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Algorithmic Optimization of Spectral and Temperature Characteristic of MTCXO

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***Abstract* - The creation of a digital thermocompensation system is possible now on a basis of single chip of typical microcontroller. In paper the problems of assuring of clean spectrum characteristics of DTCXO output signal on the basis microcontroller at usage of the cheapest DAC solution - PWM are considered. The variants of control voltage shaping accuracy increasing are considered at the expense of second and higher orders PWM realization. The optimization of choice algorithm order is carried out. Besides, the variants of ADC accuracy increasing are described at the expense of special algorithms of statistical handling.**

1. INTRODUCTION

The basic requirements for crystal oscillators used as sources of reference oscillations can be formulated as:

- high frequency stability
- low level of noises
- small overall dimensions
- small power consumption
- short setting time
- low cost

When the overall dimensions and consumed power is the determinative factor (for example, in mobile equipment) thermocompensated crystal oscillator with analogue thermal compensation (TCXO) are widely used. However, the thermal stability of such oscillators normally does not exceed 1 - 2 ppm in industrial temperature range. More stable oscillators are very laborious in tuning and much more expensive. The thermal frequency stability can be improved with using of digital methods of thermal compensation. Besides, oscillators with digital thermal compensation (DTCXO) are much more simple in tuning. In literature DTCXO with stability 0,1 - 0,3 ppm [1 - 5] are described. The circuit diagrams of typical DTCXO is shown on Fig. 1.

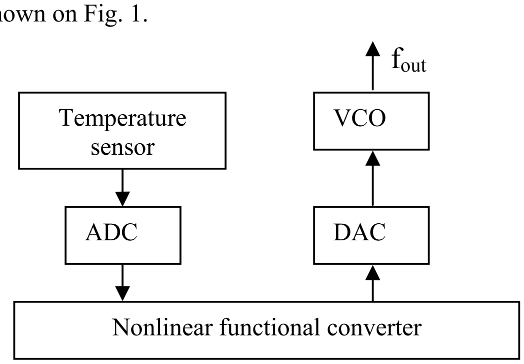

Fig. 1. DTCXO with analogue sensor

As a nonlinear functional converter both ROM end microcontroller must be used. Unfortunately, until recently the element basis permitting to create a low-cost high-stable DTCXO was absent. Discrete ADC, DAC, ROM and other element (used for implementation of structure shown on Fig 1) had a high price and high overall dimensions [6].

The situation has changed, when the small-sized microcontrollers, having in their structure the majority of units, indispensable for implementation of a digital part of shown on Fig. 1 DTCXO, have appeared at market. Really, practically each microcontroller family have a model that incorporate a built-in ADC and memory, and some of most modern devices have even a built-in temperature sensor. As the most reasonable device it is possible to propose the ATtiny15 (by Atmel) or C8051F310 (by Cygnal). The parameters of these microcontrollers are listed in Table 1.

TABLE 1

	ATtiny15	C8051F310
ADC	10-bit 4 channel	8-bit up to 8 ch.
Sample speed	15 ksps	up tp 500 ksps
Thermo sensor	no	yes
DAC	no	no
ROM	1K Flash	8K Flash
RAM	64 bytes	256 bytes
MIPS	1.6	Up to 25
Internal oscillator	1.6 MHz	24.5 MHz
External oscillator	-	0 – 25 MHz
PWM	8-bit@150kHz	8-bit@24kHz 16-bit@95Hz
Supply voltage, V	2.7 – 5.0	2.7 – 3.6
Consumption	3 mA	5mA@25MHz 1mA@5MHz
Case	Soic - 8	11 pin
Footprint, mm	8x5.5	3x3

Only one external unit that it is necessary to use for realization of DTCXO under the circuit of Fig. 1 is DAC. In microcontroller applications for code to analog conversion a pulse width modulation (PWM) is widely use. Advantages of such method are high linearity of conversion and possibility of using of built-in modules, and imperfections are necessity of tight filtration of PWM pulse sequence and additional noise in a spectrum of an output signal. It is necessary to note that tight filtration leads to delay in compensation signal and, as the supervention, leads to dynamic error increasing [7]. A research of a possibility of using of mentioned above microcontrollers for DTCXO creation without usage of external DAC represents practical interest. For this purpose it is necessary to evaluate an achievable temperature stability and side band components of output signal spectrum

1. PWM OPTIMIZATION

First of all, it is necessary to define required PWM word length. If the range of a frequency vs. temperature characteristic in a temperature interval is δF, and the desirable temperature stability is δf, the required PWM word length can be determined as

$$n \ge \frac{1}{k \cdot l} \cdot \log_2 \left(\delta F / \delta f \right),$$

where k is operating ratio of control voltage ($k < 1$), l is relative part of general temperature stability, defined by discretization of of control voltage ($l < 1$). The nonlinearity of oscillator control characteristic is also taken into account in factor k. Usually $l = 0.23 \ldots 0.5$, $k = 0.5 \ldots 0.8$. Thus, in an industrial temperature range at using of AT-cut resonators for stability ±0.3 ppm obtaining, the PWM word length is necessary within the 10 - 12 bits.

