

The invention consists in a method for flow-rate measurement by means of ultrasonic waves which are propagated in a medium flowing through a measuring tube between two ultrasonic transducers which are spaced a defined distance apart and one of which operates as a transmitting transducer which converts an electric transmission signal into an ultrasonic wave of the same frequency, while the other operates as a receiving transducer which converts the incoming ultrasonic wave into an electric reception signal of the same frequency, the flow rate being determined by measuring the phase shift between the transmission signal and the reception signal in such a way that in measurement time intervals whose duration corresponds to the phase shift, those periods of a signal of a counting frequency are counted which are substantially greater than the measuring frequency corresponding to the frequency of the transmission signal, that in a measurement cycle comprising a plurality of successive measurement time intervals, the phase angle of the start phase which the counting-frequency signal has at the beginning of each counting operation with respect to the measurement time interval is increased from measurement time interval to measurement time interval, and that the counts obtained in the course of the measurement cycle are evaluated to form a measured value with increased measurement resolution.

In this flow rate measuring method, the flow rate to be measured is derived from the phase shift between the transmission signal and the reception signal in an ultra-

sonic wave propagated in one direction if the velocity of sound is known. By determining the phase shifts in an ultrasonic wave propagated in the flow direction and an ultrasonic wave propagated against the flow direction, it is possible to obtain from the difference between said phase shifts a measurement result for the flow rate which is independent of the velocity of sound.

The invention further consists in an arrangement comprising a measuring tube through which a medium flows, two ultrasonic transducers on the measuring tube which are spaced a defined distance apart, devices for applying an electric transmission signal with the measuring frequency to one of the two ultrasonic transducers and for receiving the electric reception signal with the measuring frequency from the other ultrasonic transducer, a counting-frequency generator which generates a counting-frequency signal with a counting frequency which is substantially greater than the measuring frequency, a counter which counts the periods of the counting-frequency signal in measurement time intervals corresponding to the phase shift between the transmission signal and the reception signal, an evaluating circuit to which the count reached in each measurement time interval is transferred, and devices whereby in a measurement cycle comprising a plurality of successive measurement time intervals, the phase angle of the start phase which the counting-frequency signal has at the beginning of each counting operation with respect to the measurement time interval is increased from measurement time interval to measurement time interval.

In a method disclosed in CH-PS 628 140, the phase shift between the transmission signal and the reception signal determined for each propagation direction is converted to a DC voltage which is stored in a sample and hold circuit. The two stored DC voltages are then applied to the two inputs of a differential amplifier which thus provides at the output a differential voltage which is proportional to the difference between the phase shifts and thus to the flow rate. In addition, by forming the sum of the stored DC voltages, a voltage is obtained which is proportional to the velocity of sound in the medium. Finally, in this known method, the frequency of the ultrasonic wave is regulated at an optimum value in dependence upon the phase shifts, and the flow rate obtained from the differential voltage is corrected in dependence upon the set frequency.

This known method requires a large amount of precise analog circuitry but is still liable to the inevitable causes of errors in any analog signal processing scheme, so that the measurement accuracy obtainable is limited. Furthermore, the output signal obtained is available in the form of an analog voltage which is not directly suitable for digital evaluation, for example by a micro-computer.

The problem underlying the invention is to provide a flow-rate measurement method of the above kind which permits high-accuracy flow-rate measurements with a relatively small amount of circuitry and provides the measurement result in the form of a digital signal.

In the method according to the invention, by counting the periods of a counting frequency during the measurement time intervals corresponding to the phase shifts, the phase measurement is reduced to a digital time measurement which can conveniently and reliably be carried out with the widely used digital integrated circuits. However, there is the problem that the digital measurement has a system-inherent quantizing error which limits the resolution obtainable in the measurement because the smallest distinguishable time interval is equal to the period of the counting-frequency signal. Under normal conditions of use, the measurement time intervals corresponding to the phase measurement are very short, and they must be measured with high accuracy and correspondingly high resolution particularly when the difference of the phase shifts is evaluated for a flow-rate measurement independent of the velocity of sound. The duration of the measurement time intervals is of the order of microseconds and their difference is of the order of fractions of one nanosecond. To measure such short time intervals by digitally counting counting-frequency periods with a resolution sufficient to achieve the required accuracy, extremely high counting frequencies would be necessary and their generation and evaluation would involve a great amount of technical complexity.

In the method according to the invention, a resolution improved with respect to the counting frequency is obtained by additionally increasing the start phase in successive measurement time intervals. As a result of this measure, the number of counting-frequency periods counted in the course of a measurement time interval

varies during the measurement cycle; while k counting-frequency periods are counted for a number of measurement time intervals, $k+1$ counting-frequency periods are counted for the remainder of the measurement time intervals. The ratio of these numbers provides more accurate information on the actual duration of the measurement time intervals and consequently permits an increase in resolution beyond the quantizing step corresponding to one counting-frequency period. This increase in resolution increases with the number of measurement time intervals evaluated in a measurement cycle to form a measured value, but the measurement period required to obtain a measured value increases to the same extent. In general, however, the necessary measurement periods are so short that their duration is of no significance compared with the advantage of the increased resolution.

According to a book by P. Horowitz and W. Hill, "The Art of Electronics", New York, 1980, pages 622 and 623, it is also known from the general prior art of electronic frequency- and time-measurement technology to reduce the above-explained inaccuracy error by causing the counting-frequency signal to jitter, i.e., by allowing its frequency to slightly vary about a center frequency in a manner comparable to frequency modulation.

In the invention, however, the frequency of the counting-frequency signal is maintained constant, but the start phase of this signal is increased continuously.

Preferred embodiments and further developments of the method according to the invention and the arrangement

for carrying out the method are characterized in the subclaims. These embodiments and further developments relate in particular also to measures for determining the number of whole wavelengths on the measuring distance and to a construction of the ultrasonic transducers which is particularly suited to carrying out the method according to the invention.

Further features and advantages of the invention will become apparent from the following description of embodiments taken in conjunction with the accompanying drawings, wherein:

Fig. 1 is a schematic illustration of a flow rate measuring arrangement in which the invention is applicable;

Fig. 2 is a timing diagram of signals which occur in the flow rate measuring arrangement of Fig. 1 in one mode of operation;

Fig. 3 is a timing diagram of the same signals in another mode of operation of the flow rate measuring arrangement of Fig. 1;

Fig. 4 is a timing diagram serving to explain the digital phase measurement performed in the flow rate measuring arrangement of Fig. 1;

Fig. 5 is a block diagram of one embodiment of the flow rate measuring arrangement according to the invention;

Fig. 6 shows diagrams to explain the operation of the flow rate measuring arrangement of Fig. 5;

Fig. 7 is a more detailed block diagram of one embodiment of the controllable phase shifter contained in the flow rate measuring arrangement of Fig. 5;

Fig. 8 shows a modified embodiment of the flow rate measuring arrangement of Fig. 5;

Fig. 9 is a block diagram of another embodiment of the flow rate measuring arrangement according to the invention;

Fig. 10 is a circuit diagram of one embodiment of the difference-counting-frequency generator contained in the flow rate measuring arrangement of Fig. 9;

Fig. 11 is a block diagram of another embodiment of the flow rate measuring arrangement;

Fig. 12 shows diagrams to explain the operation of the flow rate measuring arrangement of Fig. 11;

Fig. 13 is a sectional view of a preferred embodiment of an ultrasonic transducer;

Fig. 14 is an overall view of the flow rate measuring arrangement with a measuring tube and two ultrasonic transducers and two attenuator tubes of the type shown in Fig. 13;

Fig. 15 is a block diagram of an additional arrangement for determining the number of whole wavelengths in the measuring tube;

Fig. 16 shows diagrams to explain the operation of the arrangement of Fig. 15;

Fig. 17 is an overall block diagram of the flow rate measuring arrangement, and

Fig. 18 shows diagrams to explain the operation of the inverting circuit contained in the flow rate measuring arrangement of Fig. 17.

The flow rate measuring arrangement illustrated in Fig. 1 includes a measuring tube 10 through which a medium flows in the direction of the arrow with the flow velocity V_M . To determine the volume flow rate of the medium, it suffices to measure the flow velocity V_M , since the volume flow rate is equal to the product of the flow velocity V_M and the known cross-sectional area of the measuring tube.

1 To measure the flow velocity V_M , two ultrasonic transducers 12
and 14 are mounted on the measuring tube 10 which are separated
by a precisely known distance L_M . Each ultrasonic transducer
can be operated either as a transmitting transducer or
5 as a receiving transducer. In position 1, which is shown
in the drawing, a directional changeover switch 16 connects
the ultrasonic transducer 12 to the output of a transmission
frequency generator 18 and at the same time the ultrasonic
transducer 14 to the input of a reception signal processing
10 circuit 20. In its other position, 2, the directional change-
over switch 16 connects the ultrasonic transducer 14 to the
output of the transmission frequency generator 18 and the
ultrasonic transducer 12 to the input of the reception signal
processing circuit 20. The directional changeover switch 16
15 is illustrated in the drawings symbolically by mechanical
changeover contacts, but in reality it is constructed in a
manner known per se with electronic switching elements.

The transmission frequency generator 18 furnishes at the
20 output an electrical transmission AC voltage U_S of frequency
 f_M which, in the illustrated position of the directional
changeover switch 16, is applied to the ultrasonic trans-
ducer 12. The ultrasonic transducer 12 then operates as
transmitting transducer. It is so constructed that due to
25 its excitation by the electrical transmission AC voltage U_S
it generates an ultrasonic wave of the same frequency f_M
which is propagated in the direction of the axis of the
measuring tube 10 in the medium flowing through the meas-
uring tube. Said ultrasonic wave passes in the flow direc-
30 tion to the ultrasonic transducer 14 which, in the illustrated
position 1 of the directional changeover switch 16, is
connected to the input of the reception signal processing
circuit 20. The ultrasonic transducer 14 thus operates as a
receiving transducer; it converts the incoming ultrasonic

1 wave to a reception AC voltage U_E which is supplied to the input of the reception signal processing circuit 20.

