Digitally Synthesized Power Calibration Source

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***Abstract*—A digitally synthesized source of "phantom" power for calibrating electrical power and energy meters is described. Independent sources of voltage, current, and phase angle are programmable between 0 and 240 V, 0 and 5 A, and 0 and 360 deg, respectively. The accuracy of the active and reactive power is estimated to be within ±100 ppm of the full-scale apparent power (volt-amperes).**

I. Introduction

Calibration of wattmeters and watt-hour meters has traditionally been made by comparing the meter under test (MUT) to a standard wattmeter or watt-hour meter. The advantage of this approach is that a precise knowledge of the source parameters is not required. Voltage and current amplitudes and the phase angle between them need only be known approximately and the stability of each of these parameters is not critical as long as the power or energy output of the MUT is averaged or integrated over the same period as the standard instrument. With the advent of multifunction instruments capable of measuring voltage, current, power factor, and active and reactive power, a knowledge of each of the source parameters has become advantageous. The measurement of reactive power and energy, in particular, is greatly simplified if the voltage, current, and phase angle are known and stable.

This approach has led to the development of a dual-channel sine-wave source of voltage and current which is shown in Fig. 1. Previous experience [1]–[6] led to the selection of a digital waveform generator to synthesize two separate low-level sinusoidal voltages which are programmable in amplitude and phase angle. A special voltage amplifier A1 was designed to scale the low-level voltage V_1 to typical test levels ranging from 60 to 240 V, while test currents ranging from 1 to 5 A are obtained with a specially designed transconductance amplifier $A2$ [7]. The source is controlled by a desktop computer which is linked to auxiliary instrumentation for measuring the analog and digital outputs of the MUT.

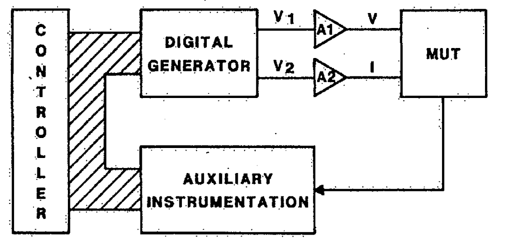

Fig. 1. Block diagram of the power calibration source.

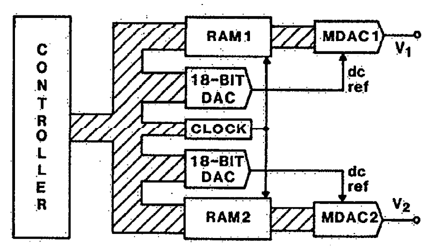

Fig. 2. Block diagram of the dual-channel digital generator.

II. Digital Generator

The heart of the power calibration source is a digital generator (see Fig. 2) which synthesizes voltages $V1$ and $V2$ in a staircase or "zero-order-hold" approximation. Two independent waveforms are reconstructed by sequentially applying digital values stored in RAM1 and RAM2 to 18-bit multiplying digital-to-analog converters (DAC's), MDAC1 and MDAC2. The values stored in memory are equally spaced samples of this waveform with up to 2048 discrete steps per period. For normal operation the stored functions are sine waves; however, any arbitrary waveforms with up to 1024 harmonic components may be stored. The phase angle between the two waveforms is modified by changing the set of function values stored in RAM2.

The theoretical resolution of the phase angle separating a pair of digitally synthesized sine waves is a function of the resolution of the processor used to calculate the sample points, the resolution of the generating DAC's (this determines to what extent each step is quantized), and the number of steps (sample points) per period. The algorithm used to calculate the sample points is performed with adequate precision in the computer to introduce negligible errors. The generating MDAC's, capable of 18-bit precision, are normally used as 16-bit converters to speed up data transmission. These MDAC's may be updated at 8-μs intervals, thus at 60 Hz, 2048 steps are used to synthesize one period. The angular resolution under these conditions (based upon computer simulations) is approximately 1 μrad.

