

# DESIGN OF CRYSTAL OSCILLATOR CIRCUITS

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## 1. PRINCIPLE CONSIDERATIONS

### 1.1. Crystal Oscillators Using Low-Frequency Crystals

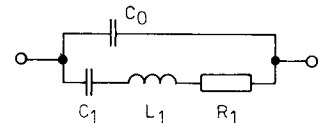
All flexure, extensional and face-shear mode vibrators, whose resonant frequency is typically less than 1 MHz are to be designated here as LF-crystals. This means that thickness-shear vibrators will not fall under this definition. Due to their sensitivity to mechanical shock, their large frequency-to-temperature characteristic and their higher price due to the complicated manufacture, the importance of this type of vibrator has dropped off considerably. The possibility of obtaining low frequencies with the aid of integrated (CMOS) dividers in conjunction with thickness-shear vibrators (AT-crystals) has also had its effect.

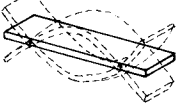
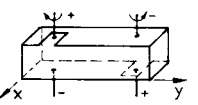
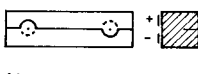
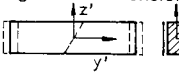
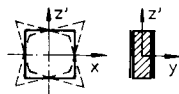
**Table 1** gives a list of the most common types of vibrators in this frequency range. The temperature response of the resonant frequency is either straight or in the form of a second order parabola

$$\frac{\Delta f}{f} = -a(T - T_{inv})^2 \quad (1)$$

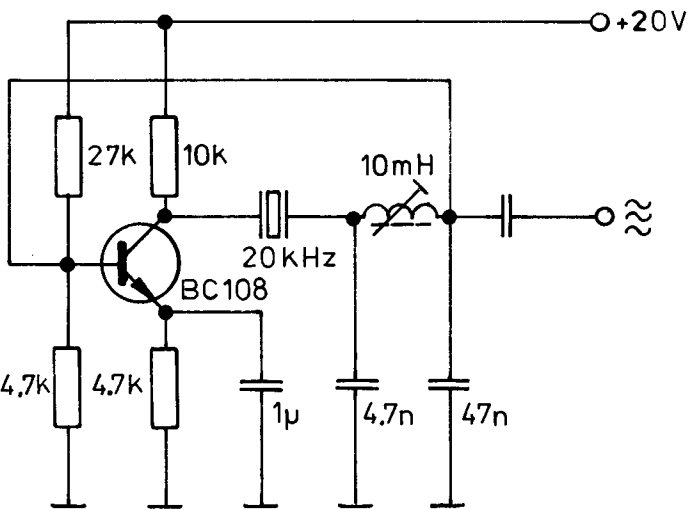
with a coefficient  $a$  of between  $2 \times 10^{-8}/\text{deg}^2$  and  $5 \times 10^{-8}/\text{deg}^2$  ( $\frac{\Delta f}{f}$  in ppm  $\triangleq 10^{-6}$ ) according to type. The inversion point  $T_{inv}$  can be varied for each vibrator by changing the design of the crystal.

