Integrated bidirectional bridge with dual RMS detectors for RF power and return-loss measurement

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Directional couplers are used in many applications to sense RF power, and may appear at multiple points in a signal chain. Inline RF power and return loss measurements are typically implemented using directional couplers and RF power detectors.

In Figure 1, a bidirectional coupler is used in a radio or test and measurement application to monitor transmitted and reflected RF power. It's also sometimes desirable to have RF power monitoring embedded in a circuit, a good example being when two or more sources are switched into the transmission path, either using an RF switch or external cables.

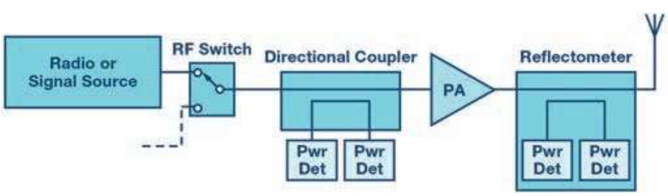

Figure 1: Measuring forward and reflected power in an RF signal chain

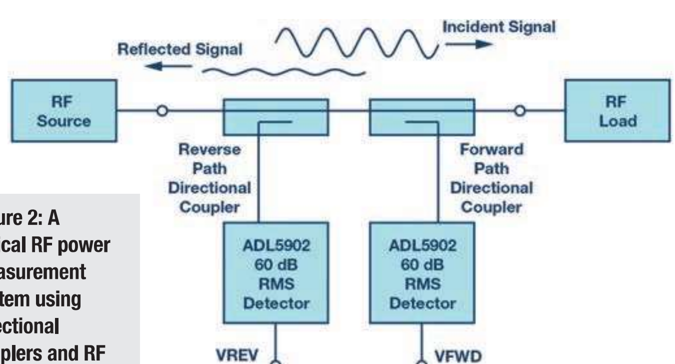

Figure 2: A typical RF power measurement system using directional couplers and RF detectors

Directional couplers have the valuable characteristic of directivity, which is the ability to distinguish between incident and reflected RF power. As the incident RF signal travels through the forward-path coupler on its way to the load (Figure 2), a small portion of the RF power – usually 10-20dB lower than the incident signal – is coupled away to drive an RF detector. Where both forward and reflected power are being measured, a second coupler is used with reverse orientation compared to the forward coupler. The output voltages from the two detectors will be proportional to the forward and reverse RF power levels.

Surface-mount directional couplers suffer from a fundamental tradeoff between bandwidth and size. While bidirectional directional couplers with one octave of frequency coverage (F_{MAX} is equal to twice F_{MIN}) are commonly available in packages as small as 6mm^2, a multi-octave surface-mount directional coupler will be

Figure 3: A connectorised directional coupler, surface-mount directional coupler and ADL5920 integrated IC with directional bridge and dual rms detectors

much larger; see Figure 3. Broadband connectorised directional couplers have multioctave frequency coverage but are significantly larger than surface-mount devices.

Figure 3 also shows the evaluation board for the ADL5920 (Figure 4), a new RF power detection subsystem from Analog Devices, with detection range up to 60dB, packaged in a 5mm × 5mm MLF package (the ADL5920 IC shown is located between the RF connectors).

Instead of sensing the forward and reflected signals using directional couplers, the ADL5920 uses a patented directional bridge technology to achieve broadband and compact on-chip signal coupling. To understand how a directional bridge works, we first need to look at the Wheatstone bridge.

Wheatstone Bridge

The notion of a directional bridge is based on the Wheatstone bridge (Figure 5), which creates zero differential voltage when balanced. In it, one of its legs' resistors is variable (R_2), with the two others (R_1 and R_3) fixed.

There are four resistors in total (R_1, R_2, R_3, R_x), where R_x is an unknown resistance. If $R_1 = R_3$ and R_2 is equal to R_x, then V_{OUT} = 0V. The bridge is considered balanced when the variable resistor is of a value such that the voltage-divide ratios on the left and right side of the bridge are equal and thereby create a zero-volt differential signal across the differential sense nodes that produce V_{OUT}.

