required gain is found by using Equation 24, repeated here for convenience in Equation 40.

$$V_{GSX} = 4R_S\frac{V_{2W}}{Z_O}\left(\frac{R_{ZO1}}{R_{a2}}\right) \qquad \text{(EQ. 40)}$$

Substituting the required voltage levels (Figure 7) for V_{GSX} (0.82049) and V_{2W} (0.60196) and rearranging to solve for R_{a2} results in Equation 41. Where: V_{GSX} / V_{2W} =1.363, and Z_0=R_{Z01}:

$$R_{a2} = \frac{400}{1.363} = 293.47 \qquad \text{(EQ. 41)}$$

The value of R_{a2} needs to be scaled by 1000 to minimize the effects of parallel resistance on the gain adjustment resistor R2. The nearest standard value for R_{a2} is 294kΩ.

R_{a3} needs to increase by (1.363) to maintain the same feedback for impedance matching Equation 42.

$$R_{a3} = (200k\Omega)(1.363) = 272.6k\Omega \qquad \text{(EQ. 42)}$$

The closest standard value is for R_{a3} is 274kΩ.

To achieve a 4-wire to 2-wire gain (DR to V_{2W}) equal to 0dB we need to decrease the input to the feedback circuit from the VFRO pin to account for the +4dB increase in the TP3057A. A simple voltage divider will decrease the 1.2276 volt input down to the required 0.60196 volts Equation 43.

$$0.60196 = \frac{R_2}{R_2 + R_1}1.2276 \qquad \text{(EQ. 43)}$$

Rearranging to solve for R_2 results in Equation 44.

$$R_2 = R_1(0.9621) \qquad \text{(EQ. 44)}$$

If R_1 equals 1kΩ then R_2 equals 962.1Ω.

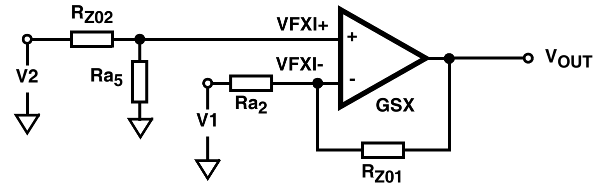

FIGURE 8. TRANSHYBRID BALANCE CIRCUIT

The closest standard value for R_2 is 976Ω.

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

The internal GSX amplifier of the TP3057A is used to perform the transhybrid balance function. For discussion purpose, the GSX amplifier is redrawn with the external resistors in Figure 8. The transfer function of the amplifier is given in Equation 45.

$$V_{OUT} = V2\frac{R_{a5}}{R_{a5} + R_{Z02}}\left(\frac{R_{Z01} + R_{a2}}{R_{a2}}\right) - V1\frac{R_{Z01}}{R_{a2}} \qquad \text{(EQ. 45)}$$

The impedance of the series parallel complex China load is equal to (multiplied by 1000):

$$R_{Z01} = R_{Z02} = 771k - j249k \qquad \text{(EQ. 46)}$$

Setting V_{OUT} equal to zero, best transhybrid balance, and rearranging the equation we get Equation 47.

$$V2\frac{R_{a5}}{R_{a5} + 771k - j249k}\left(\frac{771k - j249k + 294k\Omega}{294k\Omega}\right) = V1\frac{771k - j249k}{294k\Omega} \qquad \text{(EQ. 47)}$$

V1 is equal to (0.60196V_{RMS})(400/811) and V2 is equal to 0.60196V_{RMS}. The results of rearranging Equation 47 to solve for Ra5 and substituting in the values for V1 and V2 is shown in Equation 48.

$$R_{a5} = \frac{273k - j114k}{0.636 - j0.0302} = 420k - j199k = 465k\angle{-25.3^\circ} \qquad \text{(EQ. 48)}$$

Closest standard value for Ra5 is 464kΩ.

Specific Implementation for Australia

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

- Desired line circuit impedance is 220 + 820//120nF
- Receive gain (V_{2W}/V_{DR}) is -3.5dB
- Transmit gain (V_{DX}/V_{2W}) is 0dB
- 0dBm0 is defined as 1mW into the complex impedance at 1020Hz

$R_p = 50$, $R_S = 100$.