If the clock frequency for PWM former is equal to f_c, on output of the PWM circuit there will be a time sequence with a pulse repetition rate of $f_{pwm}=f_c / 2^n$. The spectrum of this sequence is well known. However, in this case PWM sequence is filtered and produces a frequency modulation of VCO output signal. The spectrum of FM signal differs from a spectrum of a baseband signal and hardly depend on deviation index $k = \Delta f / F$, where F is modulating frequency, Δf is frequency deviation. In this case the modulating frequency is PWM frequency, and Δf is range of VCO frequency shift (in absolute units). A conclusion from here follows, that the modulation index will depend on nominal value of output frequency: the frequency is higher - the modulation index is more. Besides, the spectrum of a PWM sequence will depend on a relative pulse durations (on-off time ratio), namely will depend on PWM controlling code. The greatest value of a first harmonic level will be at on-off time ratio equal to two; hence, in the further accounts this worst case will be used. The spectrum of a VCO output signal with nominal frequency of 10 MHz, modulated by unfiltered PWM of ATtiny 15 internal module (8-bit, 150 kHz), is indicated on Fig. 2.

For a reducing of a harmonic component level this PWM should be passed through a low-pass filter. Using of the high order filters for such purposes is non-real, as it hardly complicates the circuit (in small-sized oscillators each element is on the account). Really, it is necessary to consider variants of using of 1-st or 2-nd order passive filters. At a choice of parameters of this filter it is necessary to take into account some constructive limitations. In particular, the total resistance of the filter cannot be very large, as it augments influence of a current leakage. The ceramic capacitors with large capacitance

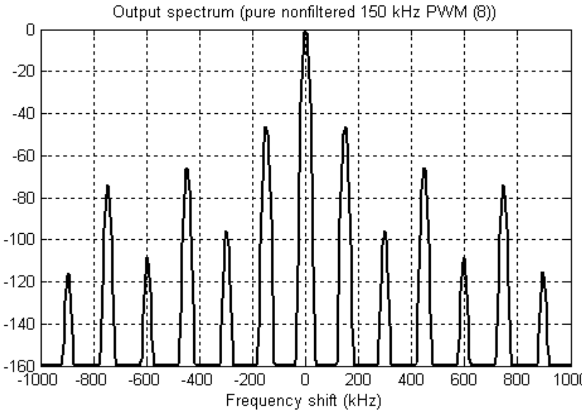

Fig. 2. 8-bit PWM spectrum

have great size, and electrolytic capacitors should not be applied because of instability and large noise. At optimization of a choice between filters of 1-th and 2 orders the criterion of equality of total sequential resistance and total parallel capacity was used. As limitations the value of 200 kOhm of total resistance and total capacities of 68 nF are accepted. From the accounts, the maximum suppression of given frequency by the filter of the second order with the mentioned above criterion of a choice of elements is achieved at equality of elements of the first and second line-ups (provided that the cutoff frequency of the filter differs from eliminated frequency more, than in 100 times). Proceeding from this, the circuits of used filters are indicated on Figs. 3a and 2b.

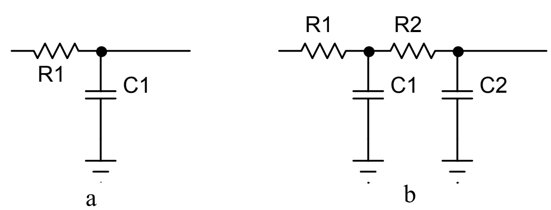

Fig. 3. LPF of 1-st and 2-nd order

On Fig. 4 the spectrum of a oscillator output signal is indicated at using of the 1-st order filter (doted line), and at using of the 2-nd order filter (solid line).

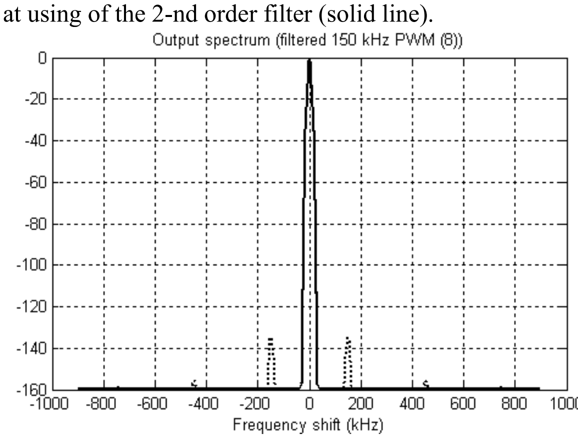

Fig. 4. Spectrum of filtered PWM sequence

It is well noticeably, that using of filter of the second order is preferable. It is necessary to note, that the data outcomes are obtained at using of 8-bit PWM with frequency of 150 kHz.
As it was mentioned earlier, for high stability obtaining, the 8-bit PWM will be insufficiently. If 16-bit PWM from C8051F310 will be used, the output signal spectrum will aggravate (Fig. 5) sharply.

