# 16 Bit Digital Multimeter Module (DMM)

This document briefly describes the 16-bit DMM that was designed to supersede the 12-bit one. Full functional compatibility was retained while simultaneously offering improved performance. This was particularly useful in specific applications where existing performance was inadequate or unstable.

Development of the 16-bit DMM was accomplished through a combination of design improvements and use of new generation integrated circuits to replace obsolete devices.

## Design Goals

Specific goals of this new design were to:

- Use a higher resolution A/D converter to enhance measurement accuracy both at low levels and, coupled with digital averaging techniques, in the presence of noise at all levels.
- Eliminate amplifier feedback instability that affected measurement of emitter current for high beta bipolar transistors. Data from this measurement is used to generate Gummel plots.
- Use reed relay switches with improved thermal emf performance to enhance current measurement accuracy of non-ideal sources encountered in such applications as the Gummel one plus realistically achieve improved low level voltage measurement accuracy.
- Reduce the number of parts in the amplifier output stage and improve heat dissipation of same for longer MTBF at high output currents.
- Improve Common Mode Rejection Ratio (CMRR) to enhance measurement accuracy when the meter low terminal is connected to potentials other than ground.

## Capacitance Based A/D Converter

The DMM-16 uses a monolithic CMOS A/D converter with an inherent sampling architecture based on the successive-approximation algorithm. Instead of the traditional resistor network, the internal DAC employs a binary weighted capacitor array with all capacitors sharing a common node as shown in figure 1. Because the bipolar mode is being used, during conversion the MSB capacitor is connected to the reference voltage, $V_R$, while all other capacitors are connected to the input voltage, $V_{IN}$.

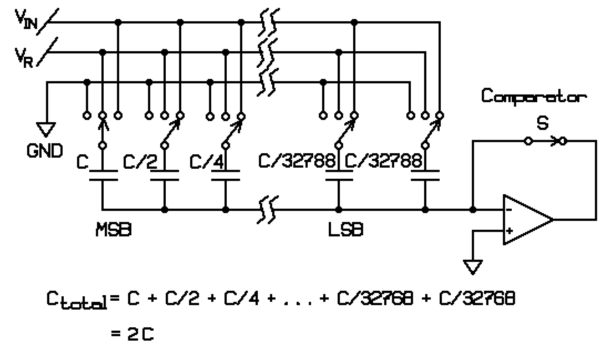


*Figure 1 - Successive Approximation A/D Converter*

While switch S is closed, charge on the input capacitor array is set by $V_{IN}$ while that on the MSB capacitor is set by $V_R$. Thus, total charge can be calculated to be:

$$Q_{total} = CV_R + CV_{IN} \tag{1}$$

When the conversion command is given, switch S opens and the capacitor array elements are disconnected from $V_{IN}$, thus trapping the fixed charge, $CV_{IN}$, on them. A/D conversion consists of sequentially connecting the ends of the capacitors to either $V_R$ or to ground under control of the successive-approximation algorithm. At the end of conversion, the array (including the MSB capacitor) is a capacitive divider between $V_R$ and ground. The divider output is at 0V because that is the target of the A/D algorithm.

Since there is 0V across those array capacitors tied to ground, fixed charge is only a function of $V_R$. If $C_1$ is the capacitance eventually tied to $V_R$ (inclusive of the MSB capacitor), total fixed charge is:

$$Q_{total} = C_1 V_R \tag{2}$$

Because equations 1 and 2 define the same charge:

$$C_1 V_R = CV_R + CV_{IN} \tag{3}$$

Collecting terms and simplifying results in a simple relationship for determining input voltage from $C_1/C$:

$$V_{IN} = \left( \frac{C_1}{C} - 1 \right) V_R \tag{4}$$

## Assuring 16-bit Linearity

Linearity of the A/D converter is the single most important measure of its quality and is typically described by a specification of Differential Non Linearity (DNL). Because each capacitor in the array actually consists of several capacitors, which are switched in or out by an on-chip controller during a calibration cycle, their respective binary weights are set with a resolution of 18 bits. In this way the DNL errors are reduced to less than 1 LSB at any point.