Figure 9 gives the reference design using the Intersil HC5503PRC SLIC and the Texas Instruments TP3057A combined PCM CODEC and filter. Also shown in Figure 9 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 887Ω ($0dBm0_{(887\Omega)}$) is calculated using Equation 49. Australia complex load @ 1020Hz is equal to 887Ω .

$$0dBm_{(887\Omega)} = 10\log\frac{V^2}{887(0.001)} = 0.936483V_{RMS} \quad \text{(EQ. 49)}$$

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

$$GAIN = 20\log\frac{0.936483V_{RMS}}{0.7745V_{RMS}} = 1.64957dB \quad \text{(EQ. 50)}$$

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

$$Adjustment = -3.5dBm0 + 1.649dBm0 = -1.85\ dB \quad \text{(EQ. 51)}$$

The voltage at the load (referenced to 600Ω) is given in Equation 52.

$$-1.85dBm_{(600\Omega)} = 10\log\frac{V^2}{600(0.001)} = 0.625971V_{RMS} \quad \text{(EQ. 52)}$$

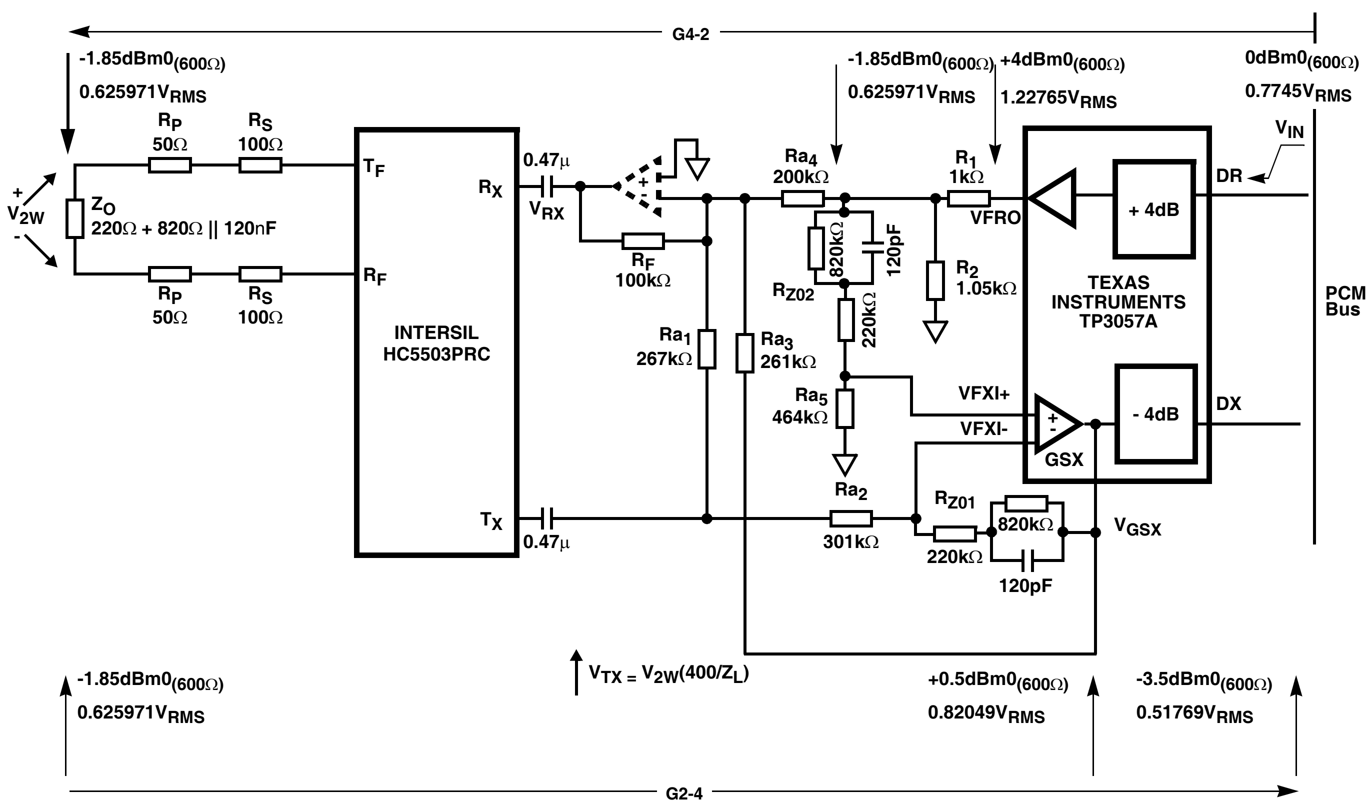

FIGURE 9. REFERENCE DESIGN OF THE HC5503PRC AND THE TP3057A WITH Australia COMPLEX LOAD IMPEDANCE

Impedance Matching

For impedance matching of the 2-wire side we set the input voltage at DR equal to zero. This effectively grounds the VFXI+ input of the GSX amplifier. To achieve a 2-wire to 4-wire gain (V_{2W} to DX) of 0dB we need to increase the gain of the GSX amplifier to overcome the -4dB loss in the TP3057A. The required gain is found by using Equation 24, repeated here for convenience in Equation 53.