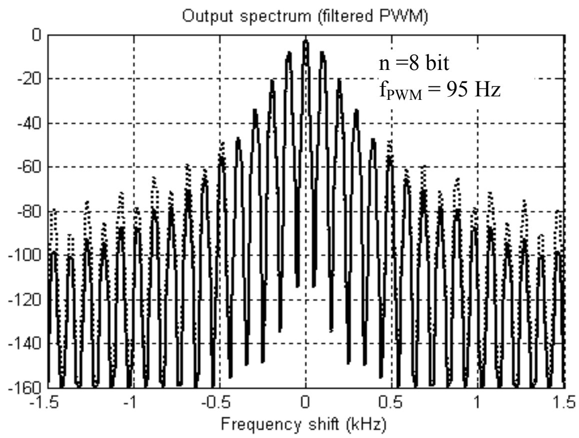

Fig. 5. Spectrum of 16-bit PWM

It is related with that a modulation index sharply increases, when the modulation frequency has decreased from 150 kHz to 95 Hz (more than in 1500 times). Apparently, that such high levels of spectral components are completely unacceptable.

For obtaining of a small index of modulation it is necessary to have enough high PWM frequency, therefore we shall use an 8 - bit PWM, and for providing an adequate accuracy we shall use an additional modulation of PWM code. In particular, to obtain DAC accuracy equivalent to 12 bit, it is necessary to make a modification a PWM code (from N to N+1) within the sequence of 16 impulses. Let's mean such sequence as “8+4” (8-bit PWM and 4-bit pulse modulation).

The modulated PWM (pulse-width modulation of the second order) is obtained by a following technique:

- the 12-bit code C is dissected into two parts: the 8-bit code N (most significant bits) and 4-bit code M (low bits), i.e. . $C=16\cdot N+M$ (N and M we shall name as parameters of PWM of the second order)
- two sets of PWM sequences are created:
  A) with M impulses of PWM of code N+1;
  B) with (16-M) impulses of PWM of code N.

The average constant component of such modulated PWM is:

$$\overline{U} = A\cdot\left(\frac{M}{16}\cdot\frac{N+1}{256}+\frac{16-M}{16}\cdot\frac{N}{256}\right)=\longrightarrow$$
$$\rightarrow A\cdot\left(\frac{N}{256}+\frac{M}{16\cdot 256}\right)=A\cdot\frac{16\cdot N+M}{4096}=\longrightarrow$$
$$\rightarrow A\cdot\frac{C}{4096},$$

where A - amplitude of PWM impulses.
The possible algorithm of second-order PWM shaping is shown at Fig. 6.

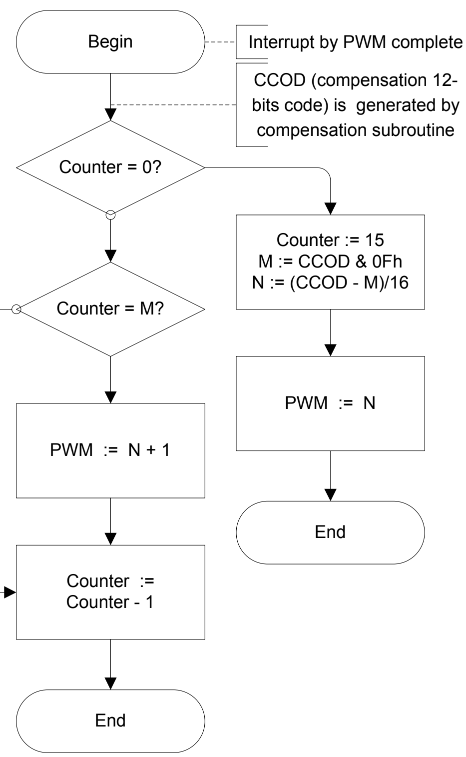

Fig. 6. Algorithm of 2-nd order PWM shaping

Let's execute the calculation of oscillator spectrum if such modulated PWM is used. In a Fig. 7 the oscillator spectrums for ATtiny15 are shown (dotted line is for 1-st order LPF, continuous line is for 2-nd order LPF). Spectrum is shown for near to carrier region.

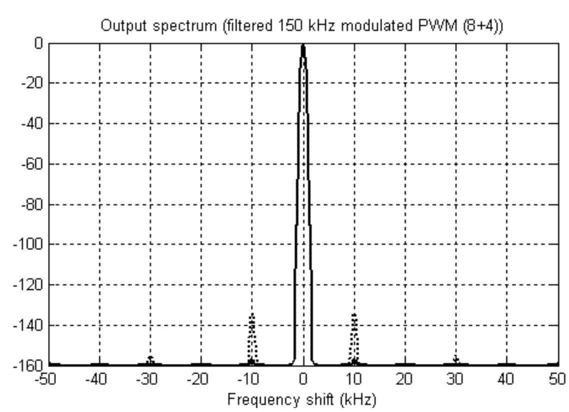

Fig. 7. Spectrum of 150 kHz 2-nd order filtered PWM

As it is visible from a Fig. 7, the levels of spectral components near to carrier are enough small (below -130 dB even for 1-st order LPF). However, short time of a PWM period (about 6 μs) at rather low productivity of ATtiny15 creates essential difficulties for program implementation of such modulated PWM. Therefore we shall consider variant of implementation of modulated PWM with usage of the microcontroller C8051F310. The spectrum of oscillator in “close to carrier” and in “far to carrier” zones are shown in Fig. 8 and Fig. 9 accordingly (dotted lines is for 1-st order LPF, continuous is for 2-nd order LPF).

