A Simple Current-Mode Quadrature Oscillator Using Single CDTA

Winai JAIKLA¹, Montree SIRIPRUCHYANUN², Josef BAJER³, Dalibor BIOLEK³,⁴

¹ Dept. of Electric and Electronic, Faculty of Industrial Technology, Suan Sunandha Rajabhat University, Dusit, Bangkok, 10300, Thailand
² Dept. of Teacher Training in Electrical Engineering, Faculty of Technical Education, King Mongkut’s University of Technology North Bangkok, Bangsue, Bangkok, 10800, Thailand
³ Dept. of Electrical Engineering, Faculty of Military Technology, University of Defense Brno, Kounicova 65, 612 00 Brno, Czech Republic
⁴ Inst. of Microelectronics, Faculty of Electrical Engineering and Communications, Brno University of Technology, Udalni 53, 602 00 Brno, Czech Republic

Abstract. This article presents a simple current-mode quadrature oscillator using a single Current Differencing Transconductance Amplifier (CDTA) as the active element. The oscillation condition and oscillation frequency can be electronically controlled. The circuit structure is very simple, consisting of merely one CDTA, one resistor and two capacitors. The proposed circuit is suitable for IC architecture. The PSpice simulation and experimental results are shown, and the results agree well with the theoretical assumptions.

Keywords
Oscillator, CDTA, current-mode.

1. Introduction

Oscillator is an important basic building block, which is frequently employed in electrical engineering applications. Among several kinds of oscillator, the quadrature oscillator is widely used because it can offer sinusoidal signals with 90° phase difference, for example, in telecommunications for quadrature mixers and single-sideband modulators [1]. Presently, the current-mode technique has been more popular than the voltage-mode technique. This is due to the requirements in low-voltage environments such as portable and battery-powered equipment. The current–mode technique is ideally suited for this purpose. Today, there is a growing interest in synthesizing current-mode circuits because of their potential advantages such as larger dynamic range, higher signal bandwidth, greater linearity, simpler circuitry, and lower power consumption [2].

A reported 5-terminal active element, namely Current Differencing Transconductance Amplifier (CDTA) [3], seems to be a versatile component in the realization of a class of analog signal processing circuits, particularly analog frequency filters [3], [4]. It is actually a current-mode element whose input and output signals are currents. In addition, the output current of the CDTA can be electronically adjusted. Besides, a modified version of CDTA, whose parasitic resistances at two current input ports can be electronically controlled, has been proposed in [5]. This circuit is called Current Controlled Current Differencing Transconductance Amplifier (CCCDTA). Another CDTA modification, called ZC-CDTA (Z Copy CDTA) is proposed in [6], providing a copy of the current flowing to the \( z \) terminal. This copy can be used as an output signal for driving an independent load.

Several implementations of oscillator employing CDTAs or CCCDTAs have been reported in the literature [7-13]. Unfortunately, these circuits suffer from one or more of the following weaknesses: They use more than one CDTA or CCCDTA and an excessive number of passive elements, which is not convenient for IC fabrication. In addition, some reported circuits use a multiple-output CDTA or CCCDTA. Consequently, the circuits become more complicated.

The purpose of this paper is to introduce a current-mode quadrature oscillator, based on a single CDTA. The oscillation condition and oscillation frequency can be adjusted electronically. The circuit construction consists of one CDTA, one resistor, and two capacitors. Finally, this oscillator has been built by means of a CDTA chip which has been fabricated in the CMOS technology [14]. The PSpice simulation and the experimental results correspond to the theoretical analyses.
2. Principle of Operation

2.1 Current Differencing Transconductance Amplifier (CDTA)

Since the proposed circuit is based on the CDTA, a brief review of CDTA is given in this Section. The characteristics of the ideal CDTA are represented by the following hybrid matrix:

\[
\begin{bmatrix}
V_p \\
V_n \\
I_x \\
I_z
\end{bmatrix} = \begin{bmatrix}
0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 \\
1 & -1 & 0 & 0 \\
0 & 0 & 0 & g_m
\end{bmatrix} \begin{bmatrix}
I_p \\
I_n \\
V_x \\
V_z
\end{bmatrix}.
\]  

(1)

In general, CDTA can contain an arbitrary number of \(x\) terminals, providing currents \(I_x\) of both directions. As an example, the symbol and the equivalent circuit of the CDTA with a pair of \(x^+\) and \(x^-\) terminals are illustrated in Fig. 1(a) and (b), respectively.