When the directional changeover switch 16 is brought into
5 the other position 2 the ultrasonic transducer 14 receives the transmission AC voltage U_S from the transmission frequency generator 18 so that it operates as transmitting transducer and generates an ultrasonic wave which is propagated in the direction of the axis of the measuring tube
10 10 in the medium flowing through the measuring tube 10. Said ultrasonic wave moves oppositely to the flow direction to the ultrasonic transducer 12 which is now connected to the input of the reception signal processing circuit 20 and operates as receiving transducer converting the incoming
15 ultrasonic wave to the reception AC voltage U_E which is supplied to the input of the reception signal processing circuit 20.

The diagram A of Fig. 2 shows the time profile of the
20 transmitting AC voltage U_S furnished by the transmission frequency generator 18 and of the reception AC voltage U_E supplied by the receiving transducer to the reception signal processing circuit 20 in the case where the directional changeover switch 16 assumes the position 1 illustrated in
25 Figure 1, i.e. the ultrasonic transducer 12 operates as transmitting transducer and the ultrasonic transducer 14 as receiving transducer. Between said two AC voltages there is a total phase shift ϕ_1 which results from the travel time required by the ultrasonic wave in the flowing
30 medium to cover the distance L_M from the transmitting transducer to the receiving transducer. This travel time depends on the length of the distance L_M and the velocity of the ultrasonic wave relatively to the stationary ultrasonic transducers 12 and 14. Since in the case of the diagram A
35 of Fig. 2 the ultrasonic wave covers the distance L_M in the

1 flow direction of the medium the relative velocity of the
ultrasonic wave corresponds to the sum of the speed of sound
c in the medium and the flow velocity V_M . The time
the ultrasonic wave requires to travel the distance
5 L_M corresponds to a total phase shift ϕ_1 , which is
given by the following formula:

$$\phi_1 = \omega_M \frac{L_M}{c + V_M} \quad (1)$$

10

where $\omega_M = 2\pi f_M$ is the known angular frequency of
the ultrasonic wave.

15

The total phase shift ϕ_1 is made up of the number m_1
of the whole wavelengths between the two transducers and
the residual phase angle ϕ_{R1} :

$$\phi_1 = \phi_{R1} + m_1 \cdot 2\pi \quad (2)$$

20

25

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In corresponding manner, diagram A of Fig. 3 shows the time
profile of the transmission AC voltage U_S and reception
AC voltage U_E for the case where the directional changeover
switch 16 is in the other position so that the ultrasonic
transducer 14 operates as transmitting transducer and the
ultrasonic transducer 12 as receiving transducer. In this
case the ultrasonic wave passes over the distance L_M oppo-
sately to the flow direction so that the relative velocity
of the ultrasonic wave corresponds to the difference be-
tween the velocity of sound c in the medium and the flow
velocity V_M . Accordingly, on passing through the distance
 L_M the ultrasonic wave undergoes a total phase
shift ϕ_2 which is given by the following formula:

$$\phi_2 = \omega_M \frac{L_M}{c - V_M} \quad (3)$$

35

1 This total phase shift ϕ_2 is in turn made up of
the number m_2 of the whole wavelengths between the two
ultrasonic transducers and a residual phase angle ϕ_{R2} :

$$5 \quad \phi_2 = \phi_{R2} + m_2 \cdot 2\pi \quad (4)$$

By measuring the total phase shift ϕ_1 , the flow
velocity V_M desired can be calculated from formula (1) when
10 the velocity of sound c in the medium is known:

$$V_M = \frac{\omega_M \cdot L_M}{\phi_1} - c \quad (5)$$

Correspondingly, by measuring the total angular phase shift
15 ϕ_2 the flow velocity V_M can be calculated according to the
formula (3) when the velocity of sound c is known:

$$V_M = c - \frac{\omega_M \cdot L_M}{\phi_2} \quad (6)$$

20 The velocity of sound c in the medium is however not con-
stant; it changes in dependence upon the properties of the
medium, in particular the temperature and density thereof.
However, with the alternate measurement of the total ang-
ular phase shifts in the flow direction and opposite to
25 the flow direction it is possible by joint evaluation of
the measurement results to determine the flow velocity V_M
independently of the velocity of sound c in the medium. If
the above equations (5) and (6) are added the result is:

$$30 \quad 2V_M = \frac{\omega_M \cdot L_M}{\phi_1} - \frac{\omega_M \cdot L_M}{\phi_2} \quad (7)$$

$$35 \quad V_M = \frac{\phi_2 - \phi_1}{\phi_2 \cdot \phi_1} \cdot \frac{\omega_M \cdot L_M}{2} \quad (8)$$

1 If for φ_1 and φ_2 the corresponding terms according to equations (2) and (4) are inserted this gives for the flow rate to be measured:

5
$$V_M = \frac{\varphi_{R2} - \varphi_{R1}}{(\varphi_{R2} + m_2 \cdot 2\pi)(\varphi_{R1} + m_1 \cdot 2\pi)} \cdot \frac{\omega_M \cdot L_M}{2} \quad (9)$$

A preferred method of determining the number m_1 or m_2 of whole wavelengths along the distance L_M between the two ultrasonic transducers will be described later. This determination can be made with relatively low measuring accuracy. Furthermore, determination in one direction suffices because with the flow velocities occurring the travel differences are very small and consequently as a rule $m_1 = m_2 = m$. To determine the flow velocity V_M it is then also necessary to measure the residual phase shifts φ_{R1} and φ_{R2} , i.e. the simple phase shifts between the transmission AC voltage U_S and the reception AC voltage U_E . The accuracy with which the flow velocity V_M can be determined thus depends in this measuring method on the accuracy with which the residual phase angles φ_{R1} and φ_{R2} can be measured.

Now, according to formula (9) the flow velocity V_M is proportional to the difference

$$\Delta\varphi = \varphi_{R2} - \varphi_{R1}$$

between the two residual phase angles. With the flow velocities V_M usually to be measured the residual phase angles φ_{R1} and φ_{R2} are of the same order of magnitude; their difference is shown exaggeratedly large in the diagrams of Figs. 2 and 3 only for clarity. Accordingly, the phase difference $\Delta\varphi$ is small compared with the magnitude of each residual phase angle φ_{R1} and φ_{R2} . This results in high demands on the measurement accuracy in measuring the residual phase angles φ_{R1} and φ_{R2} because even small errors in the measure-

1 ment of the two residual phase angles result in relatively large errors in the difference $\Delta\varphi$ of the two residual phase angles.

5 In the measuring arrangement illustrated in Fig. 1 the measurement of the residual phase angles φ_{R1} and φ_{R2} is carried out digitally. For this purpose, the reception AC voltage U_E is transformed in the reception signal processing circuit 20 to a square-wave voltage Q_E in phase there-
10 with and a transmission signal processing circuit 22 is connected to the output of the transmission frequency generator 18 and in said circuit the transmission AC voltage U_S is converted to a square-wave voltage Q_S in phase therewith. The diagrams B and C of Fig. 2 show the square-wave voltages
15 Q_S and Q_E respectively for the case where the ultrasonic wave is propagated between the two ultrasonic transducers 12 and 14 in the direction of flow and the diagrams B and C of Fig. 3 show the square-wave voltages Q_S and Q_E for the case where the ultrasonic waves are propagated between the
20 two ultrasonic transducers 12 and 14 oppositely to the flow direction. The edges of the square-wave voltages coincide with the zero passages of the AC voltages from which they are derived. Between the corresponding edges, i.e. for example between the rising edges of the square-wave voltages
25 Q_S and Q_E , there is therefore the same phase shift as between the corresponding zero passages of the AC voltage U_S and U_E ; thus, in the case of Fig. 2 this phase shift corresponds to the residual phase angle φ_{R1} and in the case of Fig. 3 to the residual phase angle φ_{R2} .

30

As is apparent from the timing diagram of Fig. 2, the residual phase angle φ_{R1} corresponds to a definite time interval Δt_1 between corresponding edges of the square-wave voltage Q_S and Q_E because with the known angular frequency ω_M the
35 following relationship is true:

1

$$\phi_{R1} = \omega_M \cdot \Delta t_1 \quad (10)$$

5

Similarly, in accordance with the timing diagram of Fig. 3, the residual phase angle ϕ_{R2} corresponds to an exactly defined time interval Δt_2 between the corresponding edges of the square-wave voltages Q_S and Q_E in accordance with the relationship

10

$$\phi_{R2} = \omega_M \cdot \Delta t_2 \quad (11)$$

Thus, to measure the residual phase angles ϕ_{R1} and ϕ_{R2} , it suffices to measure the time intervals Δt_1 and Δt_2 , respectively, which are proportional thereto.

The outputs of the two signal processing circuits 20 and 22 are connected to the two inputs of a phase detector 24 which at the output furnishes a train of square-wave pulses I_M which are represented in the diagrams D of Figs. 2 and 3 and referred to hereinafter as measuring pulses. The pulse repetition frequency of the measuring pulses I_M corresponds to the frequency f_M generated by the transmission frequency generator 18 and is referred to hereinafter as measuring frequency. Each measuring pulse I_M has a duration T_I which corresponds to the time interval Δt_1 and Δt_2 respectively between the rising edges of the two square-wave voltages Q_S and Q_E in the diagrams B and C. Such a square-wave pulse train can be generated very easily with the aid of a flip-flop which is set by the rising edge of the square-wave voltage Q_S and reset by each rising edge of the square-wave voltage Q_E .

Thus, to measure the time intervals Δt_1 and Δt_2 , which are proportional to the residual phase angles ϕ_{R1} and ϕ_{R2} , respectively, the duration T_I of the measuring pulses I_M has to be measured. This time measurement is done digi-

1 tally in the measuring arrangement of Fig. 1 as will be
explained with the aid of the diagrams of Fig. 4, where a
train of measuring pulses I_M of duration T_I furnished by the
phase detector 24 is again shown in diagram A. The output
5 of the phase detector 24 is connected to the control input
of a gate circuit 26 of which the signal input is connected
to the output of a counting frequency generator 28. The
counting frequency generator 28 generates a train of count-
ing pulses I_C having a repetition frequency f_C substantially
10 greater than the measuring frequency f_M . As an example, in
diagram B of Fig. 4 such a train of counting pulses I_C is
illustrated. The time interval of the counting pulses
 I_C corresponds to the period duration T_C with the counting
frequency f_C . For clarity, in diagram B of Fig. 4 the
15 period duration T_C of the counting pulse train I_C is exagg-
erated in magnitude with respect to the period duration T_M
of the measuring pulses I_M of the diagram A.

