A 2 MHz 3-Port Analog Isolation and Fanout Module*

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Abstract

A 3-port isolated circuit providing 1:3 fanout, buffering and amplification over a multi-megahertz bandwidth is presented. The circuit accepts a single input and drives 3 independently isolated output channels, up to +/- 10 V into 50 ohms. The input and output isolation is supplied via a dual optocoupler, and the power isolation is achieved with DC/DC converters. In each channel, a voltage feedback amplifier is used in combination with the optocoupler to form a transimpedance configuration with the gain-bandwidth product (GBP) set by a pair of resistors. The feedback amplifier linearizes the optocoupler transfer characteristics using a servo technique and also controls the circuit drift, nonlinearity, and bandwidth. The circuit has demonstrated long-term drift of <+/- .1% of full scale, and resolution to better than 9 bits. The circuit provides frequency response to true DC with an analog bandwidth variable over a range of <100 kHz to >4 MHz, and a SNR of >55 db in a 1 MHz bandwidth with < 1% THD for a 10 V amplitude sinusoidal input. With few modifications, this design is capable of providing input/output gain and bandwidth in the range of 10 - 50 MHz.

I. INTRODUCTION

In many accelerator based data acquisition systems, signal isolation is a necessary feature so ground loops are avoided. This problem commonly occurs in distributed data acquisition and also in signal distribution from a single point to multiple destinations. Many techniques exist for isolation of analog signals and they are summarized in various places.[1] A common approach is to use voltage-to-frequency converters (VFCs). However, for isolated data transmission of analog signals, because of the digital output, the information bandwidth of VFCs is generally limited to well below their carriers. Further, situations occur where the data must remain in analog form, and although analog fiber optic technology[2,3,4] is an alternative for new installations, in some cases it is not possible to install fibers, and the existing copper lines must be used. Thus, single package analog isolation amps must be considered. There exists a wide variety of components available from several manufactures, such as Burr-Brown and Analog Devices, but almost all commercially available units have bandwidth below the 100 - 200 kHz range, and many do not isolate the DC power. It is valuable to achieve isolation from the DC power sources so that the system is completely independent of local ground references, and thus the processing electronics can be located anywhere user demand requires. Addressing the need for a universally applicable 3-port analog isolation fanout/buffer with a DC - 2 MHz bandwidth for signal distribution, the following system was developed.

II. CONCEPT

A picture of the module is shown in figure 1. The form factor is a 10 HP x 3 U x 220 mm eurocard package. This format