![CDTA schematic symbol and equivalent circuit](image)

Fig. 1. CDTA (a) schematic symbol, (b) equivalent circuit.

2.2 Proposed Circuit

The proposed quadrature oscillator is designed by cascading a first-order all-pass filter and a non-inverting lossless integrator as shown in Fig. 2. Based on this block diagram, the single-CDTA quadrature oscillator can be implemented according to Fig. 3. In order to utilize the current through the capacitor \(C_2\), an auxiliary \(zc\) terminal is used. The internal current mirror provides a copy of the current flowing out of the \(z\) terminal to the \(zc\) terminal [6].

The characteristic equation of the proposed oscillator in Fig. 3 can be expressed as follows:

\[
s^2C_1C_2R + g_m = 0.
\]  

(5)

From (5), the oscillation frequency is as follows:

\[
\omega_{osc} = \frac{g_m}{\sqrt{C_1C_2R}}.
\]  

(6)

From the circuit in Fig. 3, the current transfer function from \(I_{o1}\) to \(I_{o2}\) is

\[
\frac{I_{o2}(s)}{I_{o1}(s)} = \frac{g_m}{sC_2}.
\]  

(7)

For sinusoidal steady state, equation (7) becomes

\[
\frac{I_{o2}(j\omega)}{I_{o1}(j\omega)} = \frac{g_m}{sC_2} e^{-j90^\circ}.
\]  

(8)

The phase difference \(\phi\) between \(I_{o1}\) and \(I_{o2}\) is

\[
\phi = -90^\circ,
\]  

(9)

ensuring that the currents \(I_{o2}\) and \(I_{o1}\) are in quadrature.

For the oscillation frequency, with regard to (6), equation (8) gives

\[
\frac{I_{o2}(j\omega_{osc})}{I_{o1}(j\omega_{osc})} = \frac{g_m}{sC_2} e^{-j90^\circ}.
\]  

(10)

Taking into account oscillation condition (3), one can conclude that the oscillator will provide quadrature signals of equal magnitudes.

All the active and passive sensitivities of the oscillator are low as shown in (11):

\[
S_{g_m}^{m_m} = -\frac{1}{2}, S_{R}^{m_m} = \frac{1}{2}.
\]  

(11)
3. Experimental Results

To prove the performance of the proposed circuit, the oscillator was constructed using the CDTA chip. The CDTA was designed and manufactured in the 0.7μm CMOS technology [14]. Its basic small-signal parameters are as follows:

- Input resistances of p and n terminals $R_p = 260 \ \Omega$, $R_n = 250 \ \Omega$.
- Resistance of z terminal $R_z = 3.48 \ \text{M}\Omega$.
- Resistance of $x^+$, $x^-$ terminals $R_x = 2.2 \ \text{M}\Omega$.
- Current gains from p to z terminal and from n to z terminal are 1 and -1, respectively, with a 3-dB cutoff frequency of 10 MHz.
- Transconductance of the OTA amplifier is 1.22 mA/V with a 3-dB cutoff frequency of 16 MHz.

The CDTA chip was manufactured with the fixed transconductance, with $x^+$ and $x^-$ terminals, and without the auxiliary $zc$ terminal. During the experiment, the current through the $x$ terminal was sensed via a $I/V$ converter, formed by an operational amplifier OPA 2650 with 1 kΩ feedback resistance. That is why the output voltage in millivolts corresponded to the input current in microamperes.

The quadrature oscillator was designed for an oscillation frequency of about 1MHz. The values of the external elements were designed as follows:

$$R = 1/g_m = 819 \ \Omega, \ C_1 = C_2 = 150 \ \text{pF}. \hspace{1cm} (12)$$

According to (6), the theoretical value of the oscillation frequency is 1.294 MHz.

Actually, the measured values (12), as used in the experiment, were

$$R = 829 \ \Omega, \ C_1 = 153 \ \text{pF}, \ C_2 = 156 \ \text{pF}. \hspace{1cm} (13)$$

Then the corresponding theoretical value of the oscillation frequency is 1.257 MHz.

After substituting the values from (13) to (3), we can see that the left-side value is greater than the right-side one. In other words, the current loop-gain is greater than unity, which is the condition of self-starting the oscillations.