The output of the gate circuit 26 is connected to the
20 counting input of a counter 30 of which the stage outputs
are connected to corresponding inputs of an evaluating
circuit 32 which for example may be a suitably programmed
microcomputer.

25 The gate circuit 26 is so configured that it is open for the
duration of each measuring pulse I_M applied to its control
input and in the opened state allows the counting pulses
 I_C coming from the counting frequency generator 28 to pass
to the counter 30 as is shown in diagram C of Fig. 4. When
30 no measuring pulse I_M is applied to the control input of the
gate circuit 26 the latter blocks the transmission of the
counting pulses I_C to the counter 30. The counting pulses
allowed to pass by the gate circuit 26 in the duration of
each measuring pulse I_M are counted in the counter 30 so
35 that the latter indicates a count corresponding to the
number of counting pulses output in this duration by the
counting frequency generator 28. Thus, said count is a

1 measure of the duration of the measuring pulse I_M . After
the end of each measuring pulse I_M the count of the counter
30 is transferred to the evaluating circuit 32 and the
counter 30 reset to zero.

5

The evaluating circuit 32 thus continuously receives from
the counter 30 counts which represent the duration of the
measuring pulses I_M furnished by the phase detector 24.
When the directional changeover switch 16 is in the position
10 1 (Fig. 1) the counts furnished by the counter 30 are a
measured value for the residual phase angle ϕ_{R1} . When the
directional changeover switch 16 is brought to the other
position the counts furnished by the counter 30 are a
measured value for the residual phase angle ϕ_{R2} .

15 In accordance with the above formula (9), from these two
measured values the evaluating circuit 32 can calculate the
flow velocity V_M independently of the speed of sound c in
the medium.

20 The fundamental problem caused by the quantizing error of
the digital counting method is also apparent from the
diagrams of Fig. 4: The resolution of the time measurement,
i.e. the smallest time difference which can still be dis-
tinguished, corresponds to the period T_C of the counting
25 frequency f_C . To increase the resolution and thus the
measurement accuracy it would therefore be necessary to
make the counting frequency as high as possible. However,
there are limits to this solution due to the expenditure
involved. In the case of the flow velocities to be measured
30 in practice with the ultrasonic frequencies employed the
pulse durations T_I of the measuring pulses T_M corresponding
to the residual phase angles ϕ_{R1} and ϕ_{R2} are of the order
of magnitude of microseconds and the phase differences $\Delta\phi$
to be determined in accordance with the formula (9) between
35 the residual phase angles correspond to fractions of a
nanosecond. Consequently, extremely high counting frequen-
cies f_C would be necessary in order to select each pulse

1 duration T_I with an adequately high number of counting pulses I_C .

Fig. 5 shows the block diagram of part of the
5 measuring arrangement of Fig. 1 for a modified embodiment which, with a relatively low counting frequency f_C results in a considerable increase in the resolution in the digital phase measurement and accordingly an increase in the measuring accuracy in the measurement of the flow velocity V_M . In
10 so far as the components of the measuring arrangement of Fig. 5 correspond to those of the measuring arrangement of Fig. 1 they are denoted with the same reference numerals as in Fig. 1 and will not be described again.

15 Thus, Fig. 5 again shows the two signal processing circuits 20 and 22, the phase detector 24, the gate circuit 26, the counting frequency generator 28, the counter 30 and the evaluating circuit 32. The measuring arrangement of Fig. 5 differs from that of Fig. 1 in that the measuring frequency
20 f_M is derived from the counting frequency f_C , this being done in particular manner in that the measuring frequency and the counting frequency remain constant, but in addition the start phase ϕ_S (Fig. 4) which the counting pulses I_C have at the start of each counting operation with respect
25 to the phase of the measuring pulses I_M is continuously rotated so that after n measurements a phase angular shift of 2π is achieved, i.e. one complete period of the counting frequency f_C .

30 For this purpose, a divide-by- N frequency divider 34 is connected to the output of the counting-frequency generator 28. The division factor N determines the ratio between the counting frequency f_C and the output frequency
35 f_M' of the frequency divider 34. If, for example, the counting-frequency generator 28 generates the counting frequency $f_C = 7$ MHz, and the frequency divider 34 divides by $N = 128$, a frequency f_M' of about 55 kHz is obtained.

1 The frequency divider 34 is followed by a controllable
phase shifter 36 which at its output furnishes an AC voltage
with the frequency f_M which compared with the AC voltage
5 applied to its input with the frequency f_M' has a phase
shift which depends on a phase control voltage applied to
its control input 36a. The control input 36a is connected
to the output of a phase control circuit 38 which after
each activation by a trigger signal applied to a trigger
10 input 38a furnishes at the output a ramp voltage U_R with
the time profile illustrated in diagram A of Fig. 6. The
ramp voltage U_R increases for a period T_R linearly from a
value U_{DC} to a value U_S and then jumps back to the value U_{DC} .
The period T_R corresponds to n periods of the measuring
frequency f_M ; the following relationship is thus true:

15

$$T_R = n \cdot T_M = \frac{n}{f_M} \quad (12)$$

20 The ramp voltage U_R is so dimensioned that it rotates the
start phase φ_S which the counting pulse train I_C has with
respect to the start of each measuring pulse I_M in the
course of the period T_R , i.e. in n consecutive measuring
pulses I_M , through 2π , i.e. one complete period of the
counting frequency f_C . This results in the time profile
25 of the start phase φ_S illustrated in diagram B of Fig. 6.
A constant start phase φ_0 corresponds to the initial value
 U_{DC} of the ramp velocity U_R and the start phase $\varphi_0 + 2\pi$ cor-
responds to the end value U_S of the ramp voltage U_R . Since
in each period of the measuring frequency f_M a phase measure-
30 ment operation is carried out by counting the counting
pulses I_C in the counter 30, the start phase φ_S shifts
continuously from measuring operation to measuring operation.
In the p^{th} measurement in the course of the period T_R it
has the value

35

$$\varphi_S(p) = p \cdot \frac{2\pi}{n} + \varphi_0 \quad (13)$$

(with $p = 0 \dots n$)

1 However, the phase shifter 36 no longer performs the phase
shift on the counting frequency f_C but on the frequency
5 f_M' derived therefrom, for which the following is true:

$$f_M' = \frac{f_C}{N} \quad (14)$$

10 Between the start phase φ_S at the counting frequency f_C and
the phase φ_M at the measuring frequency f_M' , the following
relation holds:

$$\varphi_S = N \cdot \varphi_M \quad (15)$$

15 The phase shift to be generated by the phase shifter 36
must thus be a factor N times smaller than the desired phase
shift of the start phase φ_S as illustrated in diagram C of
Fig. 6. Between the values U_{DC} and U_S of the ramp voltage
20 U_R the phase shift generated by the phase shifter 44 changes
by $2\pi/N$. In the p^{th} measurement it has in the course of the
period T_R the value

$$\varphi_M(p) = p \cdot \frac{2\pi}{N \cdot n} + \varphi_0 \quad (16)$$

25

(with $p = 0 \dots n$)

30 The continuous phase shift between the frequency f_M' and
 f_M during the period T_R of each ramp of the ramp voltage U_R
is equivalent to a frequency change. The following is then
true

$$f_M = f_M' \left(1 + \frac{1}{N \cdot n} \right) \quad (17)$$

35

Since, however, $N \cdot n \gg 1$

1 we have approximately

$$f_M \approx f_M' \quad (18)$$

5 The output voltage of the phase shifter 36 can represent
the transmitting AC voltage U_S directly. The transmission
frequency generator 18 of Fig. 1 is thus replaced in the
measuring arrangement of Fig. 5 by the frequency divider 34
and the phase shifter 36 in conjunction with the phase
10 control circuit 38.

The continuous variation of the start phase ϕ_S by 2π in the
course of n consecutive measurements has the effect that
during part of the measurements a predetermined number k
15 and during the remaining measurements a number $k+1$ of
counting pulses I_C , greater by 1, is counted during the
duration T_I of each measuring pulse T_M . By suitable eval-
uation of the n measured values obtained in each measuring
cycle of duration T_R the resolution and thus the measuring
20 accuracy can be considerably increased without correspondingly
increasing the counting frequency f_C . The continuous
phase shift through 2π in n consecutive stages corresponds
as it were to a division of the quantizing step corresponding
to the counting period T_C into n substeps. If it were
25 possible to completely eliminate all other random errors,
such as thermal noise, flow noise, etc., this would give
an increase of the resolution by the factor n .

The preferred evaluation of the measurement results obtained
30 in this manner reside in that the mean value of the n counts
obtained in the course of each measuring cycle is formed.
The residual phase angle corresponding to the mean value
is then employed to calculate the flow velocity V_M in
accordance with the formula (9).

35 The activation of the phase control circuit 38 is preferably
in time with the switchover of the directional changeover

1 switch 16. The n measured values for the measurement with
the flow direction and in the following measuring cycle the
5 n measured values for the measurement against the flow
direction are then obtained alternately in a measuring
cycle of duration T_R . Thus, after every two consecutive
measuring cycles the two residual phase angles required for
the calculation of the flow velocity in accordance with
equation (9) are available with increased measuring accuracy.

10 Fig. 7 shows an example of embodiment of the controllable
phase shifter 36 in greater detail. Fig. 7 also shows the
counting frequency generator 28, the frequency divider 34
and the phase control circuit 38.

15 The counting frequency generator 28 is synchronously phase-
locked by a quartz generator 40. This synchronization en-
sures that the start phase is turned exactly through one
whole period 2π of the counting frequency f_C over the n
measurements in each measuring cycle of duration T_R . Further-
20 more, the counting frequency generator 28 is preferably so
constructed that the counting frequency f_C is adjustable in
a relatively large range, for example between 2 and 11 MHz.
To obtain a phase-locked control in conjunction with an
adjustable frequency the counting frequency generator 28
25 may be provided in known manner with a phase-locked circuit
known under the designation PLL (phase locked loop). Such
a PLL circuit includes a digital frequency divider having
a division factor which is adjustable by a digital control
circuit and governs the frequency furnished at the output.
30 It is indicated in Fig. 7 that the digital control signal is
supplied to a control input 28a of the counting frequency
generator 28.