**Table 1: List of the most common non-AT crystals in the range < 1 MHz**



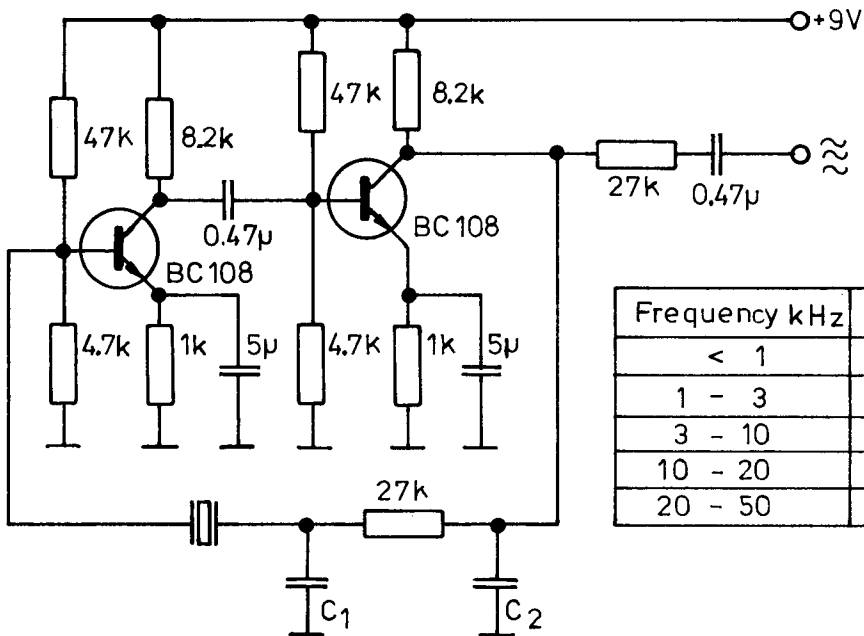
Frequency Range	Vibrator Type	Mode of Oscillation	Temperature Response	$R_1$ , [k $\Omega$ ]	$C_1$ , [fF]
800 Hz – 4 kHz	Duplex flexure vibrator	Bimetal principle  $f \sim \frac{b}{12}$	Parabola $T_{inv} = 15^\circ\text{C}$	750 – 250	250 – 50
4 kHz – 15 kHz	X-Y flexure vibrator	 $f \sim \frac{b}{12}$	Parabola $T_{inv} = + 10^\circ$ to $40^\circ\text{C}$	200 – 80	50 – 15
15 kHz – 50 kHz	H-flexure vibrator	 Also with three poles $f \sim \frac{b}{12}$	Straight – 7 to – 15 ppm/deg	20 – 8	35 – 20
50 kHz – 200 kHz	X + 5° extensional vibrator (e.g. 100 kHz standard crystal)	Longitudinal extension  $f \sim \frac{1}{T}$	Parabola (16° to 55°C)	4	60 – 30
200 kHz – 400 kHz 300 kHz – 800 kHz 000 kHz – 800 kHz	Face-shear vibrator DT-cut CT-cut SL-cut	 SL: rectangular $f \sim \frac{1}{T}$	Parabola, $T_{inv} = 0$ to $60^\circ\text{C}$ typ. $15^\circ\text{C}$	1– 5	30 – 7

The resonance resistance  $R_1$  is the most important magnitude for designing crystal oscillators. This resistance at series resonance is in the order of about 1 M $\Omega$  down to 1 k $\Omega$ , which means that it varies according to the cut and frequency range by a factor of nearly 1000. This means that it is not possible to provide a standard crystal oscillator circuit for the whole range.



**Fig. 1:**  
20 kHz crystal oscillator with impedance transformation

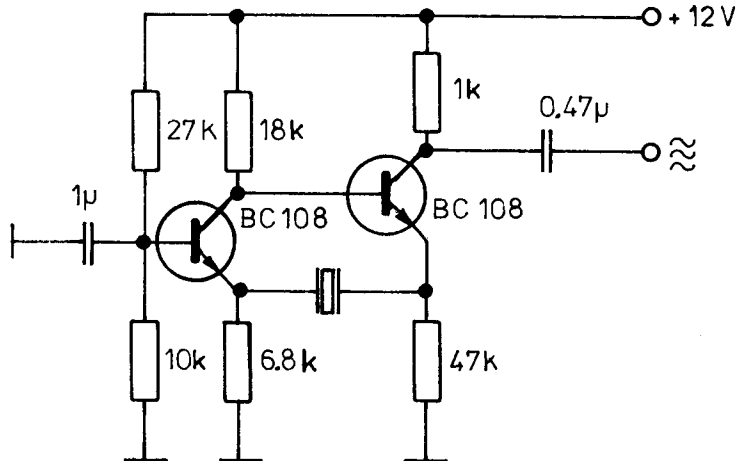
At large values of  $R_1$ , it is necessary for the amplifier stage to possess a high input impedance. This can be achieved, as shown in **Figure 1**, by using impedance transformation (1). If this is not the case, it is necessary for the voltage-divider loss (crystal –  $R_1$  : Input impedance) to be compensated for using a high loop gain, e.g. by using a two-stage amplifier as shown in **Figure 2** (2). A RC-lowpass filter is connected in the feedback link in series with the crystal. The cut-off frequency of this filter is somewhat higher than the crystal frequency. This ensures that no tendency for excitation of parasitic modes can take place, which is very prevalent with LF-crystals.