The waveform measured, namely the output of the $I/V$ converter measuring the current $I_o$ in Fig. 3, is shown in Fig. 4. The magnitude is approximately 75 μA. The measured frequency of 962 kHz is less than the theoretical value 1.257 MHz. In the following Section, an analysis of the real effects leads to an explanation.

Fig. 4. Measured output waveform in millivolts, corresponding to the current $I_o$ in microamperes.

Fig. 5. Fourier analysis of generated waveform.

4. Analysis of Real Effects

The circuit operation is affected by linear and nonlinear influences. In this Section, an analysis of real effects is done in the following structure:

a) Analysis of linear effects. The oscillation frequency and the condition of oscillation are affected by parasitic impedances of CDTA terminals, by DC current transfers from p to z and from n to z terminals, which are not generally equal to one in magnitude, and by their frequency dependence. Also, the frequency characteristic of the transconductance may have an effect.

The analysis of the above linear model of the oscillator can be performed in fully symbolic form in order to obtain formulas for the oscillation condition and for the frequency of oscillation. However, more complex high-order models do not provide such results in a closed form. That is why the symbolic analysis will be evaluated only for several main influences, and the remaining one will be analyzed separately via computer simulation.

Fig. 5 shows the measured spectrum components of the generated waveform. The magnitude of the second harmonics is more than 56 dB below the magnitude of the fundamental harmonics. The corresponding THD is 0.16%. Increasing the value of $R$ results in increasing the loop-gain and thus increasing the magnitude and also the distortion of generated signals.
b) Analysis of nonlinear effects. There are two types of important nonlinearities of manufactured CDTA: nonlinear curves $V_p = f(I_p)$ and $V_n = f(I_n)$, and nonlinear OTA characteristic $I_m = f(V_z)$. These nonlinearities are responsible for properly starting up the oscillations, for automatically achieving the oscillation condition in the steady state, and they also affect the value of the oscillating frequency.

A hand-and-paper analysis of such nonlinear effects is practically impossible. Therefore, all these characteristics were measured on the chip and transferred to PSpice as nonlinear behavioral models, built via controlled sources and the look-up table method. These models were then completed with several small-signal parameters which PSpice cannot extract from the above nonlinear curves. Such a final model has been tested and its behavior has been compared with real experiments. The coincidence was excellent. This PSpice model of the CDTA chip helped us with the analysis of nonlinear effects in the quadrature oscillator. The simulations based on this model also provide waveforms which are not easily accessible via practical measurements.

4.1 Linear Effects

The most important small-signal real influences which affect the circuit operation are modeled in Fig. 6.

![Diagram](image)

**Fig. 6.** Modeling linear effects in the oscillator.

In addition to the working passive elements $R$, $C_1$, and $C_2$ (in thick lines), several parasitic impedances show up here. The input Current Differencing Unit (CDU) is now described by the equation

$$I_z = \alpha_p I_p - \alpha_n I_n,$$

(14)

where the parameters $\alpha_p$ and $\alpha_n$ are current transfer values, deviating from one, depending on the internal circuit construction. $R_{Lz}$ and $R_{Lx}$ are loading impedances for sensing the output currents $I_{o1}$ and $I_{o2}$. In conjunction with the $R_{z}$, $C_z$, or $R_{z}$, $C_z$, pair they form a frequency-dependent current divider of terminal current $I_o$ or $I_n$. Making the influence of these parasitic impedances negligible, one should choose the load resistances as low as possible. Throughout the following test, we follow this assumption. Then the remaining non-ideal factors will be as follows:

- General values of $\alpha_p$ and $\alpha_n$. We will take them into account for the symbolic analysis in order to evaluate this influence. However, for the manufactured CDTA chip, these values are of unity-value with sufficiently high precision.
- Nonzero $R_p$ and $R_n$ resistances. $R_n$ is in series to $R$ and thus it does not change the character of this working impedance. The $R_n$ influence can be compensated by decreasing the $R$ value. This is not the case of $R_p$, which acts in series to $C_1$.
- Parasitic impedance of the $z$ terminal. $C_z$ can be considered a part of $C_2$. However, $R_z$ changes the type of the impedance, which should normally be of purely capacitive character.
- Parasitic impedance of the $x+$ terminal. The $x+$ terminal is connected to low-impedance $p$ and $n$ terminals through $C_1$ and $R$ elements. For low frequencies, the corresponding impedance level is given by $R'$ (kilo-Ohms), which is much smaller than $R_{x+}$. For the high-frequency region, this impedance level is determined by $R_n$ (hundreds of Ohms), but $C_n$ can cause a leakage of the $I_{x+}$ current to the ground. This capacitive part of parasitic impedance increases the order of the oscillator model from the original value 2 to 3. This fact considerably complicates the evaluation of the symbolic analysis. That is why we will consider this part of the model in the second stage of the analysis.