35 The phase shifter 36 is likewise formed by a PLL circuit
50. The PLL circuit 50 contains in cascade a phase compar-
ator 52, a low-pass filter 54, a summation circuit 56 and
a voltage-controlled oscillator 58 also known under the

1 term VCO (voltage-controlled oscillator). Such a voltage-
controlled oscillator furnishes of course at its output an
AC voltage having a frequency depending on a frequency
control voltage applied to its control input.

5

The phase comparator 52 receives at one input the output
signal of the frequency divider 34 and at the other input
the output AC voltage of the frequency-controlled oscillator
58. It furnishes at the output a control deviation signal
10 depending on the phase difference between its two input sig-
nals. The control deviation signal is supplied after fil-
tering in the low-pass filter 54 to the one input of the
summation circuit 56 which at the other input receives the
ramp voltage U_R from the output of the phase-control circuit
15 38. The summation voltage formed in the summation circuit
56 is supplied as frequency control voltage to the control
input of the voltage-controlled oscillator 58.

The PLL circuit 50 forces frequency identity between the
20 output frequency f_M' of the frequency divider 34 supplied
to the one input of the phase comparator 52 and the output
voltage f_M of the voltage-controlled oscillator 58 supplied
to the other input of the phase comparator 52. To ensure
that this frequency identity is retained the frequency
25 control voltage supplied to the voltage-controlled oscillator
58 must always have the constant value corresponding to the
frequency f_M . Now, however, this frequency control voltage
derives from the sum of the continuously rising ramp vol-
tage U_R and the output voltage of the phase comparator 52.
30 To ensure the frequency control voltage remains constant
the output voltage of the phase comparator 52 must therefore
be varied oppositely to the ramp voltage U_R . The necessary
change in the output voltage of the phase comparator 52 is
obtained as a result of a change in the mutual phase pos-
35 ition of its two input voltages. The PLL circuit therefore
compels a continuous shift of the phase position of the out-
put AC voltage of the voltage-controlled oscillator 52 with

1 respect to the phase position of the output AC voltage of
the frequency divider 34 for compensating the continuously
increasing ramp voltage U_R . Thus, the embodiment of the
phase shifter 36 illustrated gives exactly the effect out-
5 lined previously with reference to Fig. 5.

The phase-control circuit 38 of Fig. 7 includes a sawtooth
generator 60 and a parameter control circuit 62. The para-
meter control circuit 62 governs in particular the two
10 voltage values U_{DC} and U_S between which the ramp voltage
 U_R changes. With the parameter U_{DC} the constant phase
angle ϕ_0 can be adjusted. With the parameter U_S the max-
imum phase shift over the duration T_R of each rising edge of
the ramp voltage U_R may be set. The parameter control cir-
15 cuit 62 also receives the counting frequency f_C so that the
activation of each ramp by the trigger signal supplied to
the input 38a takes place in a defined time relationship
to the counting pulses.

20 Fig. 8 shows a modified embodiment of the arrangement of
Fig. 5 in which the phase-control circuit is formed by a
digital-analog converter 70 which converts a digital code
group furnished by the evaluating circuit 32 to an analog
voltage which is applied as phase control voltage to the
25 input 36a of the phase shifter 36. The microcomputer form-
ing the evaluating circuit 32 can thus control the variation
of the start phase ϕ_S within the course of a measuring
cycle in accordance with a predetermined program. In this
manner the exact measured value of the residual phase
30 angle may also be obtained in a way different to the form-
ation of the mean value over n consecutive measurements.
In particular, a successive approximation may be carried out
by which by a weighting method the transition from k count-
ing pulses to $k+1$ counting pulses per measuring pulse dur-
35 ation T_I is approached. The phase shifter 36 is driven in
discrete steps by the microcomputer 32 via the digital-
analog converter 70. An increase in the resolution by the

1 factor 2^p can then be obtained in p steps.

Fig. 9 shows a further embodiment of the flow rate measuring arrangement with which a continuous change of the start phase φ_S by 2π in the course of n consecutive measuring pulses I_M is obtained by employing two frequencies displaced with respect to each other. Fig. 9 again shows the gate circuit 26 to the control input of which the measuring pulses I_M of duration T_I and repetition frequency f_M are applied along with the counter 30 connected to the output of the gate circuit 26 and the evaluating circuit 32. The counting pulses I_C with the counting frequency f_C are furnished by the output 80a of a differential counting frequency generator 80 which furnishes at a further output 80b a frequency f_C' which differs by a differential frequency Δf from the counting frequency f_C , whereby:

$$f_C' = f_C \pm \Delta f \quad (19)$$

20 The measuring frequency f_M is derived by frequency division in the frequency divider 34 with the division factor N from the frequency f_C' . Accordingly, the counting frequency f_C is not an integer multiple of the measuring frequency f_M and consequently a continuous phase shift of the start phase φ_S of the counting takes place with respect to the measuring pulses I_M in consecutive periods of the measuring frequency f_M . By suitable dimensioning of the differential frequency Δf it may be achieved that the start phase φ_S changes in n consecutive periods of the measuring frequency f_M by precisely 2π , i.e. by one whole period of the counting frequency f_C . This is the case when the differential frequency Δf has the following value:

$$\Delta f = \frac{f_M}{n} \quad (20)$$

35 If, for example, the measuring frequency $f_M = 55.6$ kHz and

1 the phase rotation through 2π is to be obtained in $n = 1000$
consecutive periods of the measuring frequency f_M , the
differential frequency is:

5
$$\Delta f = \frac{55.6 \text{ kHz}}{1000} = 55.6 \text{ Hz} \quad (21)$$

For the ratio between the counting frequency f_C and the
measuring frequency f_M the following applies:

10
$$f_C = f_C' \pm \frac{f_M}{n} \quad (22)$$

$$f_C = N \cdot f_M \pm \frac{f_M}{n} \quad (23)$$

15 Fig. 10 shows an example of embodiment of the differential
counting frequency generator 80 of Fig. 9 with which two
frequencies f_C and f_C' can be obtained which differ exactly
by the desired differential frequency Δf . Said different-
20 ial counting frequency generator contains a phase comparator
81 which at one input receives a reference frequency f_{Ref}
which is preferably derived from a quartz oscillator and
is for example 1 kHz. The output signal of the phase com-
parator 81 is supplied after filtering in a low-pass filter
82 to a first voltage-controlled oscillator 83 directly and
25 to a second voltage-controlled oscillator 84 via a summa-
tion circuit 85 for frequency control. The voltage-con-
trolled oscillator 83 furnishes at the output the frequency
 f_C' which after frequency division with the division factor
 N in a frequency divider 86 (which corresponds to the fre-
30 quency divider 34 of Fig. 9) gives the measuring frequency
 f_M . Since the output voltage R_S of the frequency divider
86 is a square-wave voltage it is transformed in a signal
shaper 87 to a sinusoidal voltage of the same frequency f_M
which then forms the transmission AC voltage U_S .

35

The output of the frequency divider 86 is further connected

1 via a second frequency divider 88 to the second input of
the phase comparator 81. In this manner a PLL circuit of
known type is formed which by phase-locked control keeps
the measuring frequency f_M in a ratio to the reference
5 frequency f_{Ref} governed by the division factor of the second
frequency divider 88. The division factor of the frequency
divider 88 is preferably digitally adjustable so that the
measuring frequency f_M can be adjusted within a certain
range to a desired value. The constant ratio governed by
10 the division factor N of the frequency divider 86 is always
retained between the set value of the measuring frequency
 f_M and the frequency f_C' .

The output of the voltage-controlled oscillator 83 is fur-
15 ther connected to the one input of a phase comparator 90,
to the other input of which the output signal of the other
voltage-controlled oscillator 84 with the frequency f_C is
supplied. The output of the phase comparator 90 is con-
nected via a low-pass filter 91 to a square-wave signal
20 shaper 92 which is formed for example by a Schmitt-trigger.
At the output of the square-wave signal shaper a square-
wave signal having the frequency

$$\Delta f = |f_C' - f_C| \quad (24)$$

25 is obtained which is supplied to the one input of a phase
comparator 93. The other input of the phase comparator
93 receives the output signal of a frequency divider 94 which
divides the output frequency f_M of the frequency divider 86
30 by the factor n . The output of the phase comparator 93 is
connected via a low-pass filter 95 to the second input of
the summation circuit 85, the output voltage of which forms
the frequency control voltage for the voltage-controlled
oscillator 84. The voltage-controlled oscillator 84 thus
35 lies in a frequency-control loop which keeps its output
voltage f_C at a constant frequency distance Δf from the out-
put frequency f_C' of the voltage-controlled oscillator 83,

1 said frequency difference Δf always being kept equal to the output frequency of the frequency divider 94. Consequently:

5

$$\Delta f = \frac{f_M}{n} \quad (25)$$

10

$$f_C = N \cdot f_M \pm \frac{f_M}{n} \quad (26)$$

The circuit of Fig. 10 thus furnishes at the output of the voltage-controlled oscillator 84 a counting frequency f_C which fulfills the condition given by equation (23).

20 In all the previously described embodiments of the flow rate measuring arrangement, the counting pulses are generated continuously, but to increase the measuring accuracy in the phase difference measurement, different measures are employed to cause a systematic change of the start phase at the beginning of each measuring time interval. The embodiment of the flow rate measuring arrangement illustrated in Fig. 11 differs from these embodiments fundamentally in that in each case a counting pulse train with repetition frequency f_C is generated only during each measuring time interval, the influencing of the start phase necessary to increase the measuring accuracy being effected in that each counting train pulse starts at the beginning of a measuring time interval exactly with the phase position with which the preceding counting train pulse was terminated at the end of the preceding measuring time interval.

35 Fig. 11 again shows the phase detector 24, the counter 30 and the evaluating circuit 32 of Fig. 1. The output of the

phase detector 24 at which the measuring pulses I_M with the duration T_I appear is connected to a control input of a triangle generator 124. The output voltage U_D of the triangle generator 124 is supplied to the input of a counting pulse shaper 126 whose output provides the counting pulses I_C , which are supplied to the counter 30.

