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A LOW-DISTORTION ACOUSTIC-MEASUREMENT OSCILLATOR USING SEMICONDUCTOR JUNCTIONS AS VARIABLE-TUNING ELEMENTS

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A LOW-DISTORTION ACOUSTIC-MEASUREMENT OSCILLATOR USING SEMICONDUCTOR JUNCTIONS AS VARIABLE-TUNING ELEMENTS

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Abstract: This oscillator is of neither the function-generator nor the beat-frequency type. It is basically an RC oscillator, in which the R's are formed by transistor junctions, whose d.c. bias voltage is varied to change the frequency. The inherently exponential current/voltage relationship for such junctions results in an accurately logarithmic scale shape.

The provision of accurate compensation for the effects on frequency of variations in ambient temperature is of vital importance, but a highly satisfactory result is achieved.

The circuit lends itself to the provision of auto-sweep and warble facilities, and is currently in use by two leading British firms for production testing of loudspeakers and microphones respectively. The waveform purity - in particular the absence of high-order harmonics - is an attractive feature for such purposes.

The circuit inherently provides two equal-amplitude outputs in phase-quadrature, which lends itself to the operation of an economical type of tracking-filter system.

Introduction

The oscillator described in this paper is not of the function-generator type. It is essentially an R-C oscillator operating under linear, low-level, conditions, in which the R's of the R-C frequency-determining circuit are replaced by diode-connected transistors, or "transdiodes". Tuning is accomplished by varying the d.c. bias voltage applied to these transdiodes, thereby changing their a.c. conductance and hence the frequency. An accurately logarithmic scale shape is inherently obtained.

Circuit Design Considerations

Under the operating conditions here employed, transistors connected as in Fig. 1 follow very accurately the law represented by equation (1):-

$$I = I_0 e^{\frac{qV}{kT}} \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

where: I_0 = theoretical reverse leakage current
q = charge of an electron
k = Boltzmann's constant
T = absolute temperature

Differentiating (1) gives:-

$$\frac{dI}{dV} = \frac{q}{kT} I_o e^{\frac{qV}{kT}} \quad . \quad . \qquad (2)$$

$\frac{dI}{dV}$ is the incremental or a.c. conductance of the transdiode, which will be denoted by G_t since it is used as a tuning element. It thus follows that:-

$$G_t = \frac{q}{kT} I_o e^{\frac{qV_{dc}}{kT}} \quad . \quad . \qquad (3)$$

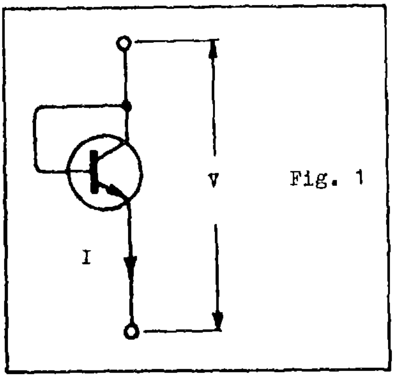


Fig. 1

where V_{dc} is the d.c. bias voltage applied to the transdiode. Equation (3) may alternatively be written:-

$$\log_e \frac{G_t \, kT}{I_o q} = \frac{qV_{dc}}{kT} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \qquad (4)$$

In any R-C oscillator with constant capacitance values in the frequency-determining network, the frequency is directly proportional to the conductance of the resistive elements in this network, provided all the conductances are varied together. It therefore follows, from equation (4), that if each of the frequency-determining resistive elements is replaced by a transdiode, or combination of transdiodes, then, provided the amplitude of oscillation is sufficiently small, variation of the transdiode d.c. bias voltage will result in a logarithmic variation of frequency with voltage. Thus a logarithmic scale shape may be obtained by employing a linear potentiometer to vary the bias voltage.

The current/voltage characteristic of a single transdiode is, of course, highly asymmetrical, but by using the tuning transdiodes in pairs in a suitable manner, a resultant symmetrical characteristic is obtained, and this allows the a.c. signal level in the oscillator circuit to be made considerably higher for a given magnitude of total harmonic distortion, thus improving the signal-to-noise ratio of the output.