- Frequency dependence of $\alpha_p$ and $\alpha_n$ transfer coefficients and of OTA $g_m$. According to the information in Section 3, the corresponding cutoff frequencies are 10 MHz and 16 MHz. For the oscillation frequencies near these values, the effect can be considerable. However, the order of circuit model is also increased. That is why we will analyze these factors separately, after the symbolic analysis.

Considering the above factors and using the notation

$$R' = R + R_n, \quad C'_2 = C_2 + C_n, \quad G_z = 1/R_{x+}, \quad G_x = 1/R_x,$$

the symbolic analysis of the model in Fig. 6 leads to the following equations:

The modified oscillation frequency

$$\omega_{osc} = \sqrt{\frac{\alpha_p g_m + (1 + R'G_z)G_x}{C_1C_2(R' + R_p + R'R_pG_z)}}.$$  

The modified condition of oscillation

$$g_m(R' \alpha_p - R_p \alpha_n) = \frac{C_1}{C_2}(1 + R'G_x) + (R' + R_p + R'R_pG_z)G_x.$$  

(16)
Setting the numerical values of the parameters from Section 3 and a short re-arrangement lead to the conclusion that several terms in (15) and (16) can be neglected. The simplified versions of (15) and (16) are as follows:

\[ \omega_{osc} = \frac{\alpha_n g_m}{\sqrt{C_1 C_2 (R + R_p)}} \Rightarrow \omega_{osc} = \frac{\alpha_n g_m}{\sqrt{C_1 C_2 (R + R_p)}} \]

\[ \omega_{osc} \approx \frac{\alpha_n g_m}{\sqrt{C_1 C_2 (R + R_p)}} \]

\[ a = \sqrt{\frac{R_p + R_n}{R}} \]

Equation (17) shows that the oscillation frequency decreases below its theoretical value due to nonzero parasitic resistances \( R_p \) and \( R_n \). For \( R_p = 260 \Omega \) and \( R_n = 250 \Omega \), and \( R = 829 \Omega \), the frequency should decrease 1.279 times, i.e. from the theoretical value 1.257 MHz to 1.072 MHz. However, the real experiments yielded the value 962 kHz (see Section 3). Obviously, there are some additional influences that decrease the oscillation frequency. For example, parasitic capacitance \( C_{para} \) decreases the frequency to 980 kHz, and the cutoff frequencies of \( a \), \( \alpha_p \), and \( g_m \) decrease it by another 15 kHz. We can conclude that such additional modifications of oscillation frequency are approximately by one order less relevant than the changes described by (17).

Equation (18) demonstrates that for the identical values of parasitic resistances \( R_p \) and \( R_n \) and for \( \alpha_p = \alpha_n = 1 \), the oscillation condition is practically not affected by the above effects.

Another analysis of the model in Fig. 6 under the above simplifications leads to the conclusion that the ratio of current \( I_{z2} \) and current \( I_{z1} \) is still described by (8), thus their phase shift is 90 degrees. However, their magnitudes will not be identical any more. Utilizing (8), (17) and (18) and a short re-arrangement yield the following formula:

\[ \frac{I_{z2}(j\omega_{osc})}{I_{z1}(j\omega_{osc})} \approx \frac{R' + R_p}{\alpha_p (R' \alpha_p - R_p \alpha_n) e^{-j90^\circ}}. \]

For ideal values of \( \alpha_p = \alpha_n = 1 \), the magnitude ratio \( b = \frac{\sqrt{R + R_p + R_n}}{\sqrt{R + R_n - R_p}} \) is as follows:

\[ b = \frac{\sqrt{R + R_p + R_n}}{\sqrt{R + R_n - R_p}} \approx 1.279. \]

The above results can be summarized in a practical formula: If the oscillation frequency decreases \( a \) times due to parasitic linear effects, where \( a \) is given by (17), then the magnitudes of generated quadrature signals will no longer be equal but their ratio will be \( b \), where \( b \) is given by (19) or (20). This formula will be confirmed in the following Subsection via PSpice simulation.