A Unidirectional Bridge

Figure 6 shows the schematic of a unidirectional bridge, and it best explains the basic operation of such a device. First, it is important to observe that a directional bridge needs to be designed for a specific Z_0 so that insertion loss is minimised. If $R_S = R_L = R = 50\Omega$, then the sense resistor of the bridge is 5Ω, a good compromise between insertion loss (< 1dB) and signal sensing.

Calculating R_{OUT} as seen looking back from the load results in an exact 50Ω port impedance while calculating R_{IN} will result in 50.8Ω port impedance ($|\Gamma| = 0.008$; RL = -42dB; VSWR = 1.016). If a signal is applied as shown at the RFIP node, since $R_{IN} \sim 50\Omega$, the voltage at RFIP is about half the source voltage. If we assume for a moment that the voltage at RFIP is 1V, then the voltage at RFOP will be about 0.902V. This voltage is further attenuated by 10/11 = 0.909 such that the negative input of the differencing amplifier is 0.82V with a resultant differential voltage of (1 – 0.82) = 0.18V. The effective forward coupling factor (Cpl) of this bridge is:

$$Cpl = 20log_{10}\left(\frac{0.18\ V}{1\ V}\right) = 15\ dB \quad (1)$$

'Balanced' in the context of the bridge means that when a signal is applied in the reverse direction (RFOP to RFIP), then the V_{FWD} detector (or Cpl port) will ideally see zero differential voltage, while

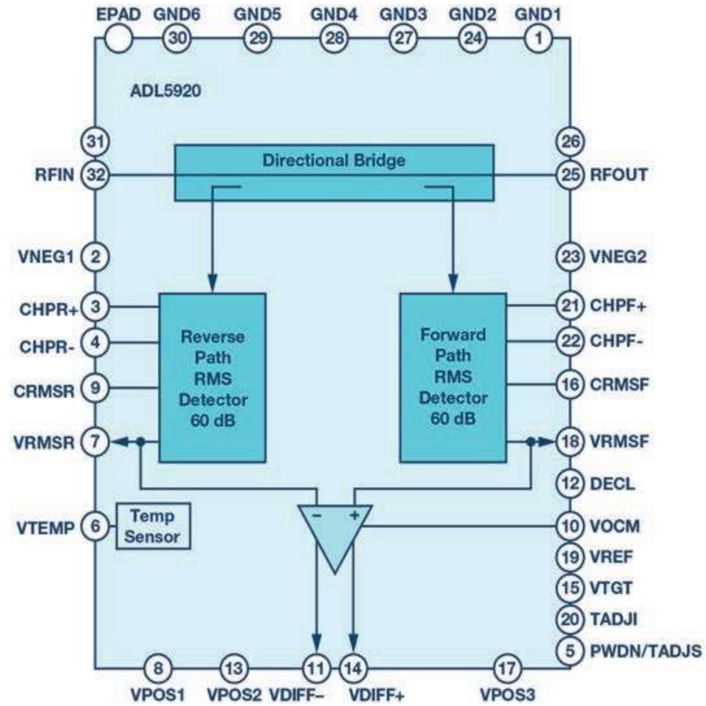

Figure 4: ADL5920 block diagram

it sees a maximum signal when the signal is applied in the forward direction (RFIP to RFOP). To get maximum directivity in such a structure, precision resistors are of utmost importance, which is why integrating them is beneficial.

In a unidirectional bridge, to determine isolation, which is needed to calculate return loss (RL), one needs to flip the device and apply the input signal to RFOP. In this case, the bridge is balanced and the plus and minus inputs to the differential amplifier are equal, since the same divide-ratios of 0.909 = [10R/(10R + R) = (R/(R+0.1R)] result in a differential voltage of (V+ minus V–) = 0V.