The basic oscillator circuit adopted is that shown in Fig. 2, which has the advantage for the present application that both tuning conductances are earthed on one side, and that both operate at the same signal level. The oscillator loop consists of two unity-gain all-pass stages, each contributing 90° phase

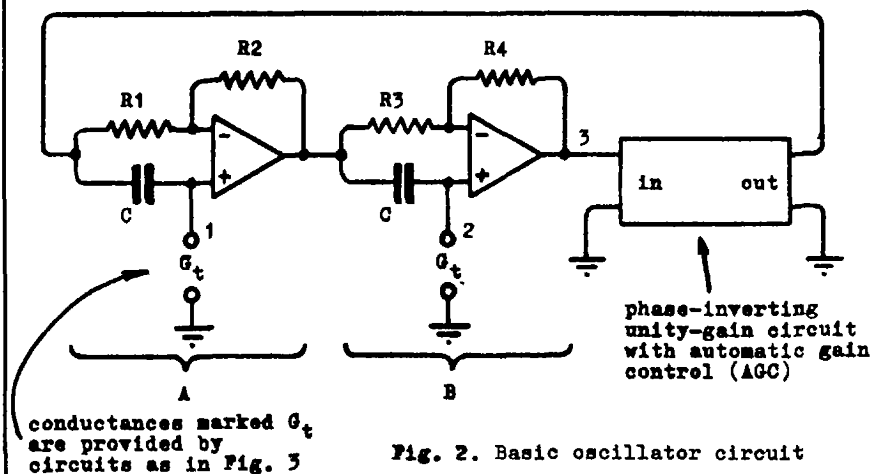
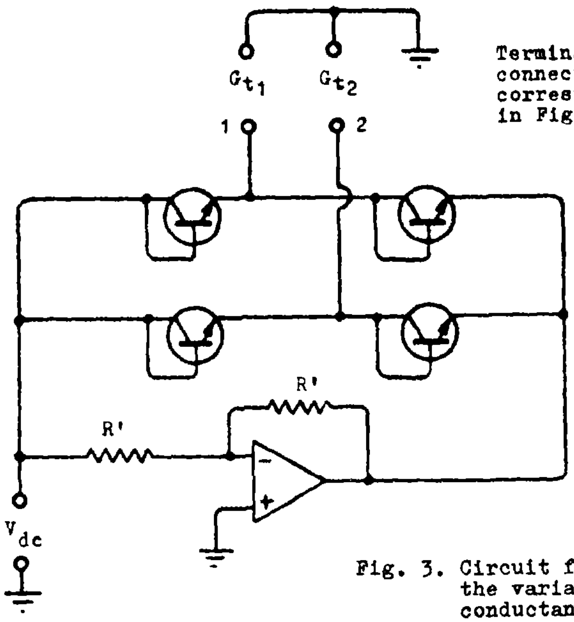


Fig. 2. Basic oscillator circuit

shift, plus a unity-gain phase-inverting circuit with FET automatic-gain-control.

The arrangement used for providing the logarithmically-variable conductances is shown in simplified form in Fig. 3. In addition to providing conductances each having a symmetrical current/voltage characteristic, as mentioned earlier, this circuit also has the virtue that variation of the tuning voltage, V_{dc}, ideally causes no changes at all in the d.c. voltage level at the terminals '1' and '2', which remain, indeed, at earth potential. This is important, for the introduction of d.c. level disturbances into the oscillator circuit of Fig. 2 is liable to lead to tuning bounce.

In the practical oscillator, the tuning arrangement shown in Fig. 3 has been somewhat elaborated. Firstly, a total of eight transdiodes is used, which permits the a.c. voltage level to be doubled for the same percentage distortion. This is considered to represent a sensible compromise between economy and excellence of signal-to-noise ratio. Secondly, a preset adjustment is provided to enable the two conductances, G_{t_1} and G_{t_2}, to be set to equality - accurate equality is unlikely to occur without such an adjustment because of production variations in the transistor characteristics. However, if set to equality at one conductance value, the two conductances remain very accurately equal as the



Terminals 1 and 2 are connected to the corresponding points in Fig. 2.