The amplitudes of V_1 and V_2 may be independently set

Manuscript received June 23, 1986.
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IEEE Log Number 8613432.

between 0 and 10 V-peak by controlling the dc reference voltages supplied to MDAC1 and MDAC2 with a second pair of 18-bit DAC's. This technique provides an amplitude resolution of approximately 38 μV ($10/2^{18}$ V). The offsets of all four DAC's may be adjusted remotely over a range of ± 500 ppm by employing four additional 8-bit DAC's. This technique provides a software trim for the dc offset and gain of each of the generated waveforms.

The frequency of the generated waveforms is a function of the number of steps per period and the sample rate (for a 60-Hz sine wave with 2048 steps the sample rate is 122.88 kHz). A programmable frequency synthesizer (clock) provides the sample strobe signal with a short-term stability of approximately 1 part in 10^8.

III. Software

The source is controlled by a desktop computer with menu-driven software written in Basic. Selectable functions are given as follows.

1. Phase: A routine used to set the phase relationship between the variable (current) channel and the reference (voltage) channel by recalculating the data set stored in RAM2. The new phase angle may be entered through the keyboard or incremented/decremented in 1 or 10 μrad steps with special function keys.

2. Offset: A routine used to set a correction angle to compensate for differential phase shifts between the two channels. Once adjusted, the phase settings are direct reading. Angle entry is similar to "phase."

3. Frequency: A routine used to set the fundamental frequency of the generated waveform by programming the internal frequency synthesizer. Again, entry is made by typing in the desired frequency or incrementing/decrementing in 10- or 100-ppm steps with special function keys.

4. Steps: A routine linked with "frequency" used to set the number of steps per period of the generated waveform. Steps ranging from 8 to 2048 (in binary increments) are selected with special function keys. While the "frequency" and "steps" routines could have been combined into one routine which selects an optimum number of steps and the synthesizer frequency, separate entries for each of these parameters provides maximum versatility.

5. Amplitude: A routine used to set the voltage and current amplitudes by adjusting the dc references to MDAC1 and MDAC2. Additional special function keys are provided to set the gain and dc offset of each of the generated waveforms. The resolution of each entry is approximately 4 ppm of the full-scale amplitude.

6. Waveform: A routine used to independently select the waveform of each channel. Special function keys are provided to select sine waves, square waves, triangle waves, ramps, or a user-generated arbitrary waveform which may be any analytic function (with frequency components limited by the number of steps per period).

7. Test: A routine used to measure and record the analog and digital outputs of up to six watt/watt-hour, var/var-hour meters with the use of an external DVM and frequency counter. This routine also calculates and prints the test results.

Additional functions allow the user to perform linearity adjustments on the waveform generator as well as to characterize and store correction coefficients for the entire system.

IV. High-Voltage Amplifier

A precision high-voltage power amplifier (to be described in a later paper) with a fixed gain of 40 was designed to boost the output voltage V_1 of the reference channel of the dual-channel digital generator to a nominal test voltage of 120 or 240 V rms. The primary design objective was to maintain the good amplitude and phase stability inherent in the digital generator. In addition, the amplifier must be capable of supplying approximately 100-mA rms to accommodate the burden requirements of electrodynamic-type meters without significant error.

A simplified circuit topology of the amplifier is shown in Fig. 3. The circuit is best understood by realizing the general symmetry of the output power stage. Two FET pairs operate on alternate polarities of the signal in what is often referred to as a "totem pole," push-pull, class B driver. Transistor pair $Q1$, $Q2$ drives the positive portion of the signal into the output load while transistor pair $Q3$, $Q4$ drives the negative portion of the signal. The transistors used for this application are N-channel power MOSFET's with 1000-V ratings. A polarity separator circuit at the output of $U1$ separates the signal into positive and negative components and steers the respective polarized signals to two dual opto-isolators. The two opto-isolators serve to couple the two polarized signals in a differential manner across the large common-mode voltage between the high-level output circuit and the low-level input circuit. The output voltage is fed back through $R2$ where it is compared with the input voltage from $R1$ at the summing junction of $U1$. Thus with sufficient open-loop gain the closed-loop gain becomes operationally equal to $-(R2/R1)$.