Bidirectional Bridge

Figure 7 shows a simplified diagram of a bidirect-ional bridge, like the one used in the ADL5920. The unit resistance R is equal to 50Ω for a 50Ω environment. So, the value of the bridge's sense resistor is 5Ω, while the two shunt networks are about 1.1kΩ each. This is a symmetric network, so the input and output resistances, R_{IN} and R_{OUT}, are the same and close to 50Ω when R_S and R_L are also equal to 50Ω.

When the source and load impedance are both 50Ω, an ohmic analysis of the internal network tells us that V_{FWD} will be quite large compared to V_{REV}. In a real-world application, this corresponds to maximum power transmission from source to load. This results in little reflected power, which in turn results in a very small V_{REV}.

Next, let's consider what happens if R_L is either infinite (open-circuit) or zero (shorted load). In both cases, repeating the ohmic analysis, we find that V_{FWD} and V_{REV} are approximately equal. This

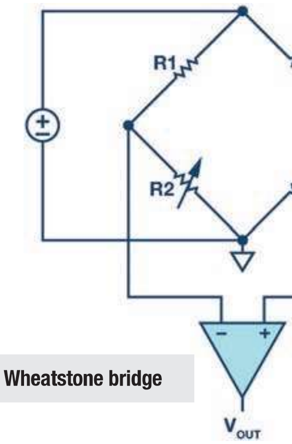

Figure 5: Wheatstone bridge

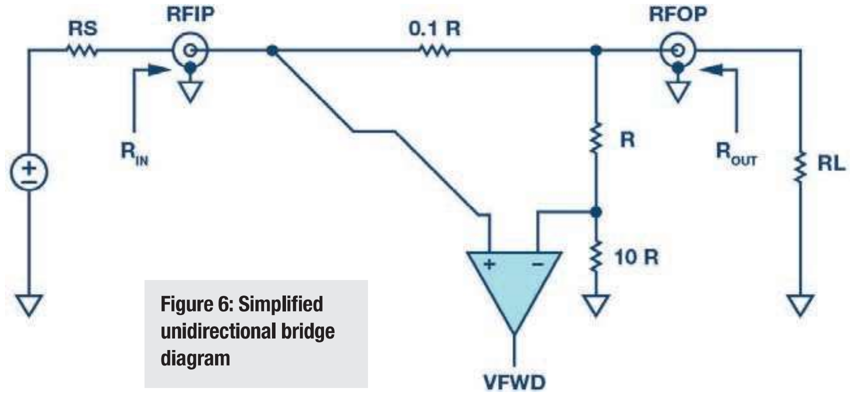

Figure 6: Simplified unidirectional bridge diagram

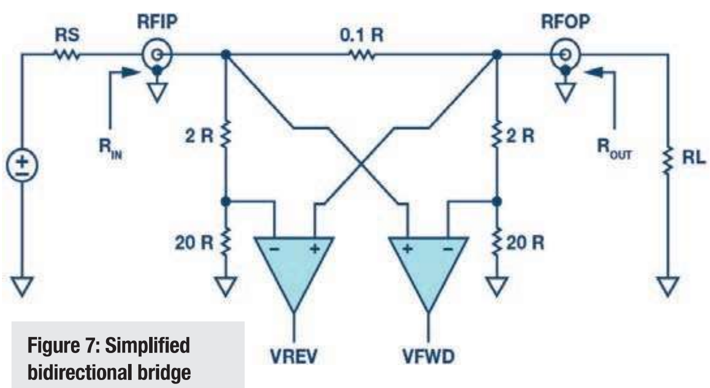

Figure 7: Simplified bidirectional bridge diagram

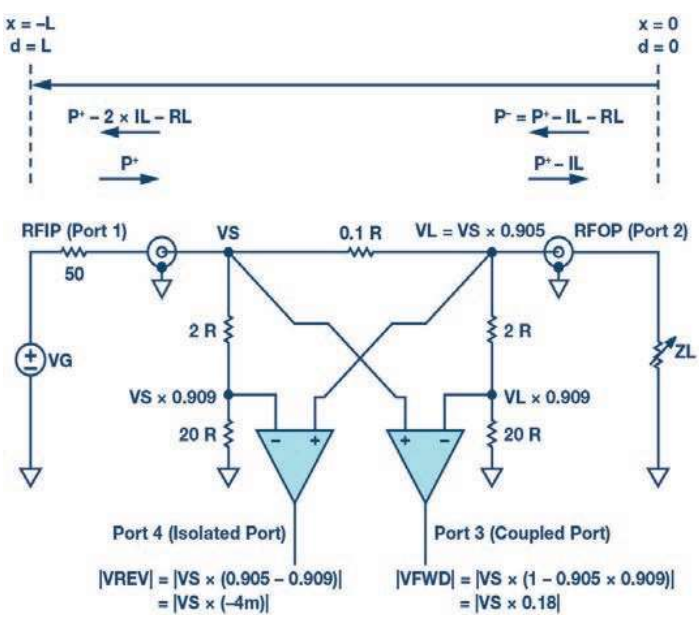

Figure 8: Simplified bidirectional bridge with signals

mirrors a real-world system where an open- or shorted-load results in equal forward and reflected power. A more detailed analysis of these scenarios follows below.

VSWR and Reflection Coefficient

A full analysis of errors in network analysis is complicated and beyond the scope of this article, but we will summarise some of the basic concepts here. An excellent resource is the application note by Marki Microwave, 'Directivity and VSWR Measurements', easily found online.