Fig. 3. Circuit for providing the variable tuning conductance.

frequency is varied. Without such equality, neither of the all-pass stages produces 90^o phase shift. (The availability of two equal-amplitude voltages in phase quadrature from the outputs of the operational amplifiers in Fig. 2 lends itself to the operation of an economical type of tracking filter employing quadrature-referenced balanced modulators and a base-band filter.) Thirdly, a desirable addition to the Fig. 3 circuit is a pair of electrolytic capacitors to earth from the extremities of the resistors labelled R', so as to avoid adding significant impedance in series with the transdiode conductances at high audio frequencies. Considerable care must be taken in the design of the operational amplifier feedback compensation arrangements if a sufficiently low output impedance is to be maintained at all working frequencies in the presence of these capacitors, for resonance can occur between the capacitance and the inductive output impedance of the operational amplifier circuit itself.

A major consideration in the design of oscillators of the present type arises from the fact that the conductance G_t as given by equation (3) is highly temperature-dependent. Temperature actually enters this equation in three places - the two shown explicitly, and also because I_0 is highly temperature-dependent. Indeed the latter effect is very much the dominant one, since I_0 approximately doubles for every 4°C temperature rise. While it would be possible to obtain satisfactory frequency stability by stabilizing the temperature of the tuning transdiodes to within a small fraction of 1°C by means of an oven, this approach would result in a fairly long warm-up time after switching on. Moreover the extra power consumption would be an unattractive feature in battery-operated versions of the oscillator. Consequently the alternative approach of using accurate temperature-compensation techniques has been preferred.

Because of the simple and fundamental nature of the basic transdiode equation (1), and the closely-matched variation of I_0 with temperature in good modern transistors of the same type and batch, it is possible to obtain very accurate temperature-compensation of frequency by using compensating transistors of the same type as those used for the tuning conductances.

With reference to Fig. 3, V_{dc} has one component (the largest component) equal to the voltage drop across a separate temperature-compensating transdiode operated at a constant, stabilized, value of direct current. If this were the only component of V_{dc}, then, with matched transistors, the direct current in the tuning transdiodes would be held constant and equal to that flowing in the compensating transdiode independently of temperature. The oscillator frequency would then be rendered independent of the large variation in I_0 caused by ambient temperature changes, but the temperature compensation would still be incomplete. This may be understood by deriving from equations (1) and (3) the relationship:-

$$G_t \approx \frac{qI_{dc}}{kT} \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

It is now evident that even though I_{dc} may have been made perfectly independent of ambient temperature in the above way, the factor kT in (5) will result in there still being some variation in frequency with temperature. Further analysis shows that if a small temperature-independent d.c. voltage of the correct magnitude is added to that provided by the compensating transdiode, then the temperature coefficient of frequency may be reduced to zero.

Having thus rendered the temperature coefficient zero at one frequency, it is necessary to arrange for it to remain zero when the frequency is shifted to other values. This may be achieved if the third component of V_{dc}, applied to shift the frequency, e.g. by means of a dial-operated linear potentiometer, is made suitably temperature-dependent by employing a further temperature-compensating transistor.

In the practical design, where, as already mentioned, eight tuning transdiodes are used, instead of the four shown in Fig. 3, the compensation arrangements must be amended appropriately. Also, because of the critical nature of the compensation required for the variation of I_0 with temperature, the tuning and compensating transistors are coupled tightly together from a thermal point of view, by means of a common heat sink. Without this precaution, the frequency is liable to be too dependent on temperature gradients within the instrument. However, by adopting the correct, though quite straightforward, techniques in all the above respects, a highly satisfactory performance is consistently obtained.