### 4.2 Nonlinear Effects

Nonlinear effects were studied by means of PSpice simulation, utilizing nonlinear models of the CDTA chip, briefly described in Section 4. Two CDTA models were used subsequently, first model No. 1 and then model No. 2.

**Model No. 1:** A simplified model of the CDU consists of linear \( R_p \) and \( R_n \) resistances, controlled source for modeling the difference of input currents, the \( z \)-terminal parasitic resistance, and single-pole frequency response with 3dB cutoff frequency. For the internal OTA, a full nonlinear model of the \( I_x = f(I_z) \) from Fig. 7, resulting from the data measured [16], was used. In addition, the output resistances of \( x \)-terminals and the 3dB cutoff frequency were also modeled.
Model No. 2: A full nonlinear model of the CDU is implemented here, utilizing the measured nonlinear characteristics \( V_p = f_p(I_p) \) and \( V_n = f_n(I_n) \) [16], [17], see Fig. 8. The other parts of the model are identical with model No. 1.

The results of the steady-state analysis by means of model No. 1 are shown in Fig. 9. Starting up the oscillations was attained for \( R = 864 \Omega \), which is in good agreement with real measurements (829 \( \Omega \)). For such a resistance value, the theoretical oscillation frequency is 1.224 MHz. The simulated value is 960.3 kHz, i.e. 1.275 times smaller. It is obvious that the generated waveforms are not of equal magnitudes. Their ratio was measured as 1.128, which roughly corresponds to the ratio 1.173, computed from (20).

As shown in Fig. 9, the peak value of the voltage at \( z \) terminal is below 200 mV. It is obvious from Fig. 7 that the transconductance amplifier works in nearly linear regime for such a voltage swing. Its equivalent transconductance is decreasing when increasing the voltage swing. The equilibrium, i.e. fulfilling the oscillation condition, is established for a concrete voltage swing. That is why the automated adjustment of the oscillation condition is due to the nonlinearity of the OTA stage.

Fig. 9 illustrates an interesting phenomenon: in spite of different magnitudes of output currents \( I_o \) and \( I_o \), the magnitudes of currents \( I_p \) and \( I_n \) are equal. It can be easily proved that \( I_p \) and \( I_n \) are also orthogonal. Their magnitudes are approximately 150 \( \mu A \).

The simulation results for CDTA model No. 2 is given in Fig. 10. The parameters of the passive elements are the same as for the previous simulation. The oscillation frequency is now 964 kHz, which is in very good agreement with the real measurement (962 kHz). Also, the magnitude of \( I_o \) (73 \( \mu A \)) now corresponds to the reality. Note that the signal levels are now lower than without detailed modeling of the CDU. This is due to the nonlinearity of \( p \) and \( n \) inputs. The \( R_p \) and \( R_n \) equivalent resistances are now dependent on signal levels, and the nonlinearity of CDU provides the automatic adjustment of the oscillation condition. It should be also noted that a DC offset appears in the waveforms, particularly in \( I_o \), which is caused by the offset of DC characteristics of the CDU (see Fig. 8).

4.3 Elimination of Real Effects

It is obvious from (17) and (20) that both real effects, i.e. decreasing the oscillation frequency and different magnitudes of generated quadrature signals, can be eliminated via increasing the working resistance \( R \) to be at least by one order higher than the parasitic resistances \( R_p \) and \( R_n \). Let us suppose that \( R \) will be increased \( k \) times. In order to preserve the designed value of the oscillation frequency (6) and also to fulfill the condition of oscillation (3), several possible operations can accompany the \( R \) increase. Two relevant operations are summarized below:

- decreasing \( C_1 \) \( k \) times, or
- increasing \( g_m \) \( k \) times, increasing \( C_2 \) \( k \) times, decreasing \( C_1 \) \( k \) times.