Travelling waves are an important concept to describe the voltages and currents along transmission lines, since they are functions of position and time. The general solution of voltages and currents along transmission lines consist of a forward-travelling wave and a reverse-travelling wave; both are functions of distance x:

$$V(x) = V^{+}(x) + V^{-}(x) \tag{2}$$

$$I(x) = \frac{V^{+}(x)}{Z_0} - \frac{V^{-}(x)}{Z_0} \tag{3}$$

In Equations 2 and 3, V+(x) represents the voltage wave travelling toward the load, while V–(x) represents the voltage wave reflected from the load due to mismatch. Z_0 is the characteristic impedance of the transmission line, and in a lossless transmission line, it is defined by the classic equation:

$$Z_0 = \sqrt{\frac{L}{C}} \tag{4}$$

The most common Z_0 is 50Ω for transmission lines. If such a line is terminated with its characteristic impedance, it appears to a 50Ω source as an infinite line, since any voltage wave travelling down the line will not result in reflections that can be sensed at the source or anywhere else along the line. However, if the load is other than 50Ω, then a standing wave that can be detected will be generated along the line, defined as the voltage standing wave ratio (VSWR).

More generally, the reflection coefficient is defined as:

$$\Gamma(x) = \Gamma_0 e^{2\gamma x} \tag{5}$$

where Γ_0 is the load reflection coefficient and γ the propagation constant of the transmission line.

$$\Gamma_0 = \frac{Z_L - Z_0}{Z_L + Z_0} \tag{6}$$

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \tag{7}$$

$$\gamma = \sqrt{(R + j\omega L)\,(G + j\omega C)} \qquad (8)$$

R, L, G and C are the resistance, inductance, conductance and capacitance per unit length of the transmission line. The return loss (RL) is the negative of the reflection coefficient (Γ) in dB. This is important to point out since reflection coefficient and return loss are frequently confused and used interchangeably.