$$V_{GSX} = 4R_S\frac{V_{2W}}{Z_O}\left(\frac{R_{ZO1}}{R_{a2}}\right) \quad \text{(EQ. 53)}$$

Substituting the required voltage levels (Figure 9) for V_{GSX} (0.82049) and V_{2W} (0.625971) and rearranging to solve for R_{a2} results in Equation 54. Where: V_{GSX} / V_{2W} =1.311, and Z_0=R_{Z01}:

$$R_{a2} = \frac{400}{1.311} = 305.17 \quad \text{(EQ. 54)}$$

The value of R_{a2} needs to be scaled by 1000 to minimize the effects of parallel resistance on the gain adjustment resistor R2. The nearest standard value for R_{a2} is 301kΩ.

R_{a3} needs to increase by (1.311) to maintain the same feedback for impedance matching Equation 55.

$$R_{a3} = (200k\Omega)(1.311) = 262.2k\Omega \quad \text{(EQ. 55)}$$

The closest standard value is for R_{a3} is 261kΩ.

To achieve a 4-wire to 2-wire gain (DR to V_{2W}) equal to 0dB we need to decrease the input to the feedback circuit from the VFRO pin to account for the +4dB increase in the TP3057A. A simple voltage divider will decrease the 1.2276 volt input down to the required 0.625971 volts Equation 56.

$$0.625971 = \frac{R_2}{R_2 + R_1}1.2276 \quad \text{(EQ. 56)}$$

Rearranging to solve for R_2 results in Equation 44.

$$R_2 = R_1(1.04037) \quad \text{(EQ. 57)}$$

If R_1 equals 1kΩ then R_2 equals 1.04Ω.

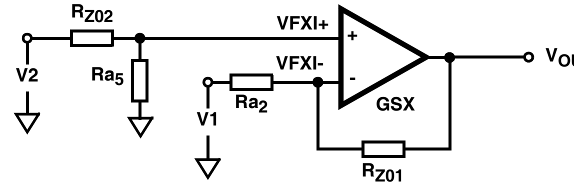

FIGURE 10. TRANSHYBRID BALANCE CIRCUIT

The closest standard value for R_2 is 1.05kΩ.

Transhybrid Balance (Z_L = 220 + 820//120nF)

The internal GSX amplifier of the TP3057A is used to perform the transhybrid balance function. For discussion purpose, the GSX amplifier is redrawn with the external resistors in Figure 10. The transfer function of the amplifier is given in Equation 58.

$$V_{OUT} = V2\frac{R_{a5}}{R_{a5} + R_{Z02}}\left(\frac{R_{Z01} + R_{a2}}{R_{a2}}\right) - V1\frac{R_{Z01}}{R_{a2}} \quad \text{(EQ. 58)}$$

The impedance of the series parallel complex Australia load is equal to (multiplied by 1000):

$$R_{Z01} = R_{Z02} = 807k - j370k \quad \text{(EQ. 59)}$$

Setting V_{OUT} equal to zero, best transhybrid balance, and rearranging the equation we get Equation 60.

$$V2\frac{R_{a5}}{R_{a5} + 807k - j370k}\left(\frac{807k - j370k + 301k\Omega}{301k\Omega}\right) = V1\frac{807k - j370k}{301k\Omega} \quad \text{(EQ. 60)}$$

V1 is equal to (0.625971V_{RMS})(400/887) and V2 is equal to 0.625971V_{RMS}. The results of rearranging Equation 60 to solve for Ra5 and substituting in the values for V1 and V2 is shown in Equation 61.

$$R_{a5} = \frac{261k - j156k}{0.659 - j0.0368} = 382k - j258k = 461k\angle{-34^\circ} \quad \text{(EQ. 61)}$$

Closest standard value for Ra5 is 464kΩ.

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