The differential amplifier which follows each dual opto-isolator was designed with discrete components. It differs from the usual integrated circuit differential amplifier in that the input is designed to have a low impedance which maximizes the bandwidth of the opto-isolators. A bias control on each differential amplifier is used to set the quiescent bias current through both output driver pairs.

The amplifier is protected against excessive load currents or short circuits as follows. As the load current increases the voltage drop across R_s rises until the drain-to-source voltage of $Q2$ or $Q4$ reaches saturation, at which point the gain drops sharply. The output voltage of $U1$ rises to try to overcome the loss in gain, and a bistable overload sensing circuit senses the increased voltage and disconnects the supply voltages from the output drivers.

The average performance characteristics observed o four amplifier units are:

1) voltage gain: 40 (nom);

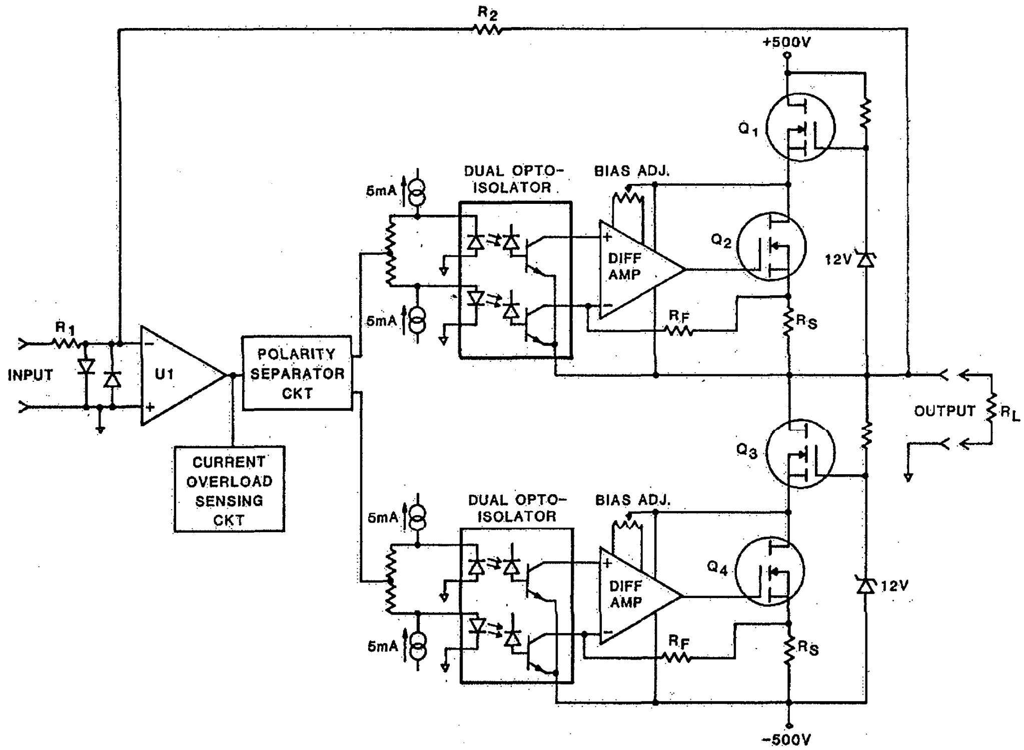

Fig. 3. Simplified diagram of the voltage amplifier.