$$RL = -20log_{10}|\Gamma_0| = 10log_{10}\frac{1}{|\Gamma_0|^2} \qquad (9)$$

Another very important definition of return loss in addition to the load mismatch above is in terms of incident and reflected power at an impedance discontinuity. This is given by:

$$RL = 10log_{10}\left(\frac{P_{incident}}{P_{reflected}}\right) \qquad (10)$$

and extensively used in antenna design. VSWR, RL and Γ_0 are related as follows:

$$|\Gamma_0| = \frac{VSWR - 1}{VSWR + 1} \qquad (11)$$

$$VSWR = \frac{|V(x)|_{max}}{|V(x)|_{min}} + \frac{1 + |\Gamma_0|}{1 - |\Gamma_0|} = \frac{1 + 10^{\frac{RL}{-20}}}{1 - 10^{\frac{RL}{-20}}} \qquad (12)$$

$$RL = -20log_{10}\left(\frac{VSWR - 1}{VSWR + 1}\right) \qquad (13)$$

Equations 14 and 15 represent the maximum and minimum of the standing-wave voltages. VSWR is defined as the ratio of the maximum to the minimum voltage along the wave. The peak and minimum voltages along the line are:

$$|V(x)|_{max} = |A|(1 - |\Gamma_0|) \qquad (14)$$

$$|V(x)|_{min} = |A|(1 - |\Gamma_0|) \qquad (15)$$

For example, in a 50Ω transmission line, if the forward-travelling voltage signal has a peak amplitude of A = 1 and the line is matched with a perfect load, then $|\Gamma_0| = 0$, there is no standing wave (VSWR = 1.00), and the peak voltage along the line is A = 1. However, if

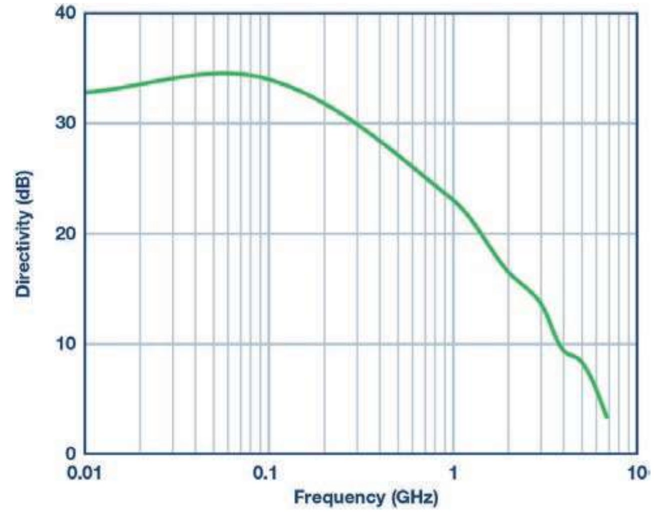

Figure 9: ADL5920 directivity vs frequency; the input level is 20dBm

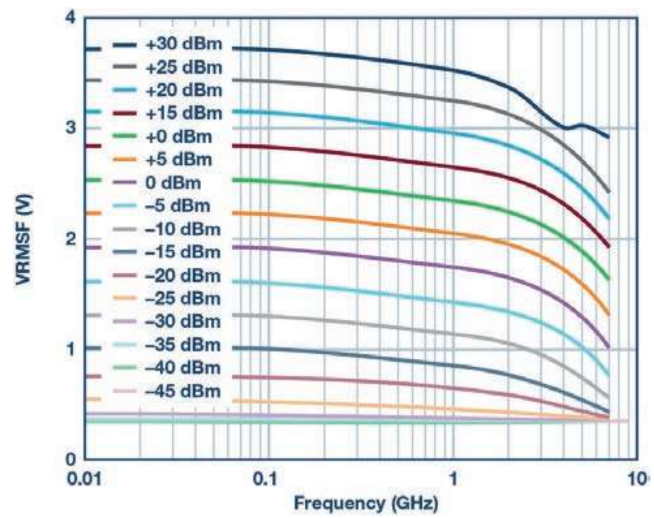

Figure 10: Typical output voltage vs frequency from forward path detector at multiple input power levels