Frequency kHz	C <sub>1</sub>	C <sub>2</sub>
< 1	2,2n	15n
1 - 3	0,68n	6,8n
3 - 10	-	4,7n
10 - 20	-	0,68n
20 - 50	-	-

**Fig. 2:** Two-stage series-resonance oscillator for 0.8 to 50 kHz

The Butler-circuit as shown in **Figure 3** has proved itself well for frequencies in excess of approximately 50 kHz. If the loop gain is not sufficient, the collector resistor should be replaced by a choke or resonant circuit.



**Fig. 3:**  
Buffer-oscillator  
for 50 to 500 kHz

## 1.2. Crystal Oscillators Equipped with Fundamental Mode AT-Cut Crystals

The most popular crystal cut is the AT-cut. AT-crystals are thickness-shear vibrators. They cover a fundamental frequency range from approximately 750 kHz to 20 MHz (and some exceptions from 500 kHz to 30 MHz).

**Table 2** gives the frequency ranges of the various different crystal shapes together with their typical equivalent data ( $1 \text{ fF} = 10^{-3} \text{ pF}$ ) which are necessary for physical reasons. The temperature response is a third order parabola whose form can be influenced by selection of the cutting angle. It is given in ppm.

$$\frac{\Delta f}{f} = a_1 (T - T_{\text{inv}}) + a_3 (T - T_{\text{inv}})^3 \quad (2)$$

where the coefficients are:

$$a_1 \approx -0.084 \times \Delta\varphi$$

$$a_3 \approx 10^{-4}$$

The inversion temperature  $T_{\text{inv}}$  is in the order of 22 to 33°C according to the range.  $\Delta\varphi = \varphi_0 - \varphi$  is the angular difference (in minutes of arc) to the so-called zero TC angle  $\varphi_0$  (at this cut angle the temperature coefficient will be zero at the inversion point).

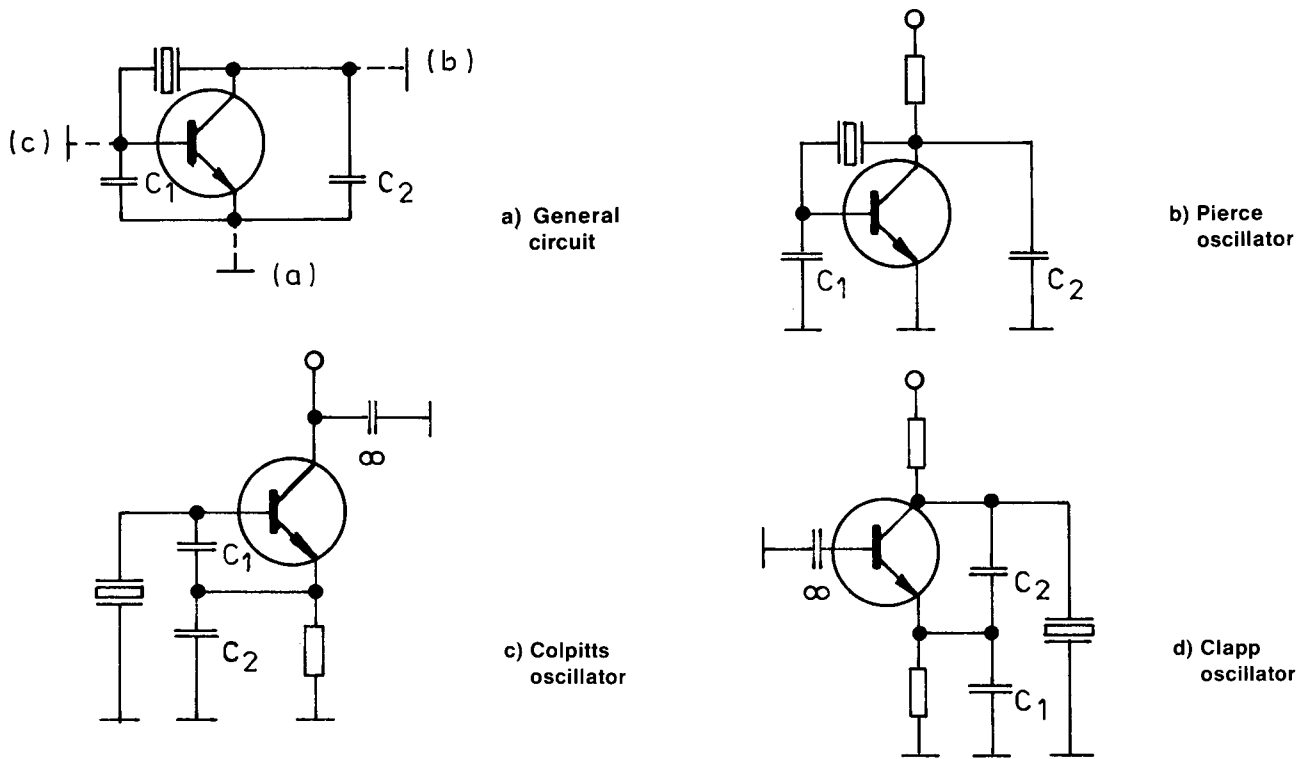
As will be seen in Table 2, the typical resonance resistance  $R_1$  will be between 10 and 500  $\Omega$  and decreases with increasing frequency.