## Stability with Bipolar Transistors

Reedholm instrumentation uses a current measurement technique that is favored industry wide because of its nearly ideal zero input voltage compliance and its comparatively fast response time. It is sometimes known as a feedback ammeter for reasons that may be appreciated by viewing the block diagram below:

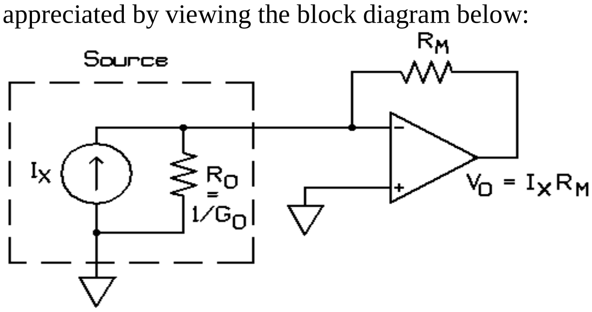


*Figure 2 - Feedback Ammeter with Near Ideal Source*

The high gain amplifier holds the input voltage at virtual ground and converts the input current to a voltage that is scaled by the metering resistor RM. This technique works very well indeed, so long as the current source is near ideal—that is, a source with near zero output conductance $G_o$, or equivalently, a virtually infinite output resistance $R_o$.

However, when the feedback ammeter is used to measure emitter current of a bipolar transistor connected through a system switching matrix, a number of non-ideal parasitic elements enter the picture as shown in Figure 3. Values of $C_1$ and $C_2$ are typical Reedholm system node capacitances.

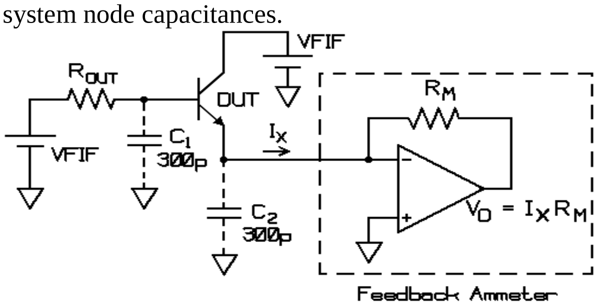


*Figure 3- Bipolar Transistor as Source*

The output conductance of this "current source" is certainly not zero! There are some system imposed parasitic capacitances, in addition to those of the DUT that unavoidably become part of a feedback amplifier system. Using a simple model for the DUT with current gain HFE, the above diagram may be redrawn as:

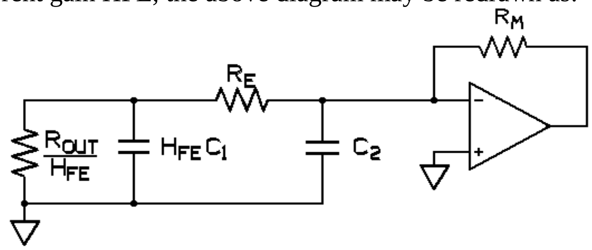


*Figure 4- Bipolar Transistor Equivalent Circuit*

From here, it can be seen that the "current source" output conductance may be considerably capacitive in nature, particularly for large $H_{FE}$ values. This relatively large capacitance causes an increase in phase shift to the point the feedback system can oscillate. The DMM-16 design has made significant improvement for performance in this application. It will maintain stability for $H_{FE} \geq 1000$.

The other effect of non-zero current source output conductance has to do with changing the step response time. Consider again the simple block diagram of the feedback ammeter shown in Figure 2. To the extent that the current source resistance $R_o$ is not infinite, the loop gain of the amplifier system is reduced by the ratio $R_o/R_M$.

This means that the same ratio both reduces the system bandwidth and increases the step response time. One of the most prevalent applications for which this is a consideration is the measurement of the emitter current of a bipolar transistor. Here the step response time may be increased by as much as 200:1. For reference, voltage step response times, which are not dependent on source impedance, are shown in Figure 5.

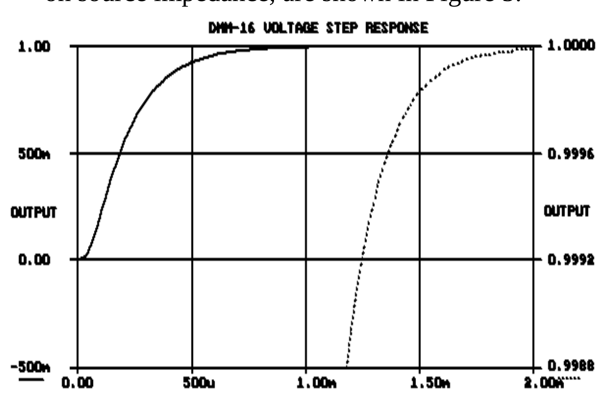


*Figure 5 – Voltage Step Response*

Current step response times and their variations with current source output impedance are shown in Figures 6, 7, and 8. Notice that the right axes show response times to small uncertainties.