2) maximum output swing: ±485 V, (340 V rms);
3) maximum output current: ±145 mA, (100 mA rms);
4) frequency response: dc to 150 kHz, 3 dB;
5) load regulation: 20-ppm change for a 100-mA load change @ 1 kHz;
6) output noise: < 50 mV rms (500-kHz bandwidth);
7) output offset: < ±1 mV;
8) input impedance: 10 k; and
9) short-term amplitude and phase instability: ±5 ppm and ±5 μrad (15-min interval).

V. Transconductance Amplifier

A precision wide-band transconductance amplifier [7] was designed to convert the output voltage of the variable channel of the dual-channel digital generator to a proportional current that remains essentially independent of the load terminal voltage. The gain factor is 1 A/V or 1 S so that the most common test current of 5 A rms is obtained with a 5-V rms input. Maintaining good amplitude and phase stability were the main design objectives for this amplifier.

A simplified diagram of the transconductance amplifier is shown in Fig. 4. The operation of the circuit can be described as follows. A voltage V_{in} applied at the input terminal causes an output current I_o through the load and a voltage drop across the shunt resistor R_s. This drop is amplified by differential amplifier U_2, and its output is fed back through R_2, where it is compared with the input voltage through R_1 at the summing junction of U_1.

Thus, in an operational sense, the output current is made proportional to the input voltage. A current booster circuit (made up of discrete components) between the output of U_1 and the output load circuit provides current amplification and high-output current drive levels. An auxiliary output current monitor from the output of U_2 provides a convenient ground-referenced voltage proportional to the output current. Assuming idealized amplifiers and components, the equation governing the output current is

$$I_0 = -V_{in}\left(\frac{R_2}{10R_1R_s}\right) \qquad (1)$$

where the factor of 10 is the gain of differential amplifier U_2. Thus, with a shunt resistance of 0.1 Ω and $R_2 = R_1$, the transconductance (I_o/V_{in}) has the convenient value of 1 S. Also, with the same parameters, the output of U_2 develops a ground-referenced voltage proportional to current with a scale factor of 1 V/A.

One of the more critical elements of the amplifier is the shunt resistor. The quality of the shunt has a direct bearing on the overall accuracy, stability, and bandwidth of the amplifier. For the purpose of this amplifier, an inexpensive, surprisingly good quality, low-reactance, four-terminal shunt was constructed by paralleling a large number of low-power metal-film resistors between two

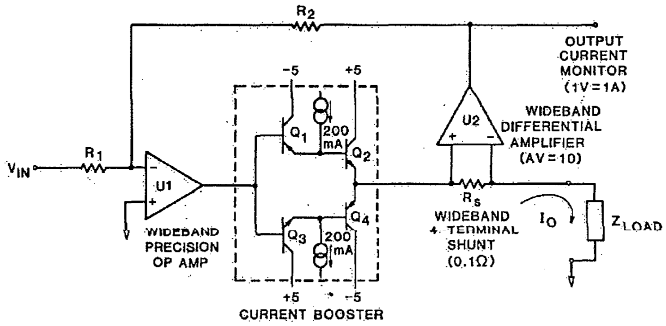

Fig. 4. Simplified diagram of the transconductance amplifier.

copper plates. A 0.1-Ω shunt was constructed by sandwiching 100 10-Ω, 0.25-W, 2-percent metal-film resistors between two 0.25-mm-(0.01-in) thick copper plates on 5-mm (0.2-in) centers in a 10 × 10 matrix.

The average performance characteristics of the transconductance amplifier are listed below:

1) transconductance: 1 S;
2) maximum output current: 8 A rms;
3) frequency response: dc-150 kHz (3 dB) @ 5 A rms;
4) load regulation: 45-ppm change for a 1-V compliance voltage change @ 5 A, 60 Hz;
5) compliance voltage: 2 V rms @ 5 A rms;
6) output offset current < ±150 μA;
7) input impedance: 5k Ω; and
8) short-term amplitude and phase instability: ±5 ppm and ±8 μrad.