**Table 2: Equivalent Data of AT-fundamental crystals**



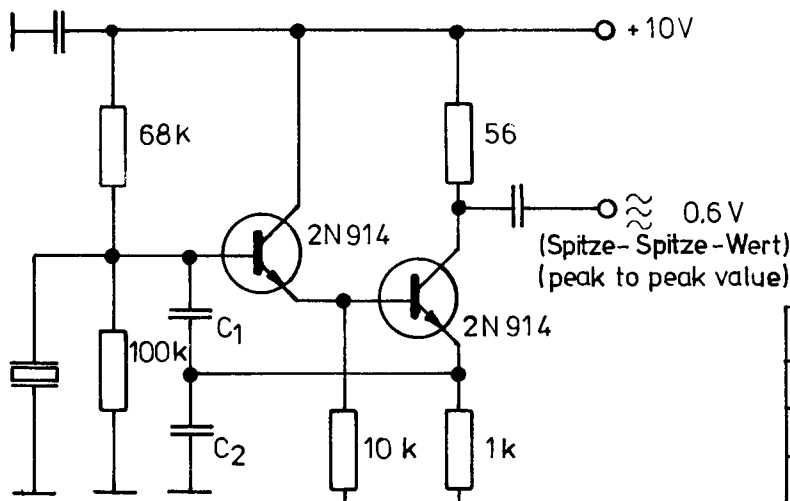
Shape of Crystal	Frequency range (MHz) for cases			Typical Equivalent Data			
	HC-6/U	HC-25/U	HC-35/HC-45	$C_0$	$C_1$	Q	$R_1$
Biconvex	0.75 - 1.5	—	—	3 - 7 pF	8 fF	> 100 000	100 $\Omega$ - 500 $\Omega$
Planoconvex	1.5 - 3	2.7 - 5.2	—	4 - 7 pF	10 fF	> 100 000	< 200 $\Omega$
Planoparallel with bevel	2 - 7	4.5 - 10.5	10 - 13	5 - 7 pF	20 fF [10 fF]	> 50 000	10 $\Omega$ - 100 $\Omega$
Plane	7 - 20 (30)	10.5 - 20(30)	13 - 20 (30)				

The most common type of circuit for fundamental AT-crystals are aperiodic oscillators, that is oscillators without additional selectivity. The most important types of parallel-resonant oscillators are Pierce, Colpitts and Clapp oscillators that can be derived from a circuit by varying the ground point (**Figure 4**). The crystal operates at a point where it exhibits the same characteristic as a high Q inductance.



**Fig. 4: Parallel-resonance oscillators for fundamental crystals (RF equivalent diagram)**

A circuit equipped with a Darlington stage is shown in **Figure 5** as an example of the successful Colpitts oscillator. Due to the high input impedance, it is possible for the divider capacitors  $C_1$  and  $C_2$  to possess large capacitance values. This means that the reaction of the transistor stage on the oscillator frequency is very low. The effective load capacitance of the crystal is represented by the series connection of  $C_1$  and  $C_2$ . In order to obtain suitable standard values of approximately 30 pF (typical range 10 pF to 50 pF), an additional capacitance of this order should be placed in series with the crystal in practice in order to align the crystal frequency.



**Fig. 5: Colpitts oscillator with Darlington stage suitable for fundamental crystals**

Frequency	$C_1$	$C_2$
3 – 6 MHz	560 pF	470 pF
6 – 15 MHz	560 pF	220 pF
15 – 30 MHz	220 pF	100 pF

A disadvantage of aperiodic oscillator circuits is the tendency to oscillate at the third or higher overtone of the crystal, or to a non-harmonic spurious resonance. In difficult cases, capacitance  $C_2$  should be replaced by a resonant circuit, which is detuned so that it is capacitive at the nominal frequency (principle of the Tritet oscillator).