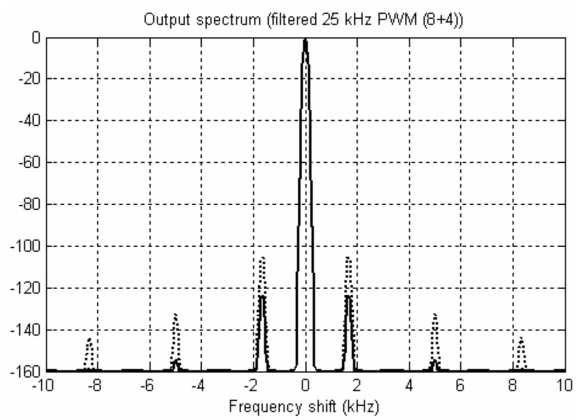

Fig. 8. Spectrum of 25 kHz 2-nd order PWM

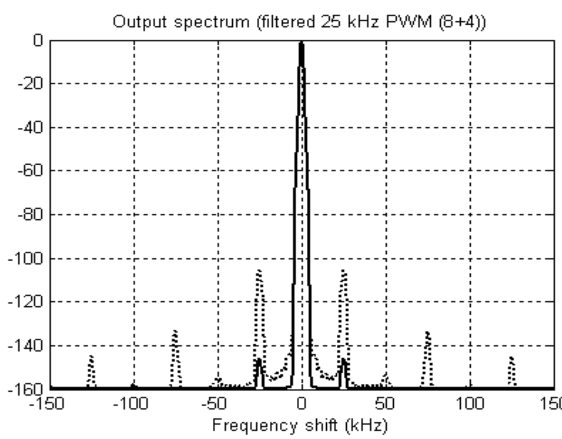

Fig. 9.Spectrum of 25 kHz 2-nd order PWM

As it is visible from these figures, at usage 2-nd order LPF is possible to obtain reasonable performances of output spectrum (below -120 dB).

It is necessary to note, that using of modulated PWM from a point of view of spectrum gives the best outcomes, than “pure” PWM of same equivalent digit capacity.

For comparison at Fig. 10 shown the spectrum of the oscillator with pure 12-bit PWM of the same clock rate, as at Fig. 8.

Here levels of spectral components are very high even at usage of a 2-nd order LPF. Thus, it is possible to make a conclusion that the modulation of PWM essentially influences on a level of spectral components of oscillator output signal.

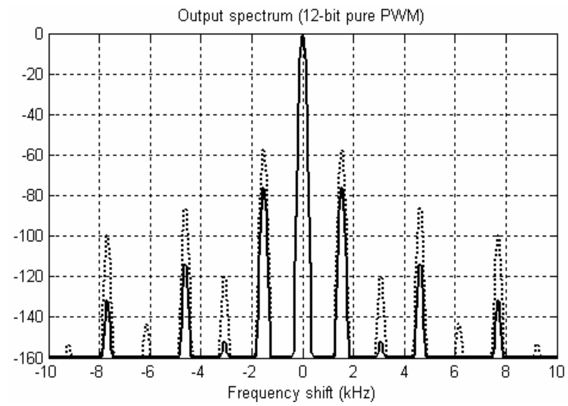

Fig. 10. Spectrum for 12-bit 25 kHz PWM

If the 12-bit code is dissected on a more parts, it is possible to obtain more complex kinds of PWM modulation. For example, if to dissect 12 bits in 3 parts, the third-order PWM will be obtain. This sequence may be produced on a following technique:

- The 12-bit code C is dissected into 3 parts: the 8-bit code N (most significant bits), Mp-bit code M (middle bits) and $(4\text{-}Mp)$- bit code K (low bits), i.e. $C=16\cdot N+2^{Mp}\cdot M+K$.
- Two sets of PWM sequences of the second order is created:
  - A) K sequences of the second order PWM with parameters N, $M+1$;
  - B) $2^{4\text{-}Mp}$ - K sequences of second order PWM with parameters N, M.

The possible algorithm of 3-rd order PWM realization is shown at Fig. 11

The average constant component of the third order PWM is:

$$\overline{U} = A \cdot \left(\frac{K}{2^{4-Mp}} \cdot \frac{2^{Mp} \cdot N + M + 1}{2^{8+Mp}} + \rightarrow \right.$$

$$\left. \rightarrow \frac{2^{4-Mp} - K}{2^{4-Mp}} \cdot \frac{2^{Mp} \cdot N + M}{2^{8+Mp}} \right) = \rightarrow$$

$$\rightarrow A \cdot \frac{16 \cdot N + 2^{4-Mp} \cdot M + K}{2^{12}} = A \cdot \frac{C}{4096}$$

For 3-rd order PWM it is possible few combination of N and M. For 8-bit basic PWM and 12-bit equivalent DAC next combination is realized: (8+3+1), (8+2+2), (8+1+3). Spectrum diagrams for each case mentioned above are shown at Figs. 12 – 14 respectively.