VI. Auxiliary Instrumentation

Most commercial standard wattmeters provide an analog output in the form of a dc voltage or current which is proportional to the measured power. In addition, many instruments provide a pulse train output, whose frequency is proportional to the measured power, which allows the user to measure energy simply by counting the number of output pulses. Energy calibrations are normally performed by totalizing the pulses of both the standard and test instruments until sufficient resolution is achieved and test periods often exceed 4 min. One of the advantages of testing with a stable power source is that a high-resolution frequency or period measurement can replace the lengthy totalizing scheme without degrading measurement accuracy.

Additional hardware to perform these measurements includes a set of programmable switches which accepts inputs from up to six test meters, a high-accuracy DVM, and a frequency counter. Power and energy measurements can be made simultaneously in intervals of 10 s per meter.

VII. Performance

The power source was originally intended to operate at 60 Hz with 120-V and 5-A sinusoidal waveforms. However, the present source is programmable between 0 and 240 V and 0 to 5 A with sinusoidal as well as arbitrary

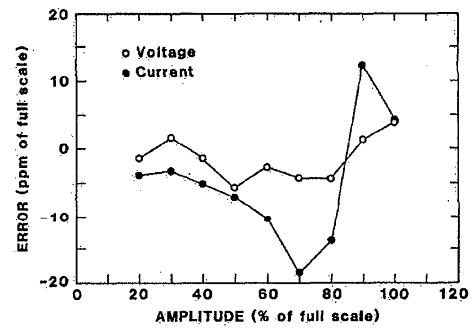

Fig. 5. Residual voltage and current integral nonlinearity after gain corrections.

waveforms at frequencies between 0.001 Hz and 100 kHz. These figures represent the limits of amplitude and frequency. Measurements, described in this paper, were performed at 60 Hz and at amplitudes between 20 and 100 percent of full scale (FS). The amplitude and phase-angle errors given below were obtained by measuring the source using a thermal wattmeter [8] and a current-comparator power bridge [9].

The amplitudes of the reference and variable channels are changed by adjusting the dc voltages supplied to MDAC1 and MDAC2. However, the output voltages V_1 and V_2 are not ideal linear functions of these dc voltages and thus a gain adjustment is required at different amplitudes. Software gain corrections for any amplitude (based on a linear fit to a few data points) reduce this voltage-dependent gain error by a factor of 5 to 10. Fig. 5 shows the residual amplitude nonlinearity, after correction, over a 5-to-1 amplitude range where 100 percent of FS represents 240 V and 5 A, respectively. Differential nonlinearity around 120 V and 5 A is shown in Fig. 6. The sample points represent a one least-significant-bit change (4 ppm of FS) of the respective scaling DAC's.

Phase-angle accuracy depends upon the initial offset (differential phase shift between channels) and the phase linearity, which is a function of quantization errors, DAC nonlinearity, and the number of steps per period. The initial phase offset between the voltage and current waveforms was measured by the power bridge and adjusted to zero at unity power factor (0°) with an uncertainty of less

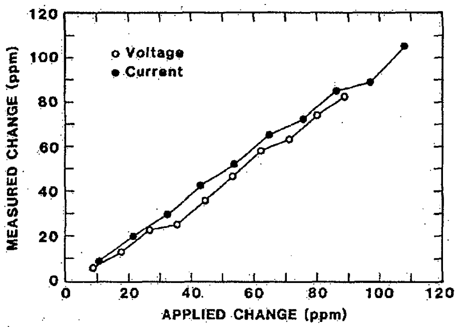

Fig. 6. Voltage and current differential nonlinearity around 120 V and 5 A.