The operation of the circuit of Fig. 11 will now be explained with the aid of the diagrams of Fig. 12. In Fig. 12, diagram A shows the measuring pulses I_M provided by the phase detector 24, diagram B the output voltage U_D of the triangle generator 124, and diagram C the counting pulses delivered by the counting pulse shaper 126 to the counter 30.

The triangle generator 124 is constructed in such a manner that it generates a triangular voltage of frequency f_C only when a measuring pulse I_M is applied to its control input. On termination of the measuring pulse I_M the triangle generator 124 is stopped, so that it reaches a hold state. This hold state is characterized by the fact that the output voltage U_D is not reset but retains the signal level reached on stopping. This signal level may be stored in a capacitor, for example. At the start of the next measuring pulse I_M the hold state is terminated and the triangle generator 124 continues the generation of the triangular voltage from the stored signal level onwards.

The counting pulse shaper 126 changes the output voltage U_D of the triangle generator 124 into counting pulses I_C which are phase-locked to that output voltage. This can be done in a simple manner by initiating the generation of a counting pulse when the voltage U_D exceeds a predetermined threshold value U_{D1} , and terminating the counting pulse when the voltage U_D drops below a predetermined lower threshold value U_{D2} . Thus, for the duration of

each measuring pulse I_M , the counting pulse shaper 126
furnishes at its output a number of
counting pulses I_C which is equal to the number of periods of
the triangular voltage generated in this duration. In con-
trast, no counting pulses are generated during the interval
between two measuring pulses I_M . By storing the signal
level of the voltage U_D it is ensured that the last count-
ing pulse period T_C which started at the end of a measuring
time interval of duration T_I but was not longer completed
is completed at the start of the next measuring time inter-
val. Consequently, with these counting methods the durations
 T_I of the measuring pulses I_M occurring successively in the
course of a measuring cycle can nevertheless follow each
other without gaps and can be measured without error by
adding up the corresponding counts. For example, in diagram
C of Fig. 12, in the first measuring time interval five com-
plete counting pulses I_C occur. The incomplete pulse gap
after the last counting pulse is completed at the start of
the second measuring time interval. Likewise, in the second
measuring time interval five complete counting pulses appear
and a started sixth counting pulse which is completed at
the beginning of the third measuring time interval. If the
counter 30 responds in each case to the rising edges of the
counting pulses I_C , in the first measuring time interval the
count "5" results and in the second measuring time interval
the count "6". Thus, a so-called "residual phase error"
which arose in system-inherent manner in the first measuring
time interval is cancelled out in the second measuring time
interval. Generally speaking this means that a residual
phase error accumulating in several consecutive measuring
time intervals is compensated in a following measuring
time interval.

In all the embodiments of the flow meter described above it
is presumed that the acoustic wave is propagated in the
direction of the axis of the measuring tube 10 from the
transmitting transducer to the receiving transducer. Of

1 course, this can be achieved not only by the arrangement of
the ultrasonic transducers shown in Fig. 1 but naturally
also by arranging them in such a manner that the major
propagation of the ultrasonic waves does not coincide with
5 the measuring tube axis as in Fig. 1.

10 The two ultrasonic transducers 12 and 14 of Fig. 1 must be
so configured that when operated as transmitting transducers
they can generate such an acoustic wave and when operated
as receiving transducers they can convert such an acoustic
wave with adequate sensitivity to an electrical receiving
signal. In addition, in particular with flow meters of
small nominal widths, it is desirable for the flow cross-
section of the flow meter not to be restricted by the ultra-
sonic transducers. Fig. 13 shows a preferred embodiment of
15 an ultrasonic transducer which fulfills these
conditions.

20 The ultrasonic transducer 130 illustrated in Fig. 13 con-
sists of an annular piezocrystal 132 which is provided at
its two end faces with metallizations 134 and 136 serving
as electrodes. The piezocrystal may consist of lead-zircon-
ium titanate, barium titanate or another suitable piezo-
electric material. The ultrasonic transducer 130 is accomm-
25 odated in a chamber 140 which is formed by recesses in a
flange 142 attached to the measuring tube 110 and in an
annular counter plate 146 secured by means of threaded bolts
144 to the flange 142. The counter plate 146 consists of
the same material as the measuring tube 10 and the flange
30 142, this material preferably being stainless steel.

To prevent the piezocrystal 132 from being attacked by a
chemically aggressive medium, on its inner peripheral sur-
face a protective layer 138 of a material resistant to the
chemically aggressive medium is applied. Depending on the
35 nature of the chemically aggressive medium the protective
layer 138 may consist of metal, such as gold, silver or

1 titanium, or of glass, plastic, ceramic or the like. It is
particularly favourable for the material of the protective
layer 138 to have an acoustic characteristic impedance
5 which lies between the acoustic characteristic impedance of
the piezocrystal 132 and the acoustic characteristic impedance
of the medium so that the protective layer effects an
acoustic impedance matching at the same time. If the medium
is not aggressive the protective layer 138 may be
omitted so that the inner peripheral surface of the piezo-
crystal 132 is then in direct contact with the medium flowing
10 through the flow meter.

The annular piezocrystal 132 is clamped between the flange
142 and the counter plate 146 with insertion of sealing
rings 150 and 152. The sealing rings 150, 152 and the
15 piezocrystal 132 with the protective layer 138, the flange
142 and the counter plate 146 have the same internal diameter
as the measuring tube 10 so that a continuously uniform flow
cross-section results which is not impaired by installation
of the ultrasonic transducer. The sealing
20 rings 150, 152 serve on the one hand to seal the clamping
point and on the other for acoustic uncoupling of the ultrasonic
transducer 130 from the tube sections adjoining in both
directions. The material of the sealing rings 150, 152
is selected so that it fulfills these two purposes as well
25 as possible and is also resistant to the media flowing
through the flow meter. To obtain good acoustic uncoupling
the acoustic characteristic impedance of the sealing rings
150, 152 should be substantially smaller than the acoustic
characteristic impedance of the measuring tube 10. Viton
30 has proved to be a particularly suitable material for the
sealing rings 150, 152. On assembling together the flange
142 and the counter plate 146 the sealing rings 150, 152
are given a predetermined compression.

35 To ensure that the ultrasonic transducer 130 generates in
the medium an acoustic wave propagated in the axial direction

1 it is excited for the phase measurement by the transmitting
AC voltage U_s applied to its electrodes 134, 136 in such
a manner that it oscillates radially, its internal diameter
and external diameter preferably varying in phase. The
5 variations of the internal diameter cause pressure fluctuations
in the medium which are propagated as ultrasonic
waves. Although the electrodes 134 and 136 lie parallel to
the desired oscillation direction a radial oscillation can
be generated in that the frequency f_M of the transmitting
10 AC voltage U_s is made equal to the radial resonant frequency
of the piezocrystal 132. This radial resonant frequency is
lower than the longitudinal resonant frequency at which the
piezocrystal 132 oscillates parallel to the axial direction,
i.e. perpendicularly to the electrodes 134 and 136.

15 Although application of an AC voltage between the electrodes
134 and 136 first causes thickness changes of the piezo-
crystal 132 perpendicularly to said electrodes, these thick-
ness changes are accompanied by contractions in the radial
20 direction which build up to radial oscillations when the
condition of radial resonance is fulfilled.

The radial resonant frequency depends on the dimensions of
the piezocrystal 132 and can therefore be adjusted by select-
25 ing these dimensions. In order to measure in accordance
with the theory a mean flow velocity independent of the
flow profile planar waves must be propagated in the measur-
ing tube 10. To ensure that planar waves can form in a tube
two conditions must be fulfilled:

- 30 1. the wavelength in the medium must be appreciably greater
than the internal diameter of the measuring tube, pref-
erably greater than 1.6 times said internal diameter;
- 35 2. the acoustic characteristic impedance (acoustic stiffness)
of the measuring tube must be substantially greater than
that of the medium.

1 The first condition is fulfilled in that the radial resonant frequency of the piezocrystal 132 is set by appropriate
choice of its dimensions to a value at which the wavelength
5 of the ultrasonic waves in the medium is greater than the internal diameter of the measuring tube. With an internal
diameter of the measuring tube of 8 mm this condition is for
example fulfilled when the radial resonant frequency of the
piezocrystal 132 and thus the frequency f_M of the trans-
mitting AC voltage U_S is about 62 kHz.

10 The second condition is fulfilled by suitable choice of the material of the measuring tube 10. For example, a measuring
tube of stainless steel has an acoustic characteristic
impedance which is large compared with the acoustic char-
acteristic impedance of any flow medium occurring.

15 The two ultrasonic transducers of the flow rate measuring arrangement have the same structure and accordingly also
the same radial resonant frequency. If the ultrasonic
20 wave generated by one ultrasonic transducer meets the other ultrasonic transducer the piezocrystal 132 thereof is
stimulated by the pressure fluctuations to radial oscillations at the radial resonant frequency. These radial
oscillations are accompanied by contractions in the long-
25 itudinal direction by which an AC voltage of the same frequency is generated between the electrodes 134 and 136 and
represents the receiving AC voltage U_E . Thus, any ultrasonic transducer constructed in the manner of Fig. 13 can
operate selectively as transmitting transducer or receiving
30 transducer.

35 Since the ultrasonic transducers are operated in resonance a very good power transmission is obtained from the transmitting
transducer to the receiving transducer. This is
of great advantage because it makes it possible to obtain
with relatively low transmitting power an adequately large

receiving AC voltage of, for example, 100 mV. A further advantage of the resonance operation of the relatively narrow band transducers is that interfering noises in the medium caused, for example, by pumps, valves or the like are filtered out directly at the measuring pickup.

Since the ultrasonic waves generated by the transmitting transducer are propagated in the flowing medium in opposite directions, it is necessary to attenuate the ultrasonic waves travelling away from the measuring tube to ensure that no measuring errors are caused by sound reflections in the conduit system containing the measuring tube. For this purpose, an attenuator tube 160 is arranged on the side of the ultrasonic transducer 130 opposite the measuring tube 10. The attenuator tube 160 consists of a steel tube 162 lined on the inside with a nonporous damping material 164. Formed integrally with the steel tube 162 is a flange 166 which is secured by means of screws 169 to the counter plate 146, with a seal 168 inserted therebetween.