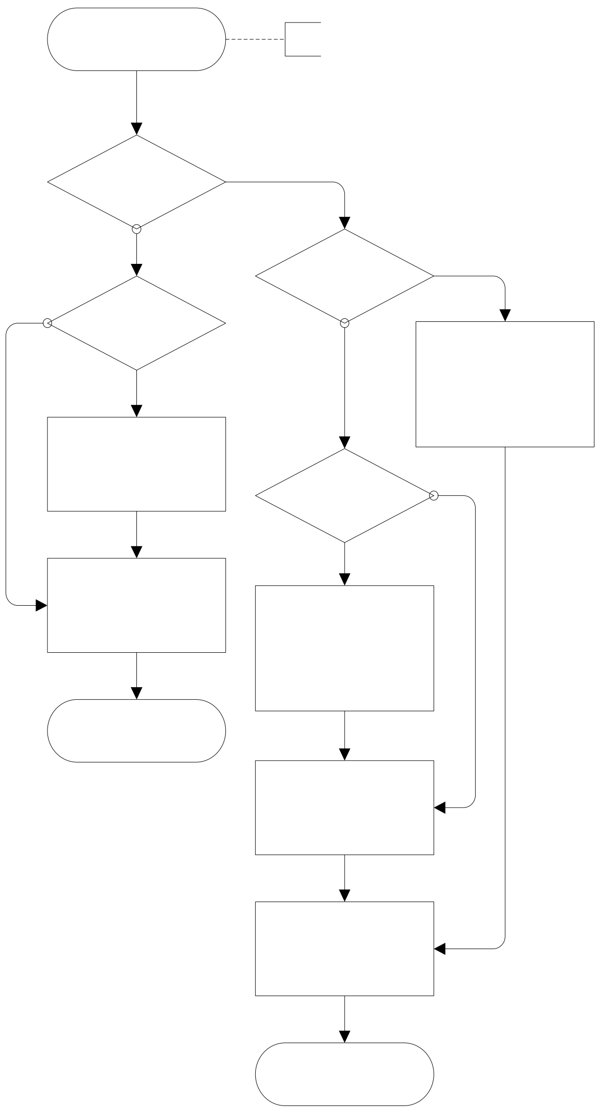

Fig. 11. Algorithm of 3-nd order PWM shaping

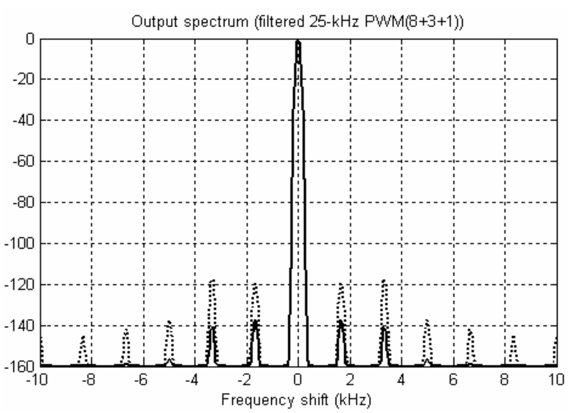

Fig. 12. Spectrum of 25 kHz 3-rd order PWM

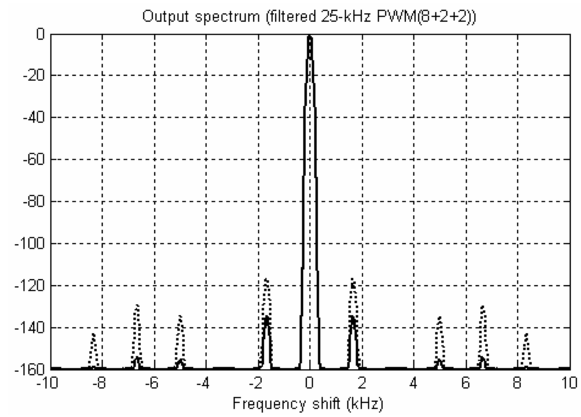

Fig. 13.Spectrum of 25 kHz 3-rd order PWM

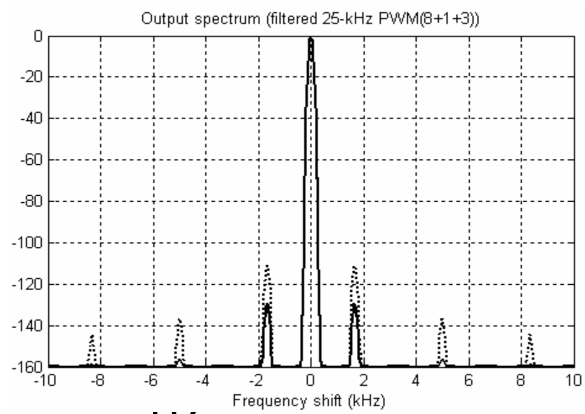

Fig. 14.Spectrum of 25 kHz 3-rd order PWM

For shaping of modulated PWM it is possible to use a principle of delta - modulation. The algorithm of PWM shaping by this principle is indicated at Fig. 15. The spectrum of output signal of the oscillator with DAC of such type is indicated at Fig. 16.