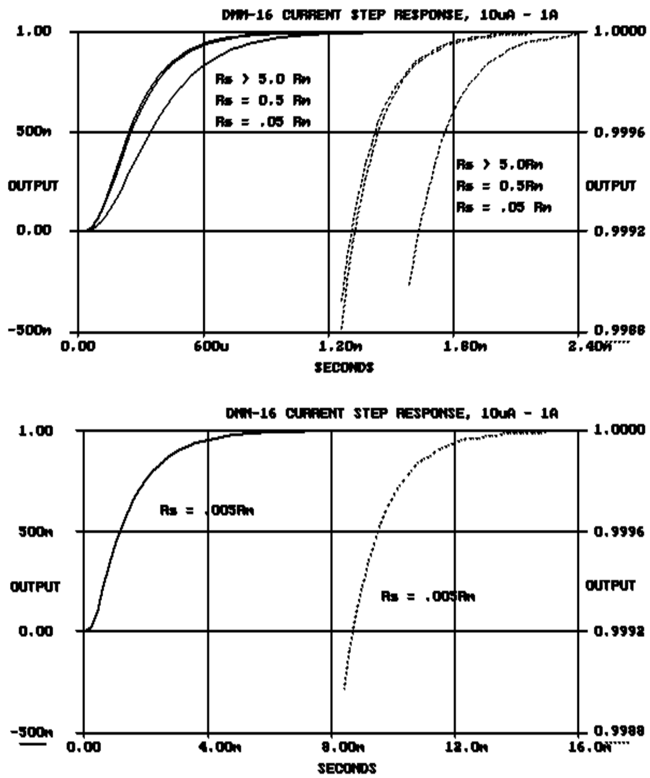


*Figure 6 - Current Step Response, 10µA - 1A Ranges*

## Thermal Voltages and Accuracy

Referring again to the block diagram in Figure 2, anything that would allow the input voltage to be non-zero would create measurement error for non-ideal current sources. For this reason, relays are specified at a maximum thermal voltage of 5µV. This ensures that the maximum current measurement error for the relatively severe requirements of the Gummel application is less than 0.02% due to thermal voltages.