Teflon has proved to be particularly well suited as damping material. Of course, the internal diameter of the steel tube 162 is dimensioned corresponding to the thickness of the damping material 164 so that the free internal diameter of the attenuator tube 160 is equal to the internal diameter of the measuring tube 10.

Since the flow meter comprises at each end of the measuring tube 10 an ultrasonic transducer of the type illustrated in Fig. 13, on each side of the measuring tube an attenuator tube is disposed so that the overall construction illustrated in Fig. 14 is obtained. Fig. 14 shows the measuring tube 10 with the flanges 142 integrally formed at its two ends and the counter plates 146, which surround the chambers in which the piezoelectric transducers are arranged. At the flanges 142 the passages 170 for the electric leads to the electrodes of the piezocrystals are also

1 illustrated. On each side of the measuring tube 10 an atten-
uator tube 160 is arranged which is secured by means of its
flange 166 to the associated counter plate 146. Each at-
tenuator tube 160 comprises at the other end a flange 172
5 for connection to the conduit system.

10 In order to be able to calculate the flow velocity V_M in
accordance with equation (9), apart from the residual phase
angles it is also necessary to know the number m_1 or m_2 of
the whole wavelengths along the distance L_M between the two
ultrasonic transducers. As explained with regard to equation
(9) it may be assumed that the number of the whole wave-
lengths in the two measuring directions is the same and
consequently:

15

$$m_1 = m_2 = m \quad (29)$$

It therefore suffices to determine the number of the whole
wavelengths in one measuring direction. Since depending on
20 the velocity of sound in the medium a different number m
results the number m must be measured. The necessary
accuracy is therefore relatively low. Errors which may
arise at the transition from m to $m+1$ or $m-1$ can easily be
determined by a plausibility test and corrected.

25

Fig. 15 shows a block circuit diagram of an arrangement
which is added to the flow rate measuring arrangement of
Fig. 1 to determine the number m of the whole wavelengths
by a travel time measuring method. This arrangement is
30 particularly suitable for the case where the ultrasonic
transducers are formed by annular piezoelectric transducers
of the type illustrated in Fig. 13. In Fig. 15, of the
flow rate measuring arrangement of Fig. 1 for simplification
apart from the measuring tube 10 with the two ultrasonic
transducers 12 and 14 and the directional changeover switch
35 16 only the transmitting frequency generator 18 is illus-
trated. The remaining components of the flow rate measuring

1 arrangement of Fig. 1 are required only for the phase measurement but not for determination of the number m of the entire wavelengths.

5 In Fig. 15, between the transmission frequency generator 18 and the transmitting-side input of the directional change-over switch 16, a changeover switch 200 is inserted which in the illustrated position 1 separates the directional change-over switch 16 from the output of the transmission frequency generator 18 and connects it to the output of a pulse generator 202. For the phase measurement the changeover switch 200 is brought into the other position 2 in which it separates the directional changeover switch 16 from the output of the pulse generator 202 and connects it to the output of the transmission frequency generator 18 so that the flow rate measuring arrangement corresponds to the illustration of Fig. 1.

20 The pulse generator 202 has a trigger input which is connected to the output of a start circuit 204. The start circuit 204 has in turn a trigger input 204a and a synchronizing input 204b connected to an output of the transmission frequency generator 18 at which a square-wave signal R_S is furnished which has the same frequency f_M and the same phase position as the transmission AC voltage U_S . If the transmission frequency generator corresponds to the embodiment of Figs. 9 and 10 such a square-wave signal is for example available at the output of the frequency divider 34 or 86, respectively. Moreover, in all the examples of embodiment there is a corresponding square-wave signal in the form of the signal Q_S at the output of the transmission signal processing circuit 22.

35 When a trigger pulse I_A is applied to the trigger input 204a of the start circuit 204 the latter sends to the pulse generator 202 a start pulse I_S which has an exactly defined

1 time position relatively to the square-wave voltage R_S
furnished by the transmission frequency generator 18 and
thus also to the transmission AC voltage U_S . On every
5 reception of a start pulse the pulse generator 202 generates
a short single pulse I_D . When the changeover switch 200 is
in position 1 said single pulse I_D is supplied to the ultra-
sonic transducer 12 or 14 which is selected as transmitting
transducer by the setting of the directional changeover
switch 16.

10 A pulse receiving circuit 210 is connected to the receiving-
side output of the directional changeover switch 16 parallel
to the reception signal processing circuit 20 (Fig. 1) (not
illustrated in Fig. 15) and includes in series an input
15 amplifier 212, a rectifier circuit 214 and a threshold
value switch 216. The output of the threshold value switch
216 is connected to a stop input 218a of a start-stop logic
218 which receives at its start input 218b the start pulse
 I_S generated by the start circuit 204. The start-stop
20 logic 218 further comprises a signal input 218c connected
to the output of a frequency doubler 220 which at its input
receives the square-wave signal R_S with the frequency f_M
furnished by the transmission frequency generator 18. The
frequency doubler supplies a square-wave signal S_R with
25 double the frequency $2 \cdot f_M$ to the signal input 218c of the
start-stop logic 218.

The start-stop logic 218 comprises a signal output 218d
connected to the counting input of a counter 222. It is
30 configured so that it is opened by a start signal supplied
to its start input 218b so that it allows the square-wave
signal S_R supplied to its signal input 218c to pass to its
signal output 218d and thus to the counter 222, and that it
is blocked by a stop signal S_S applied by the output of the
35 threshold value switch 216 to the stop input 218a so that
the further transmission of the square-wave signal S_R to the

1 counter 222 is prevented.

The stage outputs of the counter 222 are connected to corresponding inputs of the evaluating circuit 32. The
5 counter 222 counts in each case the square-wave pulses of the square-wave signal S_R occurring between a start signal and a stop signal and supplies to the evaluating circuit 32 the count indicating the number of pulses counted.

10 The operation of the arrangement of Fig. 15 will be explained with the aid of the diagrams of Fig. 16.

In Fig. 16, diagram A shows the transmission AC voltage U_S furnished by the transmission frequency generator 18, and diagram
15 B the corresponding square-wave voltage R_S of frequency f_M . Diagram C shows the square-wave voltage S_R with the double frequency $2 \cdot f_M$ furnished by the frequency doubler 220. Diagram D shows the trigger pulse I_A which is supplied to the trigger input 204a of the start circuit 204
20 and can have any random time position relative to the square-wave signal R_S . Diagram E shows the start pulse I_S generated by the start circuit 204 which starts exactly in phase with a rising edge of the square-wave signal R_S and is supplied both to the trigger input of the pulse generator 202 and to the start input 218b of the start-stop
25 circuit 218. Due to the triggering by the start pulse I_S the pulse generator 202 generates the single pulse I_D of duration T_D illustrated in diagram F (for clarity not drawn true to scale).

30

When the changeover switch 200 assumes the position illustrated in Fig. 15 and the directional changeover switch 16 is also set to the position 1 illustrated in Fig. 17, the single pulse I_D is supplied to the ultrasonic transducer 12.
35 Since a single pulse has a continuous frequency spectrum below a limit frequency which is in inverse ratio to the pulse duration, the ultrasonic transducer 12 is stimulated by the

1 single pulse I_D applied to natural resonant oscillations
at natural resonant frequencies which are contained in the
frequency spectrum of the single pulse I_D . The ultrasonic
transducer 12 thus generates a short ultrasonic wave train
5 of the type illustrated in diagram G. This ultrasonic wave
train runs through the medium flowing in the measuring tube
10 to the receiving transducer 14 at which it arrives after
a travel time T_L . The incoming wave train also stimulates
the receiving transducer 14 to natural resonant oscillations
so that it furnishes a reception AC voltage U_E in the form
of the wave train illustrated in diagram H which is supplied
to the pulse receiving circuit 210.

15 In the pulse receiving circuit 210 the reception AC voltage
 U_E is first amplified in the input amplifier 212, preferably
with low-pass filtering, and then subjected by the rectifier
circuit 214 to a full-wave rectification, resulting in
the wave train U_E' illustrated in diagram I which is app-
20 lied to the input of the threshold value switch 216. The
output signal of the threshold value switch 216 changes
from a low signal value to a high signal value when the
amplitude of its input signal exceeds for the first time
a predetermined threshold value which is set very low. As
a result, the stop signal S_S illustrated in diagram J is
25 obtained and is applied to the stop input 218a of the start-
stop circuit 218.

Diagram K of Figure 16 shows the pulse sequence which is
transferred from the output 218d of the start-stop logic
30 218 to the counting input of the counter 222. Said pulse
sequence starts in phase with the rising edge of the start
pulse I_S and also in phase with the rising edge of the
square-wave signal S_R as well as with a zero passage of the
transmitting AC voltage U_S . The counter 222 now counts the
35 number K of the pulses which are transmitted until the stop
signal S_S jumps up, i.e. during the travel time T_L . Due to
the frequency doubling this number k is twice as great as

1 the number m of the periods of the square-wave signal R_S
generated in the same interval of time by the transmission
frequency generator 18 and thus also of the transmission
AC voltage U_S . If instead of the single pulse I_D the
5 transmission AC voltage U_S had been applied to the trans-
mitting transducer 12, m periods of the transmission AC voltage
 U_S would have travelled during the travel time T_L from the
transmitting transducer to the receiving transducer so that
10 m whole wavelengths of the ultrasonic wave would have been
present in the measuring tube between the two transducers.
This is directly apparent from diagram A of Fig. 16 where m
periods of the transmission AC voltage U_S are present in the
time interval T_L . The count k reached on stopping the
counting operation thus corresponds to twice the number of
15 whole wavelengths or to the number of half wavelengths in
the measuring tube and the desired number m of the whole
wavelengths is obtained directly from the count k in acc-
ordance with the equation

20
$$m = \text{int} (k/2) \quad (30)$$

The determination of the number k of the half wavelengths
by counting with twice the frequency $2 \cdot f_M$ has the advantage
that the determination of the number m of the whole wave-
25 lengths can be checked by means of the phase measurement addition-
ally carried out. This makes it possible to detect any
error of half a wavelength in the determination of the
number of the whole wavelengths and correct it. Thus, the
transitions from m to $m+1$ or $m-1$ whole wavelengths involve
30 no problems.