All graphic diagrams of modulated PWM spectrum were obtained for cases, when the amplitude of the first harmonics is maximal. From comparison of oscillator spectrums it is possible to make a conclusion, that the best outcomes (from the point of view of a collateral component level) are obtained at use a delta - modulated PWM. It is possible to explain so that at this condition oblong impulses of a PWM-sequence are distributed on time most uniformly.

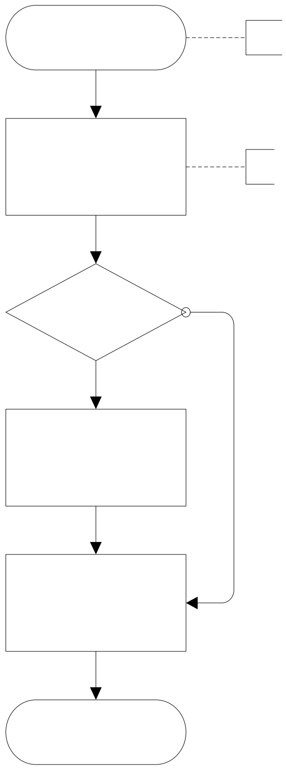

Fig. 15. Algorithm of delta-modulation formation

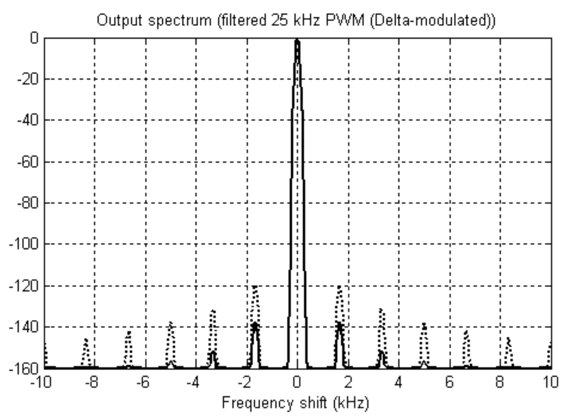

Fig. 16 Spectrum of 25 kHz delta modulations

2. ALGORITHMIC IMPROVEMENT OF BUILT-IN ADC RESOLUTION.

For deriving a small thermocompensation error is necessary a temperature measuring of high resolution. For temperature measurement it will be optimal to use a built in ADC. If working temperature range is ΔT, allowable thermocompensation error is δf and maximum slope of F-T curve is S, the ADC capacity may be determined as

$$n \geq \frac{1}{k \cdot l} \cdot \log_2 \left(\Delta T \cdot S / \delta f \right)$$

With accounting of nonlinearity of thermosensor curve and partial voltage range utilization, the ADC word length must not be less than 12.

For most popular microcontrollers the capacity of build in ADC does not exceed 10, and for C8051F310 it is equal to 8. This is quite unacceptable.

Frequently for increasing of ADC resolution, the dithering procedure is used. For successful application of this method the added noise must be compatible on amplitude with ADC quantum. The example of ADC resolution improvement from 12 to 16 bit for built-in thermosensor of Cygnal microcontroller is described in [8]. However this procedure is not acceptable for ADC improvement from 8 to 12 bit, because of nonoptimal in this case ratio of internal noise and ADC quantum.

The special technique of ADC equivalent capacity increasing by using the readings accumulation and averaging procedure is suggested in [9]. In this case the determined periodic signal with uniform voltage distribution is added to thermosensor signal. The amplitude of this signal must be equal or more that ADC quantum. The Fig. 17 illustrates the essence of this method.

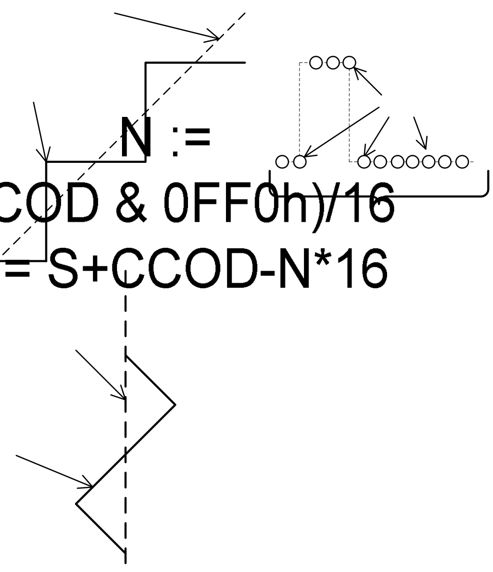

Fig. 17. The principle of ADC sygnal modulatiojn

For generation of such signal the 4-digit DAC on basis of a resistive matrix R-2R, connected to microcontroller pins, may be used, as it shown at Fig. 18.

The defect of this method is that the amplitude of a generated sawtooth signal will depend on thermoresistor resistance; in connection with this there will be an error, that nonmonotone dependent on temperature. Besides, the amount of circuit elements is rather great.