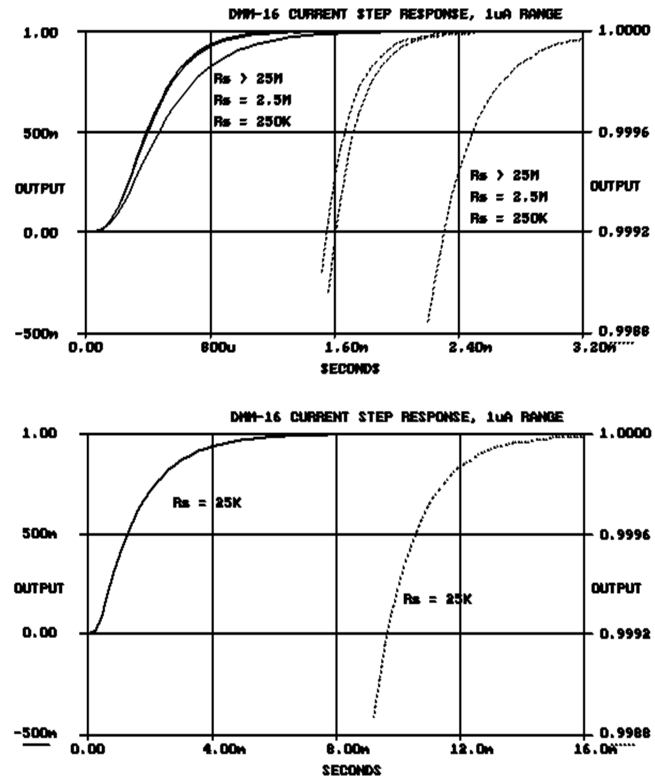


*Figure 7 - Current Step Response, 1µA Range*

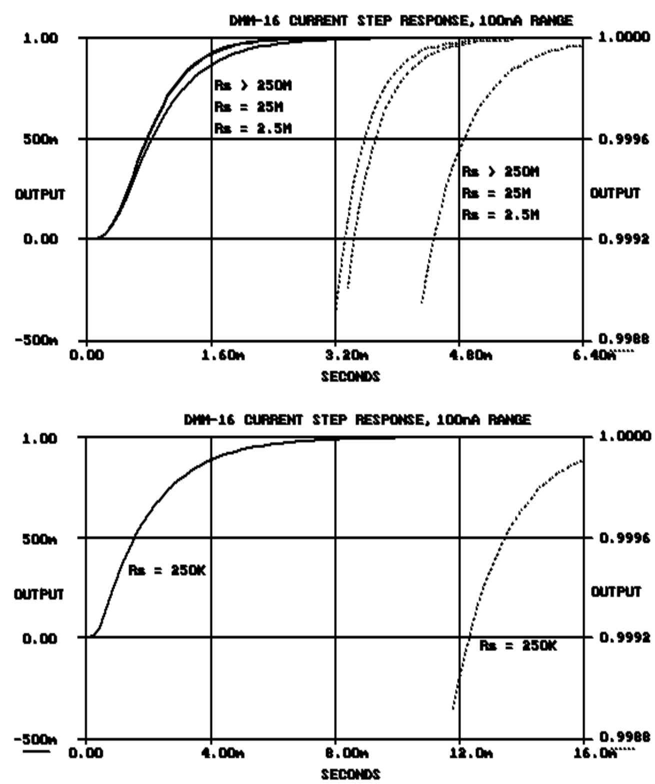


*Figure 8 - Current Step Response, 100nA Range*

## Power Output Stage

A partial block diagram of the older DMM-12 200mA output stage is shown in figure 9 for reference:

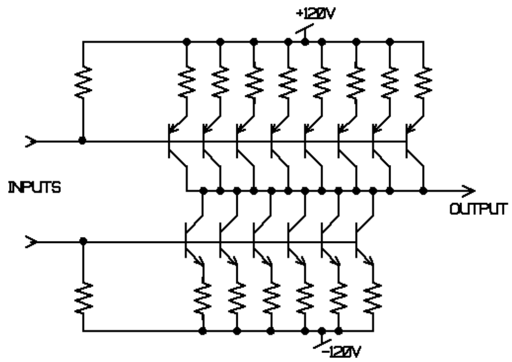


*Figure 9 – DMM-12 Output Stage*

The large number of transistors was necessary because of secondary breakdown in bipolar transistors. By spreading current load among several transistors, 200mA+ could be sourced in either polarity.

With the DMM-16, bipolar transistors were not needed for the output stage and the MOSFET's that were used do not have a thermal run-away condition. Figure 10 has a partial block diagram for comparison.

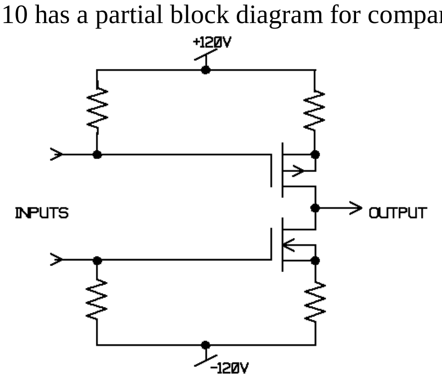


*Figure 10 - DMM-16 Output Stage*

A single transistor conducts 200mA+ in each polarity because secondary breakdown does not exist. Additionally, maximum junction temperature of output devices was reduced by greater than 20C° with larger and more conductive heat sinks.

## Common Mode Effects

Though most voltage measurements can be made with respect to system ground, in some cases it is more convenient to make floating measurements. The circuit shown below converts an unknown floating voltage to a ground referenced voltage in both the old and new Reedholm DMM's:

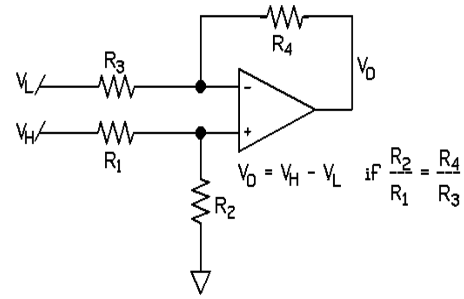


*Figure 11 – DMM-12 Difference Amplifier*

The output voltage $V_o$ equals the difference in voltage between the two inputs, so it called a "difference amplifier." Because of tolerances in the precision resistors $R_1$ through $R_4$, the conversion is not perfect. Common mode voltage-caused error due to resistor value error can be shown to be:

$$\frac{\partial V_o}{V_{cm}} = 0.5 * \left[ \frac{\partial R_2}{R_2} + \frac{\partial R_3}{R_3} + \frac{\partial R_1}{R_1} + \frac{\partial R_4}{R_4} \right] \qquad (5)$$

If the precision resistors had individual tolerances of ±0.01% and temperature coefficients of ±5ppm/°C, worst case error in output voltage can be calculated to be ±0.02% of the common mode voltage, $V_{cm}$, with a temperature variation of ±10ppm/°C of $V_{cm}$. At the maximum $V_{cm}$ = ±100V, worst-case error voltage is ±(20mV+1mV/C°).