The second method cannot be applied to the oscillator with CDTA chip, which has a fixed transconductance. Let us increase \( R \) ten times, to 8.64 k\( \Omega \) and decrease \( C_1 \) ten times, to 15.3 pF. The results of PSpice simulation for model No. 2 is in Fig. 11. Note that the oscillation frequency is now near the theoretical value, namely 1.171 MHz, and the quadrature signals are of almost equal magnitudes.
5. Conclusion

A simple quadrature oscillator based on one Current Differencing Transconductance Amplifier (CDTA), two capacitors and one resistor is described. The proposed oscillator, designed for the megahertz range, was constructed using a CDTA chip, which was fabricated in the CMOS technology. The measurements confirmed the functionality of the principle. Two problems were discussed: 1) The oscillation frequency is lower than the theoretical value. 2) The quadrature output currents are of different magnitudes although the oscillator was designed for equal magnitudes. Theoretical analysis shows the common reason for both problems: the finite values of $R_p$ and $R_n$ resistances of the CDTA. A quantitative description of these phenomena is given in equations (17) and (20). Based on this knowledge, a simple procedure is designed how to overcome the above difficulties.

For a practical implementation of the proposed topology of quadrature oscillator, it is useful if the CDTA has at least two x-terminals and one so-called zc terminal, which provides a copy of the z-terminal current. For an easy elimination of the above real influences, the current differencing unit of the CDTA should be designed with the $R_p$ and $R_n$ parasitic resistances as low as possible. The proposed value is at least one order lower than the value of the resistance of the working resistor connected in series to the n terminal.

Acknowledgements

The research described in the paper was supported by the Czech Grant Agency under grant No. 102/08/0784, and by the research programmes of BUT No. MSM0021630503 and UD Brno No. MO FVT0000403.

References


About Authors...

Winai JAIKLA received the B. Tech. Ed. degree in Telecommunication Engineering from King Mongkut’s Institute of Technology Ladkrabang, Thailand, in 2002 and M. Tech. Ed. in Electrical Technology from King Mongkut’s Institute of Technology, North Bangkok (KMITNB), in 2004. He has been with the Electric and Electronic Program, Faculty of Industrial Technology, Suan Sunandha Rajabhat University, Bangkok, Thailand since 2004. His research interests include electronic communications, analog signal processing and analog integrated circuits. He is a member of ECTI (Thailand).

Montree SIRIPRUCHYANUN received the B. Tech. Ed. degree in Electrical Engineering from King Mongkut’s Institute of Technology, North Bangkok (KMITNB), the M.Eng. and D. Eng. degrees in Electrical Engineering from King Mongkut’s Institute of Technology Ladkrabang (KMITL), Bangkok, Thailand, in 1994, 2000 and 2004, respectively. He has been with the Faculty of Technical Education, King Mongkut’s University of Technology, North Bangkok (KMUTNB), since 1994. Presently, he is with the Department of Teacher Training in Electrical Engineering as Associate Professor, KMUTNB. His research interests include analog-digital communications, analog signal processing, and analog integrated circuits. He is a member of IEEE (USA), IEICE (Japan), and ECTI (Thailand).

Josef BAJER was born in 1982. In 2005 and 2008, he received the B.Sc. degree in Electrical and Special Aircraft Equipments and the M.Sc. degree in Avionic Systems from the University of Defense Brno (UDB), Czech Republic. He is currently working towards the Ph.D. degree in Electronic and Weapons Systems at the Faculty of Military Technologies UDB. His interests include analog and digital signal processing and applications of modern active elements working in current and hybrid modes.

Dalibor BIOLEK received the M.Sc. degree in Electrical Engineering from the Brno University of Technology, Czech Republic, in 1983, and the Ph.D. degree in Electronics from the Military Academy Brno, Czech Republic, in 1989. He is currently with the Department of EE, University of Defense Brno (UDB), and with the Department of Microelectronics, Brno University of Technology (BUT), Czech Republic. His scientific activity is directed to the areas of general circuit theory, frequency filters, and computer simulation of electronic systems. For years, he has been engaged in algorithms of the symbolic and numerical computer analysis of electronic circuits with a view to the linear continuous-time and switched filters. He has published over 250 papers and is the author of a book on circuit analysis and simulation. At present, he is Professor at BUT and UDB in the field of Theoretical Electrical Engineering. Prof. Biolek is a member of the CAS/COM Czech National Group of IEEE. He is also the President of Commission C of the URSl National Committee for the Czech Republic.