Other variant of generation of a sawtooth signal is the using of the first order RC-filter, on which input the meander signal with a period, much smaller that constant time of this filter (for example, how is shown at Fig.19), is applied.

The measurement error of temperature at such method depends on amplitude of a generated sawtooth signal. The graph of the normalized error (in units of ADC quantum) depending on the normalized amplitude (from peak to peak) in same units at amount of measurements repetitions of N = 256 is indicated at Fig.20. The dotted line at this figure corresponds to a level, at which one the initial ADC word length is increased by 4 bits.

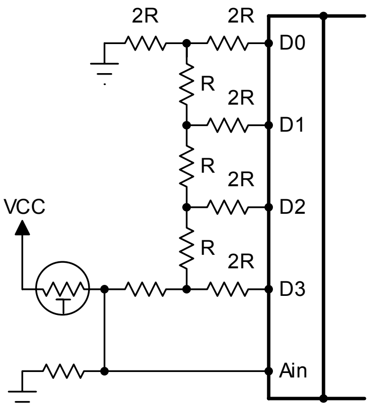

Fig. 18. Modulation by menace of R-2R matrix

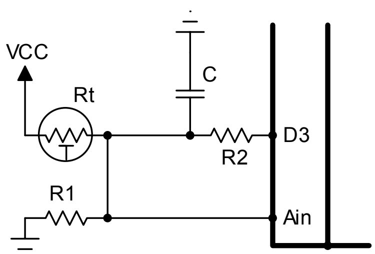

Fig. 19. Modulation by menace of RC-circuit

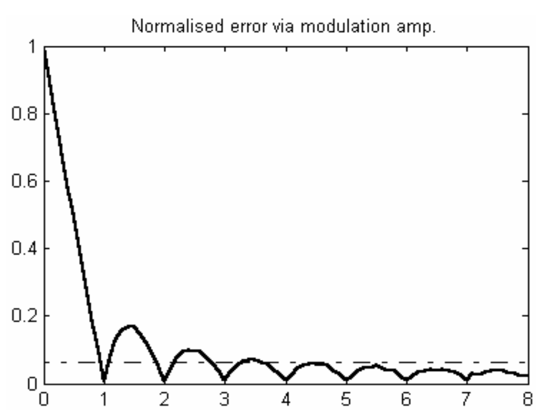

Fig. 20. Measurement error vs. modulation amplitude

As it is visible from this figure, the range of a modulation signal should not be less than 5 ADC quanta; in this case its differential nonlinearity will not render influence to exactitude of measurements. It is possible to show, that for circuit of Fig. 19 the magnitude of a modulation signal does not depend (practically) on thermoresistor resistance and is defined by the formula: $K=64*T/(R2\cdot C)$, where K is the range of a modulation signal in terms of ADC quanta, and T is the period of modulation signal. For convenience of microcontroller program implementations it is possible to select nominal of R1, R2 and C so that the frequency of ADC sampling was equal to frequency of PWM. Then the temperature measurement program can use the interruptions from PWM.

The outcomes of the given technique usage were checked for the temperature measurement circuit (Fig. 19) at R1=30k, R2=40k, C=820nF, Rt=150k (at -10 °C). The temperature coefficient of a thermoresistor is equal to 3%/°C. The frequency of sampling is 25 kHz; the number of sampling for a period of modulation is 256. At such parameters the range of a modulation signal is K=20.

The graph of a measurement temperature error for presence and absence of modulation signal are shown at Fig. 21 and Fig. 22 accordingly.

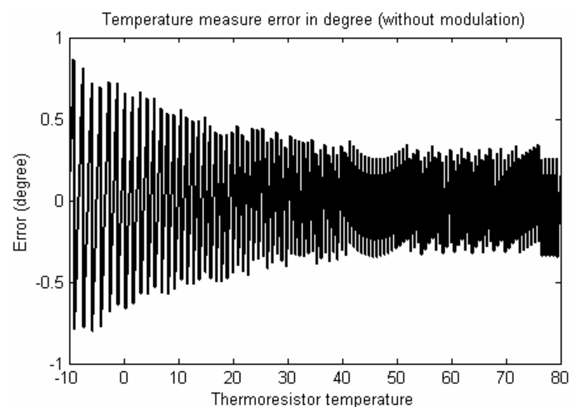

Fig. 22. Temperature measurement error without modulation

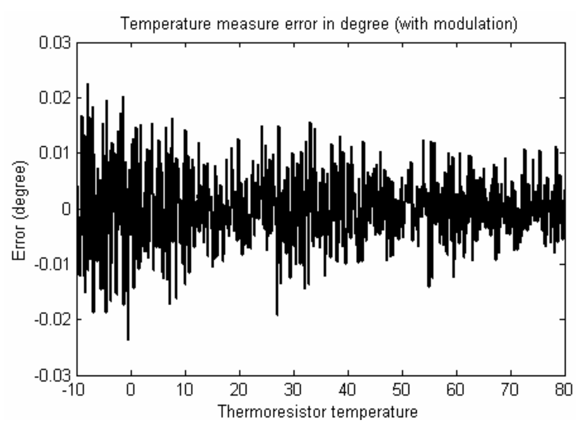

Fig. 23. Temperature measurement error with modulation

As it is visible from these figures, as a result of modulation introduction the measuring temperature error has decreased more than in 30 times. Thus the temperature measurement time will be equal to Tmeas = 256 / fpwm ≈ 10 mc, that is quite reasonable to a most cases. The algorithm of temperature measurement by means of a given technique at averaging amount N = 256 is shown at Fig. 24.