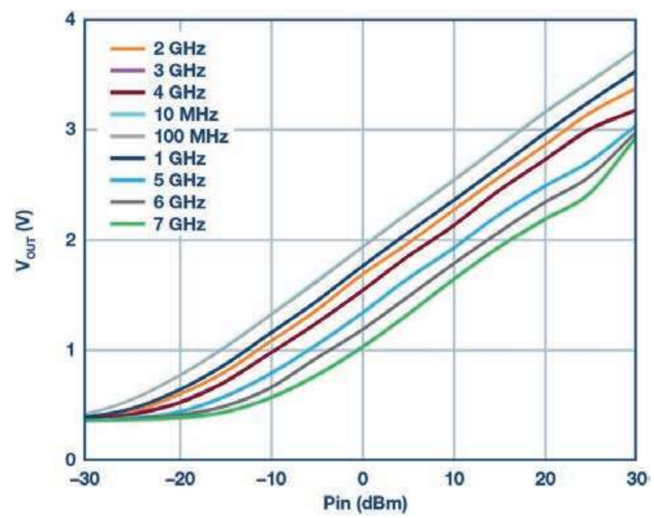

Figure 11: Typical output voltage vs input power from forward path detector at multiple frequencies

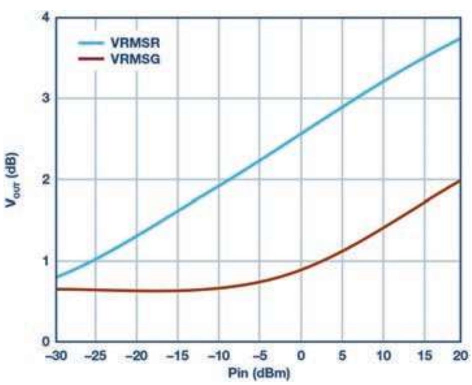

Figure 12: V_{RMSF} and V_{RMSR} output voltage vs input power at 500MHz when the bridge is driven from RF_{IN} and RF_{OUT} is terminated with 50Ω

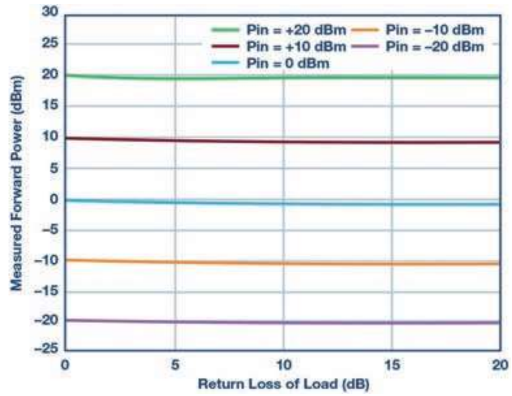

Figure 13: Measured forward power vs applied power and return loss of load, measured at 1GHz

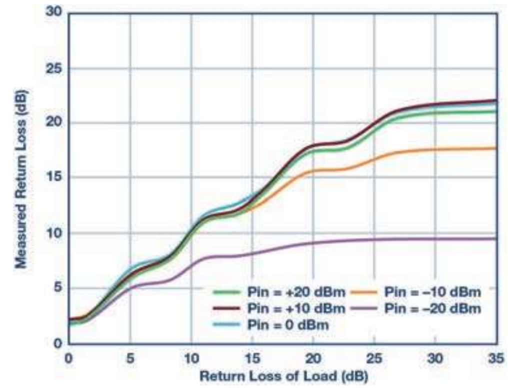

Figure 14: Measured return loss vs applied return loss and RF power, measured at 1GHz

R_{LOAD} is 100Ω or 25Ω, then $|\Gamma_0| = 0.333$, RL = 9.542dB and VSWR = 2.00, with $|V(x)|max = 1.333$ and $|V(x)|min = 0.666$.

Figure 8 is the same as Figure 7 but with signals shown in the default forward configuration and with travelling power waves shown where the reference plane is at the load. At low frequencies, where the wavelength is long relative to the physical structure, voltages and currents are in phase and the circuit can be analysed using Ohm's law.