The DMM-16 was improved over the DMM-12 by using TC matched resistor pairs to reduce temperature variation and by using a CMRR adjustment to compensate for resistor tolerance errors as shown in Figure 12.

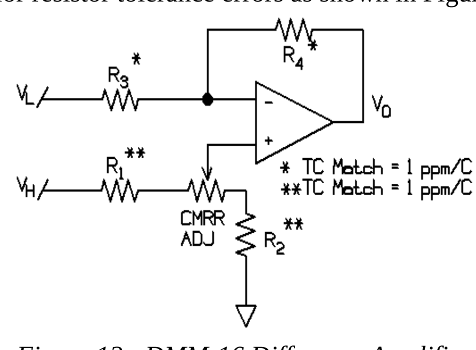


*Figure 12 - DMM-16 Difference Amplifier*

During calibration, the potentiometer is adjusted for zero output voltage with 100V simultaneously applied to both inputs. This virtually eliminates the static output error. Since the two resistor sets are TC matched, worst-case temperature variation is ±1ppm/C° of $V_{cm}$, or ±100µV/C° for $V_{cm}$ = ±100V.

## Value of New Design

The DMM-16 is fully compatible with the installed base of Reedholm test systems running supported software. True 16-bit conversion of the measured parameter is provided. Because of excellent differential linearity, measurements are made with sixteen times the resolution of the DMM-12.

Use of relays with lower thermal emf's provides commensurately better µvolt level measurements. In addition, lower thermals provide higher accuracy current measurements on non-ideal current sources. Furthermore, redesign of the feedback control loop enables emitter current measurements on bipolar transistors with betas in excess of 1000.

CMRR has been improved to reduce voltage measurement errors for floating applications. The DMM-16 is even more reliable than the DMM-12 by virtue of 20% parts count reduction and a 20°C reduction in the maximum junction temperature of the transistors in the power output stage.

Although not explicitly mentioned in the text, the DMM-16 can be calibrated without having to search for potentiometers on the board. All of them are along the board perimeter.

## DMM-16 Specifications

| Specification (18°C ≤ TA ≤ 28°C) | | | | |
|---|---|---|---|---|
| | Range | Offset Uncertainty | % of Value Uncertainty | Resolution |
| **Voltage** | 250mV | 250µV (50µV) | 0.03 | 7.8125µV |
| | 500mV | 250µV (50µV) | 0.03 | 15.625µV |
| | 1V | 300µV (75µV) | 0.03 | 31.25µV |
| | 2.5V | 500µV (100µV) | 0.03 | 78.125µV |
| | 5V | 1mV (200µV) | 0.03 | 156.25µV |
| | 10V | 2mV (400µV) | 0.03 | 312.5µV |
| | 25V | 5mV (1mV) | 0.03 | 781.25µV |
| | 50V | 10mV (2mV) | 0.03 | 1.5625mV |
| | 100V | 20mV (4mV) | 0.03 | 3.125mV |
| **Current** | 100nA | 100pA | 0.20 | 3.125pA |
| | 1µA | 300pA | 0.15 | 31.25pA |
| | 10µA | 2nA | 0.05 | 312.5pA |
| | 100µA | 20nA | 0.05 | 3.125nA |
| | 1mA | 200nA | 0.05 | 31.25nA |
| | 10mA | 2µA | 0.05 | 312.5nA |
| | 100mA | 20µA | 0.05 | 3.125µA |
| | 1A | 200µA | 0.1 | 31.25µA |

## Specification Comments:

- Maximum output current on 1A range is ±200mA. On other ranges, the maximum is 125% of range.
- Settling time to .01%: 100nA range 4 msec; 1µA range 2.3 ms; 10µA through 1A ranges 1.7msec; Volts Mode: 1.6msec to .01%.
- CMRR Voltage: 5µV/V (106dB)
  CMRR Current: 1 ppm of range per volt, 10µA-1A
  2 ppm of range per volt, 1µA
  6 ppm of range per volt, 100nA
- Accuracy on the lowest three current ranges is determined with digital averaging approximating line cycle integration.
- Accuracy of current measured on a given range is proportional to range error and a percentage of current being measured. For example, measuring 50µA on the 100µA range:
  50µA ± (20nA + .05% of 50µA) = 50µA±45nA
- Range Error shown in parentheses ( ) applies for an 8-hour period after auto zero, and for ±1°C.
- When measuring currents from sources with non-zero output conductance, add the following amounts to the error specification:

$$\pm\frac{830\,ppm\ \ of\ \ value + 151\mu A}{mho}$$