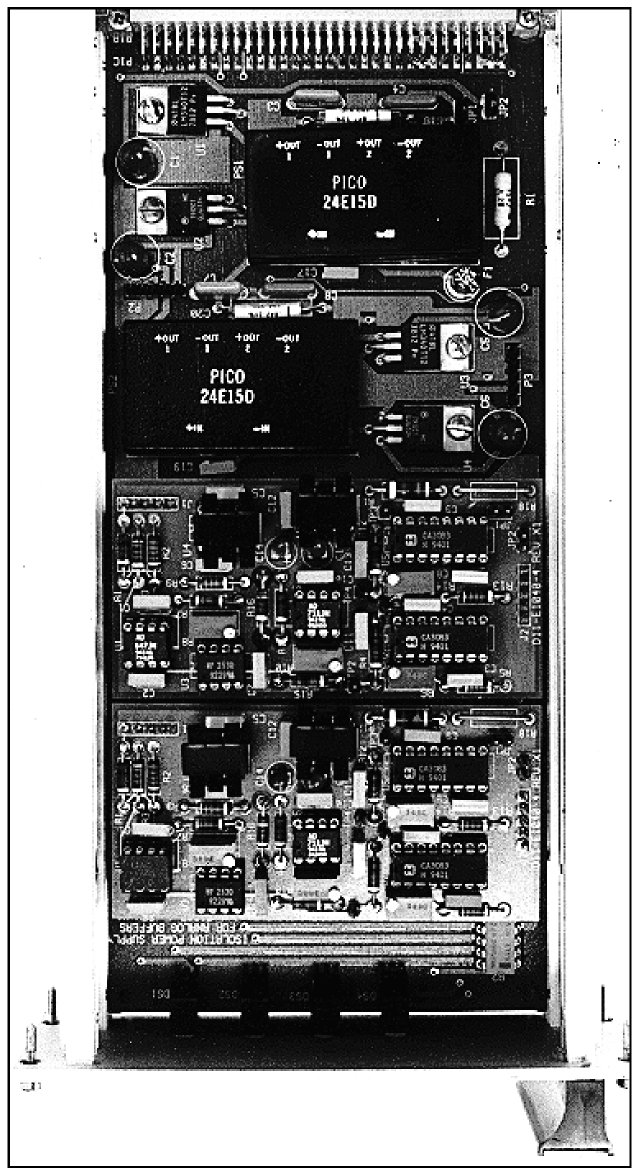

Figure 1. Module populated with 2 of 3 daughter cards

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was selected by considering criteria relating to the application of the circuitry and is not essential to the design. The signal isolation, buffering, and amplification are performed on each daughter card using +/-12 rails. The rail voltages are generated on the host card using 4 DC/DC converters (PICO 24E15D, $80), which use a common +24 VDC supply. The converters supply an unregulated +/- 15 VDC with > 200 mA per output, and switch at nominally 20 - 40 kHz. The converters have internal filtering for noise reduction on the analog supply lines, and were selected based on current output, size and cost. The unregulated converter outputs caused problems when output stages drive large signals into low impedances, so fixed 12V regulators were added to each channel. In addition to regulation, they further reduced the power line ripple and noise seen by the circuitry. One converter powers the input circuit on each of the three daughter cards, and the remaining three each drive the output circuitry of a single daughter card. A common input circuit is replicated on each daughter card to ease the packaging design and simplify the input to output isolation circuitry design. The returns for the input and output circuit power are supplied separately by the signal source and loads. Thus, the DC power for each channel is completely isolated from any local rack or crate grounds. As a result, all the circuit ports, the 3 outputs, input, and local power are pairwise isolated, and true 3-port isolation has been achieved in each channel.

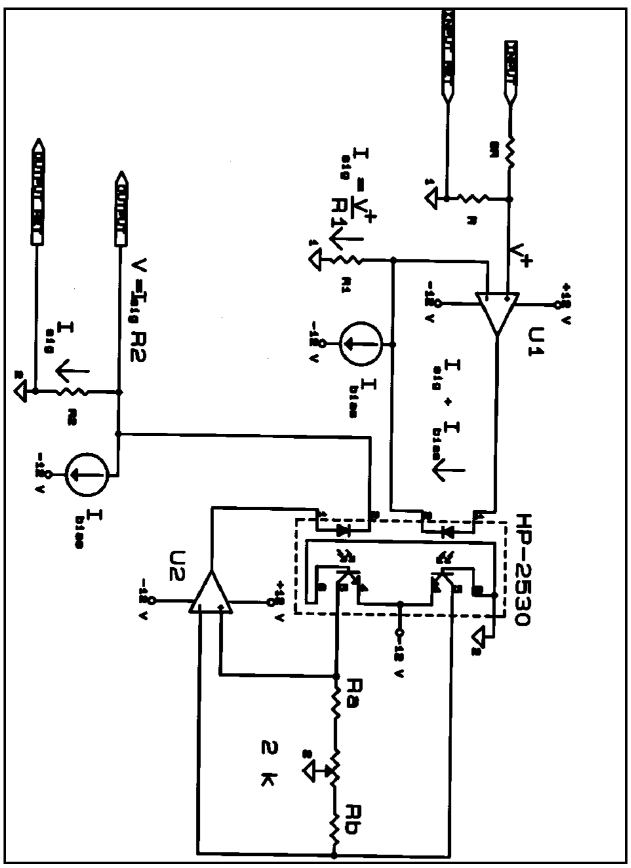

Figure 2. Simplified daughter card schematic

The conceptual design (figure 2) of the daughter card was adapted from a Hewlett-Packard application note[5] and consists of an input attenuator, followed by an opamp stage that linearly converts the input voltage to a current, and optically couples it across the isolation barrier. The input signal is recovered by sensing the difference in the D1 and D2 forward currents as a differential voltage using the resistors Ra and Rb and opamp U2. The feedback action of U2 forces the opamp output to move until its differential input is maintained at virtual ground. At this point the diode forward currents match, and assuming matched coupler halves, the collector currents are matched. The signal portion of D2's forward current flows through R2 to generate the output voltage Vout=Vin(R2/R1). Thus by sizing the ratio of R1 and R2 "noise-free" voltage gain can be achieved. This imbalance does not alter the cancellation properties of the circuit. In the design presented here, gains of up to 10 have been achieved in this way, and higher gains are feasible. The current sources bias the optocouplers into a "linear region", and the exact value of the current selected depends on the input and output swings desired as well as the minimum SNR and maximum distortion tolerable.

III. DETAILED DESIGN

The circuit accepts a 20 Vp-p input and produces a 2 Vp-p swing across the resistor R1. Using a 470 ohm resistor for R1, this sets the diode signal current swing at about 4 mAp-p. Empirically this value was found to produce low distortion (< 1%) and good dynamic range (> 55 db in a 1 MHz bandwidth). For applications requiring less than 20 Vp-p input swings, the attenuator and/or R1 should be set so that the current swing through R1 is in the 2-4 mAp-p range. Larger full-scale swings increase the distortion when the coupler halves have mismatches, and smaller swings reduce the usable dynamic range by limiting the modulation of the optical carrier. The total current flowing in diode D1 is Isig+Ibias. The Isig component is provided solely by the input signal. Thus, the input amplifier U1 (AD711) acts as a voltage-to-current converter for the input signal and should be able to drive > 20 mA into low impedance loads for good performance. The nonlinear resistance of the LED in the feedback path of U1 does not effect the circuit linearity since, to a good approximation, the current through the diode is linearly related to the non-inverting voltage.

LED bias currents > 5 mA are required to provide good performance, otherwise the SNR is sacrificed for lower distortion. Experimentally, bias currents in the 7 - 10 mA range have been adequate to provide the best trade-offs. The bias currents are generated by Wilson current sources implemented with an RCA CA3083 transistor pack.