Though the simple theory for this type of oscillator predicts a perfectly logarithmic frequency variation for a linear variation in tuning voltage, it is found in practice that the frequency becomes slightly less than the ideal value at the top end of the tuning range, especially if the latter extends, say, an octave above the top of the audio band. The causes of this departure from the ideal law are (a) series resistance within the tuning transdiodes and (b) the presence of unwanted slight phase lags in the operational amplifier circuits used in the oscillator loop. By good design these effects can be reduced to a minimum, but some small departure from the ideal tuning law will nevertheless remain. This error may be very satisfactorily and quite economically removed by feeding the oscillator output to a simple diode-pump frequency-discriminator circuit, whose d.c. output is arranged to augment the d.c. tuning voltage progressively as the frequency rises, producing a significant correction only near the top end of the tuning range.

The oscillation amplitude is stabilized by having a peak rectifier on the oscillator output, in association with an FET as the variable-gain element. This is much more satisfactory than a thermistor, partly because the FET can operate at the same small a.c. signal level, of about 15mV rms, as the rest of the basic oscillator circuit. Though the use of a sample-and-hold type of rectifier was considered during the design, in the event a straightforward peak rectifier with suitably long time-constant was found entirely satisfactory.

For negligible tuning bounce, it is important that the variation in loop gain with frequency, in the absense of AGC, should be made as small as possible by using good operational amplifiers with optimum frequency compensation. The tendency is for the gain to rise slightly at the high-frequency end of the tuning range. This happens because the dominant-lag compensation gives the amplifiers an inductive output impedance, so that the capacitive loading of this by the input impedance of the all-pass networks causes a slight amplitude rise. Adding emitter followers after each i.c. operational amplifier, within the local feedback loop, reduces this effect, but it is also advantageous to provide a facility in the phase-inverting stage of the complete oscillator (see Fig. 2) whereby a small and adjustable amount of high-frequency gain roll-off can be introduced. In the absence of such precautions, unwanted supersonic oscillation may occur, for the all-pass networks ideally provide unity gain up to indefinitely high frequencies.

The distortion introduced by the tuning transdiodes is mainly third harmonic, though at very low frequencies a small amount of second harmonic is caused by the flowing into these transdiodes of the bias current of the all-pass stage operational amplifiers, but the use of low-bias-current types minimises this effect. The third-harmonic distortion, however, is of predictable magnitude, and a worthwhile reduction in the total output distortion is achieved by introducing, elsewhere in the circuit, an appropriate amount of approximately antiphase third-harmonic distortion.

Practical Aspects

The main oscillator circuit, as outlined above, is built on a printed-circuit board measuring 120 X 110 mm, and requires ±9 to ±15V unregulated, and ±5.6V regulated, supplies. The original prototype instrument is housed in a standard instrument case, with a panel 245 X 135 mm, and is battery-operated. An alternative mains unit can be inserted in place of the batteries, and operates on any mains voltage from 100V to 250V without requiring resetting. The instrument incorporates a Blumlein integrator to give an auto-sweep facility, together with circuits to provide frequency markers and warble-tone.

Performance Specification

Experience with about twelve instruments has shown that the circuit is straightforward to test and set up, provided the right logical procedure is followed. It is believed that the following performance specification could be met in larger-scale production. Regular use by two leading British firms for production testing of loudspeakers and microphones respectively has shown that excellent and trouble-free performance is obtained from the design.

Frequency Coverage The main tuning dial has a logarithmic scale calibrated from 20Hz to 20kHz. A fine-tuning knob provides a variation of ± 5% at all settings of the main dial. Using the internal auto-sweep facility, or by applying an externally-generated ramp voltage, the frequency may be swept from 10Hz to 40kHz.

Frequency Accuracy Provided the instrument is correctly set up against the internal frequency markers, using the front panel preset adjustments, frequency errors will not exceed ± 3% at any dial setting. In practice, these adjustments very seldom require to be re-made.

Frequency Stability A warm-up frequency drift of approximately - 0.8% occurs during the first ten minutes. The variation in frequency with ambient temperature does not exceed ± 0.2% per °C for the slow temperature variations met in normal practical use. On battery operation, variation from ±7V to ±10V gives less than ±0.3% frequency variation at any dial setting. On mains operation, the supply voltage may be varied from 100V rms to 250V rms without appreciable effect on frequency.

