RMS-DC CONVERTERS

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Abstract

RMS or Root Mean Square is a fundamental measurement of the magnitude of an AC signal. RMS converters are used in voltmeters for RMS voltage or current measurement. The principles of the computational method using the mathematical relation are mostly used in modern technology in integrated technique. But, the basic definition for measuring true RMS value is defined as the amount of the power dissipated on known active load. True RMS-DC converters can be built using pair of thermocouples. This solution has disadvantages due to the inertness of the thermocouples, that can be eliminated by substituting the thermocouples with diodes or transistors.

In this paper are described RMS-DC converters, realized on the computational method and thermal conversion, the both methods are compared and in both cases are achieved similar results.

Key words: RMS, converter, thermocouples

1. INTRODUCTION

The RMS value is a significant parameter which describes the energy content of a signal. RMS measurement techniques based on a signal's peak or average magnitude value may offer circuit simplicity but do not yield accurate results for nonsinusoidal signals.

One popular technique for RMS measurement is the computational method, which uses the mathematical relation for determine the RMS voltage or current value.

The other method uses thermal conversion, a technique for RMS measurement is to convert the energy of the signal to be measured into heat and then develop and measure the DC signal which produces an equal amount of heat. For example, an AC signal of 1V RMS will produce the same amount of heat in a 1V DC signal. True RMS converters can perform with high accuracy $\pm 0.05\%$ (to 10MHz) and $\pm 2\%$ (to 100MHz).

2. RMS-DC CONVERTER BASED ON COMPUTATIONAL METHOD

RMS-DC converter based on the computational method is defined with the mathematical relation:



This means squaring the signal, taking the average, and obtaining the square root, as it is shown on Figure 1.



The direct method of computation has a limited dynamic range because the stages following the squarer must try to deal with a signal that various enormously in amplitude. These practical limitations restrict this method to inputs which have a maximum of approximately 10:1 dynamic range.



Generally better computing method uses feedback to perform the square root function indirectly at the input of the circuit as shown in Figure 2.

Fig. 2. Computational method using feedback (Implicit method)

Divided by the average of the output, the average signal levels now vary linearly with the RMS level of the input. This increases the dynamic range of this circuit as compared to explicit RMS circuits.

AD637 converter uses an implicit method of RMS computation, has high accuracy, an extended frequency response. On figures 3 and 4 are presented simulated output waveforms for different input waveforms(triangle and square input voltages) for AD637.



This converter uses inverting low pass filter stage to provide a buffered output voltage whose averaging time constant is independent of the input signal level. The averaging time is the which time in the RMS converter holds the input signal during computation; it directly affects the accuracy of the RMS measurement.

Fig. 3 Simulated output waveform for triangle input waveform





True RMS measurement is a universal language among waveforms, allowing the magnitudes of all types of voltage or current waveforms to be compared to one another and to DC. These converters are smart rectifiers; they provide an RMS reading accurate regardless of the type of waveforms being measured.

Fig. 4 Simulated output waveform for symmetrical square input waveform

As an example, if an average responding converter is calibrated to measure RMS value of sine-wave voltages, and then is used to measure either symmetrical square waves or DC voltages, the converter will have a computational error 11% higher than the true RMS value.

3. TRUE RMS-DC CONVERTER BASED ON THERMAL CONVERSION

In this paper is described a thermal technique of RMS measurement using resistortransistor chip; the base-emitter junction of a bipolar transistor is used to sense the change in temperature of the chip due to the power dissipated by a resistor (Figure 5).



Fig.5 RMS-DC Converter using a pair resistor-transistor

The power dissipated by the heater resistor R_1 due to the input signal U_i heats the input pair producing change in the base-emitter voltage of T_1 . This generates an error voltage which is amplified by the operational amplifier A_1 , so the output voltage heats the resistor R_2 . In moment when the potentials of the operational amplifier differential input are equal, the circuit is in equilibrium so $U_{0rms} = U_{irms}$. Then DC output voltage is proportional to the AC input voltage.

From the Ebers - Moll relationship for forward biased junction

$$V_{BE} = \frac{kT}{q} \ln \frac{I_E}{I_S},\tag{1}$$

 I_s is the saturation current.

$$I_s = BT^3 e^{-\alpha V_{bo}/kT}, (2)$$

By combining (1) and (2) the junction equation can be written in terms of the physical constants and the base-emitter voltage at a specific emitter current and temperature (I_{E0} and T_0).

$$V_{BE} = \frac{kT}{q} \ln \frac{I_E}{I_{E0}} \left(\frac{T_0}{T}\right)^3 + \frac{T}{T_0} \left(V_{BE0} - V_{b0}\right) + V_{b0}$$
(3)
$$V_{BE0} = V_{b0} \text{ at } I_{E0} \text{ and } T_0$$

Differentiating (3) with respect to temperature yields the junction temperature coefficient, and for constant emitter current is

$$\frac{dV_{BE}}{dT}\Big|_{I_E=const} = \frac{V_{BE0} - V_{b0}}{T_0} - \frac{3k}{q} \left(1 + \ln\frac{T}{T_0}\right)$$
(4)

The temperature coefficient is $-2mV/^{\circ}C$ and to have nonlinearity of less than 2% for temperature between 0 and 100°*C*.