The ports are defined as shown, with the input (Port 1) at RFIP, output (Port 2) at RFOP, coupled port (Port 3) at V_{FWD}, and isolated port (Port 4) at V_{REV}. Since the structure is symmetric, the ports are reversed when a signal is reflected at Z_L or applied to RFOP.

In the case of a matched load and generator voltage connected to Port 1 (RFIP), and with $Z_S = Z_L = Z_0 = R = 50Ω$:

$$V_L = V_{S+}\left[\frac{Z_{OUT}}{Z_{OUT}+0.1R}\right] = V_{S+} \times 0.905 = V_{S+} \times |S21| \quad (16)$$

$$Z_{OUT} = Z_L \mid\mid (2R + 20R) = R \mid\mid 22R = \left(\frac{22}{23}\right)R \quad (17)$$

and V_L/V_{S+} is the insertion loss, L_1, or IL in dB.

$$IL = -20log_{10}|S21| = -20log_{10}L_1 = 0.87\ dB \quad (18)$$

The attenuation factor α for the two shunt legs on either side of the main line resistor of 0.1 × R is:

$$\alpha = \frac{20R}{(20R + 2R)} = \frac{20}{22} = 0.909 \quad (19)$$

The equations in Figure 8 for $|V_{REV}|$ and $|V_{FWD}|$ show the values for those voltages with a signal applied in the forward direction. These equations indicate a fundamental directivity limit for the simplified schematic due to non-ideal rejection of 33dB at the isolated port.

$$D = 20log_{10}\left(\frac{|V_{CPL}|}{|V_{ISO}|}\right) = = 20log_{10}\left(\frac{|0.18|}{|-0.004|}\right) = 33\ dB \quad (20)$$

Figure 8 shows that the directivity of the bidirect-ional bridge in the linear domain is determined by:

$$D_L = \left(\frac{1 - L_1 \times \alpha}{L_1 - \alpha}\right) \quad (21)$$

which means that to increase directivity, α needs to equal the insertion loss, L_1. In silicon, the peak directivity is typically better than the simplified diagram would indicate (Figure 9).

If Z_L is not equal to Z_0, as is normally the case, the coupled and isolated port voltages, which are complex, would be:

$$V_{CPL} = V_{S+}[1 - L_1 \times \alpha] + V_{L-}[L_1 - \alpha] \quad (22)$$

$$V_{ISO} = V_{L-}[1 - L_1 \times \alpha] + V_{S+}[L_1 - \alpha] \quad (23)$$

where V_{S+} is the forward voltage at Port 1 (node V_S) and V_{L-} is the reflected voltage from the load at Port 2 (node V_L). Θ is the unknown phase of the reflected signal:

$$V_{L-} = V_{S+} \times L_1 \times |\Gamma_0|e^{j\Theta} \quad (24)$$

Substituting Equation 24 for V_{L-} in Equations 22 and 23 and using Equation 21 to simplify the result, plus the fact that:

$$V_{FWD} = V_{S+}[1 - L_1 \times \alpha] \tag{25}$$

results in complex output voltages:

$$V_{CPL} = V_{FWD}\left\{1 + \frac{L_1 \times |\Gamma_0|e^{j\Theta}}{D_L}\right\} \tag{26}$$

$$V_{ISO} = V_{FWD}\left\{L_1 \times |\Gamma_0|e^{j\Theta} + \frac{1}{D_L}\right\} \tag{27}$$

From Equations 26 and 27 we can observe that for $D_L >> 1$:

$$\left|\frac{V_{ISO}}{V_{CPL}}\right|_{max,\ min} = \sqrt{\frac{\left(\frac{1}{D_L}\right)^2 \pm 2\left(\frac{L_1 \times |\Gamma_0|}{D_L}\right) + (L_1 \times |\Gamma_0|)^2}{1 + 2\left(\frac{L_1 \times |\Gamma_0|}{D_L}\right) + \left(\frac{L_1 \times |\Gamma_0|}{D_L}\right)^2}} \rightarrow L_1 \times |\Gamma_0| \tag{28}$$