Amplitude Stability The variation of output voltage with ambient temperature does not exceed ± 0.25%/°C, or ± 0.02dB/°C. The variation of output voltage with frequency does not exceed ± 0.1dB from 20Hz to 40kHz. The frequency may be swept over the whole dial range in times down to 2 seconds without any bounce in the output amplitude becoming visible on a CRO. Unwanted amplitude modulation at full-deviation warble does not exceed a modulation depth of 1% at any dial setting.

Distortion The total harmonic distortion, at any output level, and with any value of resistive load down to zero, is less than 0.05% from 40Hz to 10kHz, and less than 0.1% from 20Hz to 40kHz. The distortion is mainly 2nd. and 3rd. harmonic.

Noise The unweighted random noise voltage, measured over a 20kHz noise bandwidth, which is superimposed on the output sine-wave voltage, is approximately 70dB below the level of the latter.

Markers Markers are provided at 100Hz, 1kHz and 10kHz, with a frequency accuracy of ±1%. With the marker button pressed, the markers appear as pips on the output waveform as the frequency sweeps through the corresponding frequencies. The markers are also available on pins on an auxiliary output socket. One of these pins provides a unidirectional marker of approximately 2V amplitude; if the oscillator frequency is held constant at a marker frequency, this pin gives out a positive d.c. voltage of 2V. Another output pin provides a bidirectional marker waveform having a fundamental frequency equal to the oscillator frequency and an excursion of approximately ±2V.

Auto-Sweep A logarithmic frequency sweep may be initiated, from any frequency to which the dial is set, by operating the "sweep/dial" switch. A further switch selects either an ascending or a descending sweep, at either a slow or a fast rate. The fast and slow sweep times are adjustable by screwdriver presets on the front panel, which cover ranges, for three decades of frequency, of at least 11 to 22 seconds and 45 to 90 seconds respectively. The autosweep range extends from just under 10Hz to just over 40kHz. Coincidently with the start of the autosweep, a narrow 5 volt positive-going pulse is provided from an output socket.

External Sweep The frequency may be swept logarithmically by applying an externally-generated ramp voltage to the appropriate pin. A positive-going ramp gives a rising frequency, with a sensitivity of 0.9 to 1.1 V/decade.

Auto-Sweep Ramp Output The internally-generated autosweep ramp voltage is made available at the auxiliary output socket, and is negative-going, 0.9 to 1.1V/decade, for rising frequency. Operation of the tuning dial does not provide an output frequency-dependent voltage.

Warble The warble frequency is approximately 8Hz and the frequency-modulation waveform is triangular. The percentage frequency deviation is independent of the sine-wave frequency to which the oscillator is set, and is adjustable from zero to a maximum peak-to-peak deviation of $\frac{1}{3}$ octave (approx. $\pm$ 13%).

Attenuator Etc. The internal resistance looking into the output socket is $600\Omega \pm 2\%$ at all settings of the coarse and fine attenuators. The source voltage is nominally 1.0V rms at 0dB, and the maximum attenuation is 60dB.

Quadrature Outputs The departure from 90° relationship of the two 15mV rms quadrature outputs does not exceed $\pm 2^{\circ}$ from 20Hz to 10kHz, and is not greater than $\pm 4^{\circ}$ at 20kHz.

Acknowledgements Etc.

Before developing the present oscillator, I had been using the fig. 2 configuration to produce warble-tone by means of an active modulator-controlled variable tuning resistance. The suggestion to use semiconductor-diode tuning in this circuit for the purpose of obtaining a logarithmic frequency scale, was made by C. J. V. Moore, of KEF Electronics Ltd., after reading a "Circuit Idea" from Kamil Kraus in Wireless World, August 1974 - which did not, however, include the logarithmic-law feature. The first quick experimental model giving an approximately logarithmic scale was made by Mr. Moore, who had in fact thought of such possibilities before seeing the above Wireless World, which triggered him into action.

No arrangements have yet been made for marketing the design. British patent application no. 07320 of 1977 (Baxandall and Moore) covers the features described in the present paper.