Generally speaking, the positive feedback should not be greater than required for starting and maintaining stable oscillation. In the case of the Colpitts circuit, the values of  $C_1$  and  $C_2$  can be taken from the following equations:

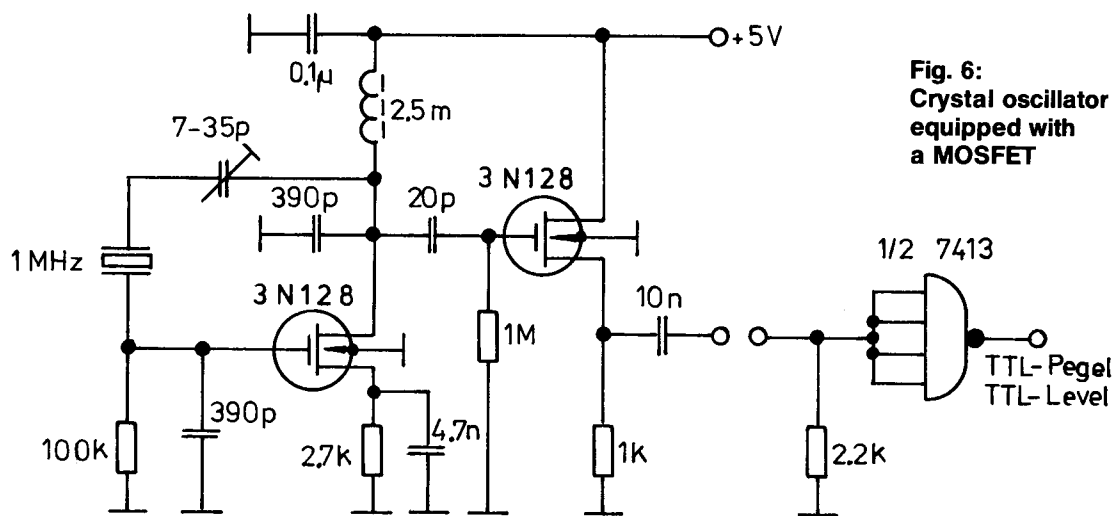
$$\frac{C_1}{C_2} = \sqrt{\frac{r_{be}}{r_a}} \quad (3)$$

$$C_1 \cdot C_2 = \frac{g'_m}{\omega_0^2 R'_1} \quad (4)$$

where:

- $r_{be}$  is the (RF) impedance between base and emitter (of the Darlington)
- $r_a$  is the AC-output impedance (measured at the common emitter)
- $g'_m$  is the transconductance ( $= \frac{1}{R_{in}}$  with an emitter follower)
- $R_1$  is the resonant resistance of the crystal transformed by the load capacitance (see 4.1.2.1., equ. 17)

**Figure 6** gives an example of a Pierce oscillator for 1 MHz equipped with a MOSFET (4). A TTL output level is available if the output of the crystal oscillator drives a Schmitt-Trigger (7413). Such an oscillator is suitable for being used as a clock for frequency counters.



**Fig. 6:**  
Crystal oscillator  
equipped with  
a MOSFET

### 1.3. Crystal Oscillators with Overtone AT-Crystals

If a thickness-shear vibrator is excited at an overtone, the crystal disk will oscillate in several subdisks in anti-phase (see illustration at Table 3). Only odd overtones can be excited. The fundamental frequency of an AT-crystal is inversely proportional to the thickness of the disk. For instance, a fundamental crystal for 30 MHz will have a thickness of approximately 55 μm. If this crystal is now excited at the third overtone, e.g. at 90 MHz, the electrical effective sub-disk thickness will be a third, which amounts to approx. 18 μm.

However, the overtone frequency is not exactly a multiple of the fundamental mode frequency; but this so-called anharmonicity will become less and less with the higher order overtones. For this reason, it is relatively simple to operate crystal oscillators even up to frequencies in the order of 300 MHz, although the usual upper frequency limit is 200 MHz/ninth overtone; one can operate the crystal at the eleventh or thirteenth overtone, which is virtually exactly 11/9 or 13/9 times the ninth overtone. However, a crystal with the highest possible fundamental-frequency should be selected (20 MHz to 30 MHz), so that the overtone modes are spaced far from another.