In the ADL5920, the voltages V_{REV} and V_{FWD} are mapped via two 60dB-range linear-in-dB rms detectors into voltages V_{RMSR} and V_{RMSF} that are (V_{ISO}/V_{SLP}) and (V_{CPL}/V_{SLP}) in dB, respectively. So, the differential output of the device V_{DIFF} in dB represents:

$$\frac{V_{DIFF}}{V_{SLP}} = \frac{VRMSR - VRMSF}{V_{SLP}} = \frac{V_{L_1} + V_{|\Gamma_0|}}{V_{SLP}} \tag{29}$$

where V_{SLP}, the detector slope, is about 60mV/dB.

Using the voltage-to-dB mapping of Equation 29 in Equation 28:

$$20log_{10}\left(\frac{VRMSR}{V_{SLP}}\right) - 20log_{10}\left(\frac{VRMSF}{V_{SLP}}\right) = 20log_{10}(L_1) + 20log_{10}|\Gamma_0| \tag{30}$$

And using Equation 9 in Equation 30 gives:

$$P_{REV} - P_{FWD} = -IL - RL \tag{31}$$

$$RL = P_{FWD} - P_{REV} - IL \tag{32}$$

Figure 10 shows the response of the forward power-sensing rms detector when the ADL5920 is driven in the forward direction. Each trace corresponds to the output voltage vs frequency for a specific power level. While the plot stops at 10MHz, operation at frequencies to 9kHz has been verified. In Figure 11, the same data is presented as output voltage vs input power, with each trace representing a different frequency.

When the ADL5920's RF_{OUT} pin is terminated with a 50Ω resistor, there should be no reflected signal. Therefore, the reverse path detector should not register any detected reverse power. However, because the directivity of the circuit is non-ideal and rolls off with frequency, some signal will be detected in the reverse path.

Figure 12 shows the voltage measured on the forward and reverse path detectors at 500MHz with RF_{IN} swept and RF_{OUT} terminated with 50Ω. The vertical separation between these traces relates directly to the directivity of the bridge.

Figure 13 shows the effect of varying the load on the measurement of forward power. Defined power levels are applied to the RF_{IN} input, and the return loss of the load on RF_{OUT} is varied from 0-20dB. As expected, with the return loss in that range, power measurement accuracy is quite good. But, as the return loss is reduced to below 10dB, the power measurement error starts to increase. It is notable that for a return loss of 0dB, the error is still around 1dB.

In Figure 14, the ADL5920 is used to measure the return loss of the load, also at 1GHz. A known return loss is applied to the RF_{OUT} port. V_{RMSF} and V_{RMSR} are measured, and the return loss is back-calculated.

There are several points to note about this plot. First, it shows that the ADL5920's ability to measure return loss degrades as return loss improves, which is due to the directivity of the device. Second, note how measurement accuracy degrades as the drive power drops, which is due to the limited detection range and sensitivity of the ADL5920's on-board rms detectors. The third observation relates to the apparent ripple in the traces. This is caused by the fact that each measurement is being taken at a single return-loss phase. If the measurement was repeated at all return loss phases, a family of curves would result, with vertical width roughly equal to the vertical width of the ripple.

Applications

With the ability to measure inline RF power and return loss, the ADL5920 is useful in multiple applications. Its small size means it can be dropped into many circuits without a significant space impact. Typical applications include in-circuit power monitoring at RF power levels up to 30dBm, where insertion loss is not critical.

The return loss measurement capability is typically used in applications where an RF load is being monitored. This could be a simple circuit to check that an antenna has not been damaged or broken off (that is, catastrophic failure). However, ADL5920 can also be used to measure scalar return loss in materials analysis applications. This is most applicable at frequencies below 2.5GHz where directivity (and thereby measurement accuracy) is greater than 15dB. EW