For the correct determination of the number m of the whole
wavelengths the arrival of the acoustic wave train at the
receiving transducer must be detected as rapidly as possible
35 in a manner which can be reproduced as exactly as possible.
This is the purpose of the full-wave rectification in the
rectifier circuit 214. The signal U_E' of the diagram I ob-

1 tained by the full-wave rectification has a fundamental
oscillation of twice the frequency of the original reception
signal U_E of diagram H so that exceeding of the response
value of the threshold value switch 216 can be detected
5 correspondingly earlier.

10 If the ultrasonic transducers are of the type illustrated
in Fig. 13 with annular piezoelectric transducers there is
a further advantageous possibility of avoiding errors in
the determination of the number m of whole wavelengths in
the measuring tube. As explained above, the annular piezo-
electric transducers for the phase measurement are preferably
operated with the radial resonance frequency. For deter-
15 mining the number m of the whole wavelengths in the measur-
ing tube it has however been found advantageous to excite
the transmitting transducer by the single pulse in such a
manner that it oscillates not only at the radial resonant
frequency but also at the longitudinal resonant frequency
20 which is about three times as high as the radial resonant
frequency. This can be achieved by suitable dimensioning of
the duration T_D of the single pulse by making the latter
so short that the limit frequency of its frequency spectrum
lies above the longitudinal resonant frequency. When the
annular piezoelectric transducer is stimulated by such a
25 short single pulse, at the start of the oscillation firstly the
longitudinal resonance is predominant and then subsequently
the radial resonance. Due to the higher longitudinal res-
onant frequency at the start of the wave train the pulse
transmission has steeper edges so that the danger of measure-
30 ment errors is still further reduced.

Fig. 17 shows the overall block diagram of a
flow rate measuring arrangement which contains both the
circuit section for the exact phase measurement and the
35 circuit section for the determination of the number m of
the whole wavelengths in the measuring tube. The circuit

1 section for the exact phase measurement corresponds to the
example of embodiment of Fig. 9 with a differential fre-
quency generator 80. The circuit section for the determ-
ination of the number m of the whole wavelengths in the
5 measuring tube corresponds to the example of embodiment of
Fig. 15. The circuit blocks in Fig. 17 corresponding to
those of Fig. 9 or Fig. 15 are denoted by the same reference
numerals as therein and will not be described again.
To simplify the block diagram, some connections have not
10 been completely drawn but have been replaced by arrows indi-
cating the destination or origin of the respective signals.

Furthermore, the case assumed in Fig. 17 is the one in which
the evaluating circuit 32 is formed by a microcomputer (μC)
15 which also furnishes the control signals necessary for the
execution of the measuring method. To simplify the illustration,
the connections for the transmission of signals from and to
the microcomputer are not shown, either, but have been re-
placed by arrows with the designation " μC ".

20 In the flow rate measuring arrangement of Fig. 17, two add-
itional measures have been taken which have so far not
been mentioned.

25 The first measure consists in the fact that in the receiving section
of the circuit an amplitude monitoring circuit 224 is provided
to which the receiving AC voltage U_E coming from the di-
rectional changeover switch 16 is supplied. A further in-
put of the amplitude monitoring circuit 224 receives the
30 square-wave voltage Q_E from the output of the reception
signal processing circuit 20. The amplitude monitoring
circuit 224 checks whether the amplitude of the reception
AC voltage U_E is large enough to permit correct evaluation.
The square-wave voltage Q_E is supplied for controlling the
35 amplitude monitoring. If the amplitude of the receiving AC
voltage U_E drops below a predetermined value the amplitude
monitoring circuit 224 delivers a signal to the microcomputer
32 which thereupon issues a warning.

1 The second measure consists in the fact that between the transmissior
signal processing circuit 22 and the phase detector 24 an
inverter circuit 226 is inserted which is controllable by a
5 signal furnished by the microcomputer 32 in such a manner
that it either allows the square-wave voltage Q_S to pass
unchanged or inverts it. The inverter circuit 226 permits
measurement of very small residual phase angles lying in
the vicinity of 0° . With such small residual phase angles
10 the duration T_I of the measuring pulses I_M become so small
that it can no longer be determined with the steps described
above. When the microcomputer 32 detects this state it in-
itiates inversion of the square-wave voltage Q_S . This cor-
responds to a phase shift through 180° , and as a
15 result, the residual phase angle becomes 180° greater and
accordingly the duration T_I of the measuring pulses $1/(2 \cdot f_M)$
longer.

This effect will be apparent from the diagrams of Fig. 18:
Diagrams A, B and C of Fig. 1 show the square-wave voltage
20 Q_S and Q_E and the measuring pulses I_M corresponding to the
illustration in the diagrams B, C, D of Figs. 2 and 3 but
for a very small residual pahse angle so that the time in-
terval Δt and accordingly the duration T_I of the measuring
pulses I_M is very small. The diagrams D, E and F of Fig.
25 18 show the same signals for the case where the square-wave
voltage Q_S is inverted. It is apparent that in this case
the measuring pulses I_M have a considerably greater duration
 T_I , which permits easy measurement.

30 Between the reception signal processing circuit 20 and the
phase detector 24 a delay circuit 228 is inserted which
compensates the delay caused by the inverter circuit 226
so that the signals Q_S and Q_E have at the inputs of the
phase detector 24 the same relative phase position as at
35 the outputs of the processing circuits 20 and 22. Of course,
the position of the circuits 226 and 228 may be interchanged

1 so that instead of the signal Q_S the signal Q_E is inverted.

With the aid of the overall block circuit diagram the complete sequence of a measuring operation under the control of the microcomputer 32 can be explained. A complete
5 measuring operation takes place in the following phases:

1. Determination of the number of whole wavelengths

- 10 a) The microcomputer 32 brings the directional changeover switch 16 and the changeover switch 200 into the respective position 1 and sends a trigger pulse I_A to the start circuit 204.
- 15 b) Synchronously with a rising edge of the square-wave voltage R_S the start circuit 204 sends a start pulse I_S to the pulse generator 202 so that the latter emits a single pulse I_D which via the switches 200 and 16 is applied to the ultrasonic transducer 12. Simultaneously,
20 the start pulse I_S opens the start-stop logic 218 so that the counter 222 begins to count the pulses of the square-wave signal S_R .
- 25 c) When the amplitude of the amplified and rectified receiving signal U_E' reaches the response value of the threshold value switch 216 the start-stop logic 218 is blocked by the stop signal S_S so that the pulse counting in the counter 222 is terminated. The count reached is transferred to the microcomputer 32 which calculates there-
30 from the number m of the whole wavelengths in the measuring tube. Thereafter the counter 222 is reset to zero.

35 2. Measurement of the residual phase angle in the flow direction

- a) The microcomputer 32 brings the changeover switch 200

1 into the position 2 but leaves the direction changeover
switch 16 in the position 1 so that the transmission
AC voltage U_S is applied to the ultrasonic transducer
12.

5

b) After expiry of a waiting time, after which the oscilla-
tions of the transmitting transducer 12 have become stable,
the first residual phase measurement is made by pulse count-
ing during a measuring time interval of duration T_1 in the
counter 30, whereupon the count reached is transferred
to the microcomputer 32 and the counter 30 reset to
zero.

10

c) The microcomputer 32 checks whether the duration T_1 of
the measuring pulses I_M suffices for a proper calcul-
ation or whether an inversion is required. If required,
the inverting circuit 226 is reversed.

15

d) The measurement of the pulse duration T_I of the measur-
ing pulses I_M by pulse counting in the counter 30 is
repeated in n successive measuring time intervals in
which an angular phase rotation of the start phase φ_S
through 2π takes place. After each measuring time in-
terval the count of the counter 30 is transferred to the
microcomputer 32 and the counter 30 reset to zero.

20

25

e) After expiry of the n measuring time intervals the
microcomputer 32 determines from the n counts obtained
the mean residual phase angle φ_{R1} by averaging.

30

3. Measurement of the residual phase angle against the flow direction

a) The microcomputer 32 brings the directional changeover
switch 16 into the position 2 and leaves the changeover
switch 200 in the position 2 so that the transmission
AC voltage U_S is applied to the ultrasonic transducer 14.

35

- 1 b) After expiry of a waiting time, after which the oscilla-
tions of the transmitting transducer 14 have become stable,
the measurement of the pulse duration T_I of the measuring
pulses I_M by pulse counting in the counter 30 is repeated
5 in n consecutive measuring time intervals in which an angular
phase rotation of the start phase ϕ_S through 2π takes place.
After each measuring time interval the count of the
counter 30 is transferred to the microcomputer 32 and
the counter 30 reset to zero.
- 10 c) After expiry of the n measuring time intervals the
microcomputer 32 determines from the n counts obtained
the mean residual phase angle ϕ_{R2} by averaging.

15 4. Calculation of the flow velocity

The microcomputer 32 calculates from the measured values m ,
 ϕ_{R1} and ϕ_{R2} obtained and from the predetermined values for f_M and L_M
the flow velocity V_M in accordance with the equation (9) and
20 from the latter the volume flow rate by multiplication by
the flow cross-sectional area. The results are optically
displayed in suitable manner and/or converted to signals
suitable for data transmission, for example to a frequency-
proportional signal.

25 A complete measuring operation has then been performed. It
should be pointed out that the determination of the num-
ber m of whole wavelengths need not necessarily take place
at the beginning of each measuring operation because said
30 number changes only slowly, if at all. In addition, a
gradual change of the number m of whole wavelengths is also
apparent from the progressive change of the measured resid-
ual phase angle. It would therefore suffice to repeat the
measurement of the number m of whole wavelengths in relat-
35 ively large intervals of time, for example every 10 seconds.