Wilson sources were used to get much higher output impedances than achievable with simple two transistor sources. Two transistor sources have output impedances limited to 130/Iout by the Early Effect, where Wilson sources provide approximately β times more impedance and less sensitivity to β variations[6]. The increased impedance provided by the Wilson sources enabled better bandwidth characteristics when R1 > 100 ohms. Each source is independently adjustable to aid circuit matching. Under matched conditions, the nonlinear transfer characteristics of the optocouplers are nulled out by the feedback action of U2, however mismatches between components (i.e.

optocoupler current transfer gain) limit the lower bound on the distortion reduction achievable. However, even this may be compensated for to some degree by selecting the proper ratio of Ra and Rb. In most cases some experimentation is required in selecting the circuit parameters to achieve an optimum balance between maximum distortion, bandwidth, and SNR. For very wideband applications of this circuit, such as extensions to the 10 MHz region, the pole created by R1 and the capacitances of the opamp and current source begin to limit the useful range of the device and compensation measures may be necessary. Also, different opamps will be necessary to handle the larger bandwidth.

A voltage feedback opamp U2 (PMI OP42), senses the current from one side of the coupler, and develops an output voltage to maintain the virtual ground at its input. The virtual ground is achieved when the U2 output is sufficient to drive D2 such that current coupled through to Rb nulls the differential input of U2. Nominally Ra and Rb are equal so that with the virtual ground condition, the forward currents through each LED are the same (assuming matched optocouplers and bias currents), and thus the current flowing through R2 is identical to the input circuit's Isig current. Because Isig is linearly dependent on the input voltage, the input and output signals are linearly related by Vout=Vin(R2/R1). However, the current through R2 is independent of the resistance in that branch, and hence voltage gain for Vin can be achieved by increasing the ratio of R2 to R1 beyond unity. This is a very low noise way to achieve voltage gain without adding extra components. The mismatch between R1 and R2 does not effect the linearity or bandwidth. Further, in some applications an intentional imbalance in Ra and Rb can be introduced to compensate for gain mismatches in the optocouplers, or to add an additional gain control parameters to the circuit.

The optocoupler and U2 opamp combination form a transimpedance amplifier. The feedback amplifier so formed controls the nonlinearity introduced by the optocoupler transfer curves, drift in the operating points, and overall system bandwidth. The current mixing action of the transimpedance amplifier is performed by using the collector resistors Ra and Rb to convert the optocoupler output currents to voltages, and then using U2 to sense the differential signal across Ra and Rb. The collector resistance sets the gain-bandwidth product (GBP) of the transimpedance amplifier at roughly αRAf with a DC open loop gain of $RA\alpha$. The expressions were derived assuming a single-pole opamp model, using α as the current transfer ratio of the coupler, A as the DC open loop gain of the opamp, R as the value of the collector resistor (either Ra or Rb), and f as the unity gain crossover frequency of U2. The expression for the GBP is a loose approximation, but shows the role of the collector resistance. In fact, the increase in the GBP by αR can cause oscillations to occur if U2 is not selected with sufficient phase margin to maintain stability for the desired value of R. Thus careful consideration of the opamp and possible R values is warranted.

The output amplifier (not shown in the schematic) sensing the signal across R2, is a fixed gain current boosted amplifier. The amplifier used is the Analog Devices AD711 with an Elantec EL2003 current buffer enclosed in the feedback loop. The output gain is designed at 1.6 which limited the output stage bandwidth to the desired 2 MHz. If wider a bandwidth is required, opamps like the AD847 can be used to replace the AD711's and OP42. The AD847 has been tested in the circuit and maintains stability. The current driver, EL2003, is used because it provides short circuit proof operation and will typically drive >200 mA into 50 ohms. In addition, locations for back termination resistors have been included for applications requiring them. The circuit has been tested driving cables in excess of 1000 feet.

IV. RESULTS

The circuit was tested for offset and drift. The drift test was performed over several days under loaded output conditions with the input shorted. The drift was measured at < +/- 10 mV under laboratory conditions using a strip chart recorder with an ambient temperature range of 15 - 27 degrees Centigrade. Preliminary testing indicates that this drift is note strongly dependent on the ambient temperature, but rather local heating effects in the opto-isolator. The drift was measured after the device experienced a 10 - 15 minute warm-up time. During this time, the output offset drifts approximately 50 mV. Readjustment after this phase using the 2K pot (figure 2) nulled the output.

Frequency domain tests of the circuit 's response were made with a network analyzer. The tests showed that the circuit configured for overall unity gain with AD711's achieves a cutoff frequency of approximately 2 MHz for a 1 Vp-p input. Using AD847's a cutoff of approximately 4 MHz is achieved. The noise levels in these implementations were measured using an oscilloscope assuming a peak crest factor of 4, and the results of 5 mVrms (AD711) and 10 mVrms (AD847) were observed. If the optocoupler is changed to one of the faster single units, a bandwidth over 30 MHz can be achieved. We demonstrated this in the lab using the two single HP4562 optocouplers. In this case however circuit balancing is more difficult and drift increases.

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