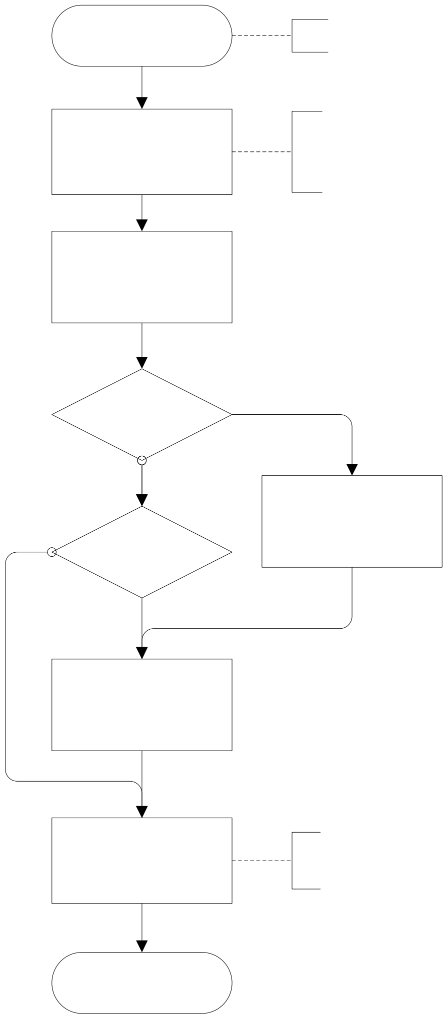

Fig. 24. Algorithm of temperature measuring

CONCLUSION

The materials, reviewed above, display, that the existing now microcontrollers can be successfully used for inexpensive high stable TCXO construction: they have small dimensions, low power consumption and good functional filling.

When tight requirements to a spectrum of a oscillator output signal are not produced, the absence of built-in DAC can be compensated at the expense of PWM module. Using of high-order PWM algorithms or delta - modulation algorithm allows to reduce a level of harmonics in a spectrum of a output signal up to a level - 140 dB even at use of filters with a high cut-off frequency. Thus a resolution capability of equivalent DAC achieves a required value (12 bit). By using of smoothing filters with lower cut-off frequency or at reduction of equivalent word length up to 10 (that is sufficient for many cases) the level of side component can be reduced up to - 150 … 160 dB, that corresponds to a noise pedestal level.

As the temperature processes are rather slow, to increase of analog to digital conversion accuracy the algorithms of measurements sequence averaging can be effectively used. Thus an equivalent analog to digital conversion exactitude can reach required magnitude without device cost increasing. Actually, it is quite enough to have a built-in 8-bit ADC for measurement of temperature with 0.030C accuracy.

One of the most perspective controllers from the point of view of use in DTCXO for today is C8051F310 of firm Cygnal. It is quite possible to expect, that some new applicable controllers will appear soon.

REFERENCES

[1]. A. Miayama, Y. Ikeda, S. Okano. A new digital temperature compensated crystal oscillator for a mobile telephone system. 1988 AFCS, pp 327 – 333.

[2]. D. Habic, D. Vasiljevic. Tempereture compensation of crystal oscillators using microcontroller – µCTCXO. 1994 IEE FCS. pp 587 – 593.

[3]. S. Deno, C Hahnlen, D. Landis. A low cost microcontroller compensated crystal oscillator. 1997 IEEE FCS. pp. 954 – 960.

[4]. S. Deno, C Hahnlen, D. Landis. A low cost high stability microcontroller compensated crystal oscillator. 1998 IEEE FCS. pp. 353 – 360.

[5]. W. Zhou, Y. Wang, L. Bai, H. Zhou, C. Liu, J. Li, J. Jia. A MCXO test system and its function in MCXO perfomances. 2001 IEEE FCS. pp 794 – 798.

[6]. A. Kosykh, A. Terentjiev. A comparative estimation of low-cost DTCXO structures. Proceedings of International symposium on acoustoelectronics, frequency control and signal generation. 1998, Moskow – St. Petersburg. pp. 63-67.

[7]. A. Kosykh, B. Ionov. Dynamic temperature model and dynamic temperature compensation of crystal oscillators. IEEE Transactions on Ultrasonics, Ferroelectrics and Frequency Control. p. 370-374. May, 1994

[8]. Improving ADC resolution by oversampling and averaging. CYGNAL Applcation note AN018

[9]. Patent № 1084938 (USSR), H 03 B 5/32 Bulletin № 13, 1984. /Thremocompensated Crystal Oscillator. V. Bagaev, A. Kosykh, A. Lepetaev.