- 1 -

C L A I M S:

1. A method for flow-rate measurement by means of ultrasonic waves which are propagated in a medium flowing through a measuring tube (10) between two ultrasonic transducers (12, 14) which are spaced a defined distance apart and one of which operates as a transmitting transducer which converts an electric transmission signal (U_S) into an ultrasonic wave of the same frequency (f_M), while the other operates as a receiving transducer which converts the incoming ultrasonic wave into an electric reception signal (U_E) of the same frequency, the flow rate being determined by measuring the phase shift between the transmission signal (U_S) and the reception signal (U_E) in such a way that in measurement time intervals whose duration (T_I) corresponds to the phase shift, those periods of a signal (I_C) of a counting frequency (f_C) are counted which are substantially greater than the measuring frequency (f_M) corresponding to the frequency of the transmission signal (U_S), that in a measurement cycle (T_R) comprising a plurality of successive measurement time intervals (T_I), the phase angle of the start phase (ϕ_S) which the counting-frequency signal (f_C) has at the beginning of each counting operation with respect to the measurement time interval (T_I) is increased from measurement time interval to measurement time



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interval, and that the counts obtained in the course of the measurement cycle (T_R) are evaluated to form a measured value with increased measurement resolution.

2. A method as claimed in claim 1 wherein the start phase (ϕ_S) of the counting-frequency signal (I_C) is changed in the course of the measurement cycle (T_R) by 2π .

3. A method as claimed in claim 1 or 2 wherein the measuring frequency (f_M) is derived from the counting frequency (f_C) by frequency division, and wherein in each measurement cycle (T_I) a phase shift (ϕ_M) is imparted to the measuring frequency (f_M) with respect to the counting frequency (f_C) which increases from period to period of the measuring frequency (f_M).

4. A method as claimed in claim 3, wherein the phase shift (ϕ_M) imparted to the measuring frequency (f_M) increases steadily in the course of the measurement cycle (T_R).

5. A method as claimed in claim 4 wherein the phase shift (ϕ_M) imparted to the measuring frequency (f_M) in each period of the measuring frequency is varied in dependence upon preceding measurement results for the purpose of successive approximation.

6. A method as claimed in claim 1 or 2 wherein the measuring frequency (f_M) is derived by frequency division from a frequency (f_C) which differs by a predetermined difference frequency (Δf) from the counting frequency (f_C).

7. A method as claimed in claim 6 wherein the difference frequency (Δf) is equal to the measuring frequency (f_M) divided by the number (n) of measurement time intervals (T_I) per measurement cycle (T_R).



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8. A method as claimed in claim 1 wherein the counting-frequency signal (I_C) is generated only for the duration (T_I) of each measurement time interval, wherein the phase of the counting-frequency signal (I_C) is stored at the end of each measurement time interval, and wherein the generation of the counting-frequency signal (I_C) is continued at the beginning of each following measurement time interval with the stored phase.

9. A method as claimed in claim 8 wherein the counting-frequency signal (I_C) is derived from a triangular voltage (U_D) which is generated only for the duration (T_I) of each measurement time interval and is continued at the beginning of each measurement time interval from the signal level reached at the end of the preceding measurement time interval.

10. A method as claimed in any one of the preceding claims wherein upon detection of a duration (T_I) of the measurement time intervals which is too short for correct measurement of the phase shift, an additional phase shift of 180° is imparted to one of the two out-of-phase signals (U_S , U_E).

11. A method as claimed in claim 10 wherein the phase shift between two square-wave signals (Q_S , Q_E) derived from the transmission signal (U_S) and the reception signal (U_E) is measured, and wherein, to obtain the additional phase shift of 180° , one of the two square-wave signals (Q_S , Q_E) is inverted.

12. A method as claimed in any one of the preceding claims wherein, to determine whole wavelengths of the



ultrasonic wave in the medium flowing through the measuring tube (10) between the two ultrasonic transducers (12, 14), the transmitting transducer is stimulated by a short pulse (I_D) to transmit a wave train, and wherein the periods of a signal (S_R) with the measuring frequency (f_M) or an integral multiple ($2 \cdot f_M$) thereof are counted during the travel time (T_L) of the wave train from the transmitting transducer to the receiving transducer.

13. A method as claimed in any one of the preceding claims wherein the amplitude of the reception signal (U_E) is monitored, and wherein upon detection of an amplitude insufficient for proper evaluation, a signal is sent to the evaluating circuit (23).

14. An arrangement comprising a measuring tube (10) through which a medium flows, two ultrasonic transducers (12, 14) on the measuring tube (10) which are spaced a defined distance (L_M) apart, devices for applying an electric transmission signal (U_S) with the measuring frequency (f_M) to one of the two ultrasonic transducers and for receiving the electric reception signal (U_E) with the measuring frequency from the other ultrasonic transducer, a counting-frequency generator (28) which generates a counting-frequency signal (I_C) with a counting frequency (f_C) which is substantially greater than the measuring frequency (f_M), a counter (30) which counts the periods of the counting-frequency signal (I_C) in measurements time intervals (T_I) corresponding to the phase shift between the transmission signal (U_S) and the reception signal (U_E), an evaluating circuit (32) to which the count reached in each measurement time interval (T_I)



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is transferred, and devices (34, 36, 38; 32, 70; 80, 34; 100, 120; 124, 126) whereby in a measurement cycle (T_R) comprising a plurality of successive measurement time intervals (T_I), the phase angle of the start phase (ϕ_S) which the counting-frequency signal (I_C) has at the beginning of each counting operation with respect to the measurement time interval (T_I) is increased from measurement time interval to measurement time interval.

15. An arrangement as claimed in claim 14 wherein the output of the counting-frequency generator (28) is connected to a frequency divider (34) whose division factor (N) is equal to the ratio between the counting frequency (f_C) and the measuring frequency (f_M), and which is followed by a controllable phase shifter (36) which imparts to the output of the frequency divider (34) a phase shift determined by a phase control signal (U_R), and wherein the output of the controllable phase shifter (36) is used for the electric transmission signal (U_S).

16. An arrangement as claimed in claim 15 wherein the phase control signal (U_R) is a ramp signal which changes linearly in the course of each measurement cycle.

17. An arrangement as claimed in claim 15, comprising a digital-to-analog converter (70) which converts digital code groups into the phase control signal (U_R).

18. An arrangement as claimed in any one of claims 15 to 17 wherein the controllable phase shifter (36) is formed by a PLL circuit.



23. An arrangement as claimed in any one of claims 14 to 17, comprising a pulse generator (202) which, to determine the number (m) of whole wavelengths of the ultrasonic wave in the measuring tube (10), applies a short single pulse (I_D) to one of the two ultrasonic transducers (12, 14), so that the latter is stimulated to transmit a short wave train, a counter (222) which is started upon application of the single pulse (I_D) to count a counting signal (S_R) with the transmit frequency (f_M) or an integral multiple ($2 \cdot f_M$) thereof, and a pulse-receiving circuit (210) which is connected to the other ultrasonic transducer and stops the counter (222) upon arrival of the short wave train at the other ultrasonic transducer.

24. An arrangement as claimed in claim 23 wherein the pulse-receiving circuit (210) contains an input amplifier (212), a rectifier circuit (214), and a threshold switch (216) in series.

25. An arrangement as claimed in claim 23 or 24 wherein the counting signal (S_R) is applied to the counter (222) through a start-stop logic (218) which is enabled by a start signal (I_S) initiating the generation of the short single pulse (I_D) and is inhibited by a stop signal (S_S) provided at the output of the pulse-receiving circuit (210).

26. An arrangement as claimed in any one of claims 23 to 25, comprising a frequency doubler (220) which receives a square-wave signal (R_S) with the transmit frequency (f_M) and delivers the counting signal (S_R) with twice the transmit frequency ($2 \cdot f_M$).



27. An arrangement as claimed in any one of claims 14 to 26, comprising signal-conditioning circuits (20, 22) which convert the reception signal (U_E) and the transmission signal (U_S) into square-wave voltages (Q_E , Q_S) in phase therewith, and a phase detector (24) which receives the square-wave voltages (Q_E , Q_S) and provides measuring pulses (I_M) whose duration (T_I) corresponds to the phase shift between the input signals of the phase detector (24) and determines the duration of the measurement time intervals.

28. An arrangement as claimed in claim 27 wherein between the output of one of the two signal-conditioning circuits (20, 22) and the associated input of the phase detector (24), an inverter (226) is inserted which is operated by a control signal provided by the evaluating circuit (32) to invert the applied square-wave voltage (Q_S or Q_E) when the evaluating circuit (32) determines that the duration (T_I) of the measuring pulses (I_M) provided by the phase detector (24) is less than a predetermined value.

29. An arrangement as claimed in any one of claims 14 to 28, comprising an amplitude-monitoring circuit (224) which receives the AC reception voltage (U_E), and supplies a signal to the evaluating circuit (32) when the amplitude of the AC reception voltage (U_E) is less than a predetermined value.

30. An arrangement as claimed in any one of claims 14 to 29 wherein each of the ultrasonic transducers (12, 14) is an annular piezoelectric transducer (130) which is



installed coaxially with the measuring tube (10) and, to measure the phase shift, is stimulated by a measuring frequency (f_M) equal to its radial resonance frequency.

31. An arrangement as claimed in claim 30 wherein the radial resonance frequency of the annular piezoelectric transducer (130) is so chosen that the wavelength of the ultrasonic wave propagated in the medium is greater than 1.6 times the internal diameter of the measuring tube (10).

32. An arrangement as claimed in claim 30 or 31 wherein the acoustic characteristic impedance of the measuring tube (10) is substantially greater than the acoustic characteristic impedance of the medium flowing therein.

33. An arrangement as claimed in any one of claims 30 to 32 wherein, to determine the number of whole wavelengths in the measuring tube (10), the annular piezoelectric transducer (130) is excited by a short single pulse (I_D) into vibrations at its longitudinal resonance frequency.

34. An arrangement as claimed in any one of claims 30 to 33 wherein the inner surface of the annular piezoelectric transducer (130) is covered with a protective layer (138) resistant to aggressive media.

35. An arrangement as claimed in any one of claims 30 to 34 wherein between the annular piezoelectric transducer (130) and the adjacent tube sections, seals (150, 152) are inserted whose acoustic characteristic impedance is substantially lower than the acoustic characteristic impedance of the tube sections.



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36. An arrangement as claimed in any one of claims 30 to 35 wherein on both sides of the measuring tube (10), attenuator tubes (160) are mounted whose inner surfaces are lined with a nonporous damping material (164).