Neglecting the effects of the sense transistor base currents in the base resistors R_b , the base-emitter voltages of the sense transistors can be related as

$$V_{be1}(s) = V_{be2}(s) + H(s)U_0(s)$$
(5)

where H(s) is the AC feedback transfer function:

$$H(s) = \frac{sR_bC_fA_2}{1 + sC_f(R_b + R_f)}$$
(6)

Combining (5) and (6) yields

$$U_0^2(s) = \frac{A_{T01}(1+s\cdot\tau_{T2})}{A_{T02}(1+s\cdot\tau_{T2})} \frac{U_i^2(s)}{1+s\tau_f} + \frac{V_{BE1} - V_{BE2}}{A_{T02}}$$
(7)

where A_{T0} is the thermal gain of the transistors. Its typical value is $A_{T0} = 10 \, mV / V^2$, and τ_T is thermal time constant. τ_f is a filter time constant defined as:

$$\tau_f = \frac{R_b C_f A_2}{A_{T02} U_0(s)}$$
(8)

The frequency dependence of the RMS-DC converter may be nonlinear since the transfer function time constant is inversely proportional to the output voltage if A_2 is linear. This nonlinear frequency dependence produces an unsymmetrical step response and a low frequency cutoff for accurate RMS converter which is proportional to the RMS value of the input. These characteristics are undesirable if fast response is necessary and a nonlinear AC feedback should be employed.

By making A_2 a square law amplifier

$$A_2 u_0(t) = \rho \cdot u_0^2(t)$$
(9)

The time constant will be independent of the output; combining (8) and (9) the filter time constant for square law A_2 , τ_{fs} is

$$\tau_{fs} = \frac{R_b C_f \rho}{A_{T02}} \tag{10}$$

For sinusoidal inputs $u_i(t) = \sqrt{2} \cdot U_{irms} \cos \omega t$, the output as a function of the input frequency is

$$u_{0}(t) = U_{irms} \cdot \left[1 + \left[\frac{1 + 4\omega^{2} \tau_{T1}^{2}}{(1 + 4\omega^{2} \tau_{T1}^{2})(1 + 4\omega^{2} \tau_{fs}^{2})} \right]^{1/2} \cos^{2} \omega t \right]^{1/2}$$
(11)

This is valid only for filter time constant which has square law amplifier $A_2u_0(t) = \rho \cdot u_0^2(t)$, as it is shown on figure 6b. However, an approximate solution for linear A_2 can be obtained by relating the output of the linear amplifier to that of the square law amplifier

$$U_f = A_2 U_0 \quad U_{fac} = A_2 U_{0ac} \text{ for linear } A_2 \tag{12}$$

$$U_{f} = \rho \cdot U_{0}^{2} U_{fac} \approx \rho U_{0dc} U_{0ac} \text{ for square law } A_{2}$$
(13)

The RMS value of the ripple at the output is then

$$U_{ripple} = \frac{\sqrt{2A_{T02}E_{0dc}E_{irms}}}{8\pi f R_b C_f A_2} \quad \text{linear } A_2 \tag{14}$$

$$U_{ripple} = \frac{\sqrt{2}A_{T02}E_{irms}}{16\pi f R_b C_f \rho} \text{ square law } A_2$$
(15)

In figure 6(a) the RMS value of the output ripple is simulated versus the RMS value of the input for an input frequency of 20Hz. As, expected, the output ripple with linear AC feedback is proportional to the square of the RMS value of the input where it is linearly proportional to the RMS value of the input with square law AC feedback. The square law does, therefore, provide a low frequency cutoff which is independent of the signal level.



Fig.6 (a) Output ripple versus input for a 20Hz input for both linear and square A_2 (b) Output voltage as a function of time for sinusoidal input signal $R_b = 1k\Omega, C_f = 2\mu F$, and $A_2U_0 = U_0$ or $A_2U_0 = 0.25 \cdot U_0^2$)

4. CONCLUSION

Some advantages of implicit computation (AD637) over other methods are fewer components, greater dynamic range and generally low cost. A disadvantage has less bandwidth than either thermal or explicit method. True RMS-DC converters are capable of producing high accuracy over a wide range of levels, waveforms and frequencies.

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