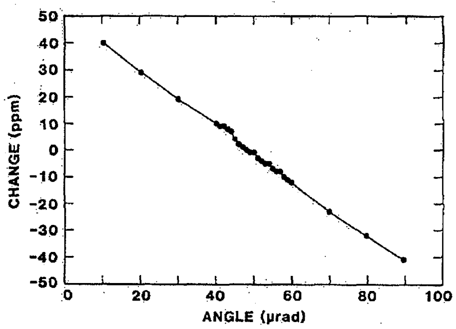

Fig. 7. Phase differential nonlinearity: generator angle versus change in power indication of a TDM wattmeter at zero power factor.

than 10 μrad. Subsequent power measurements, with constant voltage and current, between $+90°$ and $-90°$, indicate that the integral phase nonlinearity in this range is less than 20 μrad.

Fig. 7 shows the differential phase linearity at zero power factor (90°) as measured by a time-division-multiplier (TDM) wattmeter [10]. This plot not only confirms the computer simulation predictions of 1-μrad phase resolution of the digital generator, but demonstrates the potential of TDM wattmeters for performing extremely precise measurements around zero power factor as well.

Once the three parameters (voltage, current, and phase angle) have been adjusted and corrected, the major concern becomes stability. Measurements at 120 V, 5 A over a three-week period are given in Fig. 8. The precision of these measurements was approximately 10 ppm in amplitude and 10 μrad in phase. The current drift of 80 ppm has been attributed to aging of the 0.1-Ω shunt in the transconductance amplifier. A simple dc calibration of this amplifier is useful in detecting the gain drift due to the shunt and the results may be applied as an additional gain correction to improve long-term current stability.

Finally, the source was evaluated over a three-week period as a power calibrator. Measurements were performed at 120 V and 5 A at a number of phase angles between $\pm 90°$. The source was adjusted at the beginning of the testing period and used to calibrate a TDM wattmeter over the next 20 days without further adjustments. Measurements were also performed on the TDM wattmeter using the power bridge, and an envelope which encloses all of the differences between the source and the bridge, using the TDM wattmeter as a transfer standard, is plotted in Fig. 9. There is a direct correlation between these differences and the current drift from Fig. 8 as the data for both plots were collected during the same period. If corrected for this drift, the maximum power differences fall within a ± 30-ppm band, as shown in Fig. 10. These figures include the short-term drift of the TDM wattmeter between the source and bridge calibrations.

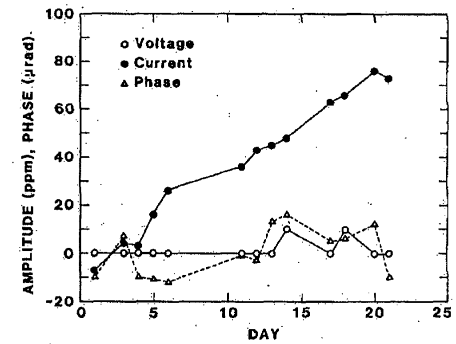

Fig. 8. Long-term stability of the source voltage, current, and phase angle as measured by a thermal wattmeter.

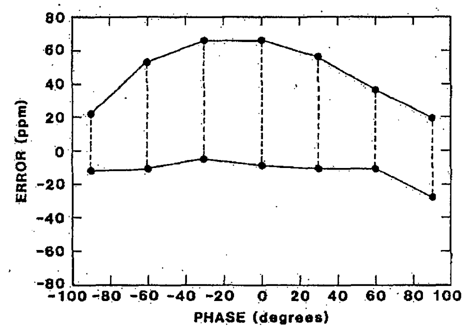

Fig. 9. Maximum differences between the power calibration source and the power bridge using a TDM wattmeter as a transfer standard.

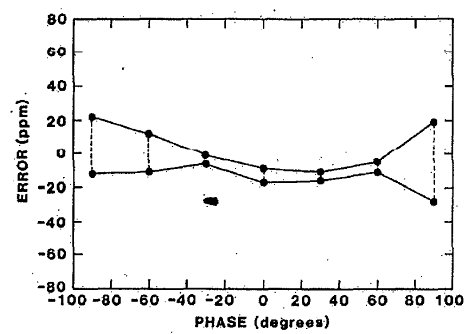

Fig. 10. Maximum differences between the power calibration source and the power bridge after correcting for the current drift.

VIII. Conclusions

An accurate and precise source of synthetic power for calibrating watt/watt-hour and var/var-hour meters at the 100-ppm level has been described. The source consists of a dual-channel digital waveform generator followed by direct-coupled, high-voltage, and transconductance amplifiers to provide signal levels of 60–240 V and 1–5 A at any phase angle. Control is provided by a desktop computer and auxiliary instrumentation supports the calibration of up to six test instruments. The uncertainties of source parameters at power frequencies are:

1) voltage < 30 ppm of FS;
2) current < 50 ppm of FS (requires a periodic monitor of the current amplifier gain);
3) phase < 20 μrad at FS voltage and current (degrades slightly at lower amplitudes); and
4) power (active and reactive) < 100 ppm of FS volt-amperes.

While the source is normally operated under sinusoidal conditions, future applications will utilize its ability to synthesize arbitrary waveforms with dc components. A direct coupled system permits calibration of gain factors at dc where, in general, the resolution and accuracy of measuring instrumentation is greater. Extrapolating results at dc to power frequencies is reasonable and offers the possibility of a calculable ac source based on dc measurements. Furthermore, experience with a direct coupled system has focused attention on the ever present ground loops that always seem to evolve in a system. In fact, monitoring and eliminating dc offsets created by ground loops has turned out to be a good technique for assuring that ground loops do not create measurement errors which might otherwise go undetected.

Acknowledgment

The authors wish to thank R. Palm for his valuable assistance in fabricating printed circuit boards and assembling the source components, and Dr. R. Turgel for providing general consultation as well as computer models to analyze phase performance.

References

[1] J. R. Kinard and L. A. Harris, "Wattmeter calibration at zero power factor using digitally generated sinewaves," *IEEE Trans. Instrum. Meas.*, vol. IM-25, no. 4, Dec. 1976.
[2] R. S. Turgel and N. M. Oldham, "High-precision audio-frequency phase calibration standard," *IEEE Trans. Instrum. Meas.*, vol. IM-27, no. 4, Dec. 1978.
[3] N. M. Oldham and R. S. Turgel, "A power factor standard using digital waveform generation," *IEEE Trans. Power App. Syst.*, vol. PAS-100, no. 11, Nov. 1981.
[4] C. Vastrade, R. DeVre, M. Couvreur, and R. Huybrechts, "Calibration of high accuracy wattmeters with a standard generator," *IEEE Trans. Instrum. Meas.*, vol. IM-34, no. 2, June 1985.
[5] H. Ahlers, L. Rahf, H. Hartwig, and J. D. Sievert, "Programmable digital two-channel function generator for testing power meters," *IEEE Trans. Instrum. Meas.*, vol. IM-34, no. 2, June 1985.
[6] N. M. Oldham, "A power calibration standard based on digitally synthesized sinewaves," *IEEE Trans. Power App. Syst.*, vol. PAS-104, no. 11, Nov. 1985.
[7] O. B. Laug, "A wide-band transconductance amplifier for current calibrations," *IEEE Trans. Instrum. Meas.*, vol. IM-34, no. 4, Dec. 1985.
[8] G. Schuster, "Thermal instrument for measurement of voltage, current, power, and energy at power frequencies," *IEEE Trans. Instrum. Meas.*, vol. IM-29, no. 3, Sept. 1980.
[9] N. M. Oldham and O. Petersons, "Calibration of standard wattmeters using a capacitance bridge and a digital generator," *IEEE Trans. Instrum. Meas.*, vol. IM-34, no. 4, Dec. 1985.
[10] P. Miljanic, B. Stojanovic, and P. Bosnjakovic, "The development of a high precision time-division power meter," in *CPEM 84 Dig., Proc. Conf. Precision Electromagnetic Meas.* (Delft, the Netherlands), 1984, pp. 67–68.