**Table 3: Equivalent data of AT-overtone crystals**



Over-tone	Frequency range (MHz) for cases			Typical Equivalent Data			
	HC-6/U	HC-25/U	HC-35/HC-45	$C_0$	$C_1$	Q	$R_1$
3	18 – 60 (80)	20 – 60 (90)	27 – 60 (90)	5 – 7 pF [2 – 4 pF]	2 fF [1 fF]	$> \frac{4 \times 10^6}{f \text{ (MHz)}}$	20 $\Omega$ [ 40 $\Omega$ ]
5	40 – 115(130)	40 – 115(150)	50 – 125		0.6–0.8 fF[0.4 fF]	$> \frac{5 \times 10^6}{f \text{ (MHz)}}$	40 $\Omega$ [ 80 $\Omega$ ]
7	70 – 150	70 – 150	70 – 175		0.3 – 0.4fF[0.2 fF]		100 $\Omega$ [150 $\Omega$ ]
9	150 – 200	150 – 200	150 – 200		0.2 – 0.3fF[0.1 fF]		150 $\Omega$ [200 $\Omega$ ]

[ ] = HC-35/HC-45

The typical equivalent data are given in **Table 3**. The motional capacitance  $C_1$  reduces as a square of the overtone n:

$$C_1 \text{ typ} \sim \frac{1}{n^2} \quad (5)$$

The attainable Q-value will also fall on increasing frequency. For this reason, the  $R_1$  values will increase, and will be in the order of typically 20 to 200  $\Omega$ .

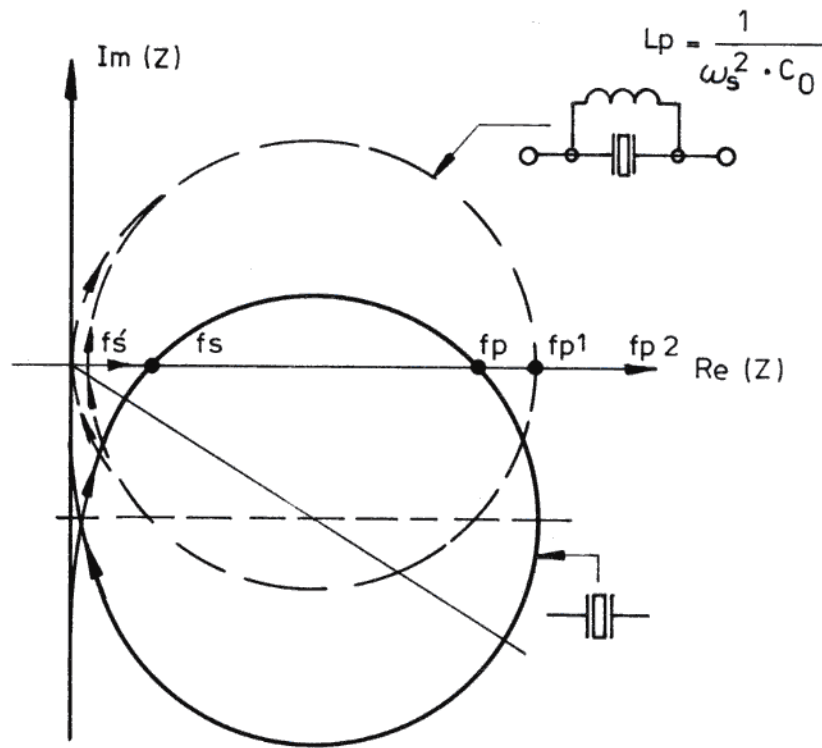
On increasing frequency, the static capacitance  $C_0$  will form an ever increasing bypass for the crystal. The results of this can be seen in **Figure 7**. Given is the locus of the complex crystal impedance. In the vicinity of parallel and series resonance, it will represent a circle, which cuts the real axis at  $f_s$  and  $f_p$ . The spacing of the center point of the circle from the real axis will become greater, the lower the reactive resistance of  $C_0$ . At low values of  $X_{C_0}$ , the phase slope in the vicinity of series resonance will be lower, and especially the phase deviation will be less in the inductive direction. Finally, it can happen that the circumference no longer cuts the real axis, which means that no real resonant point is present, at which the crystal is purely ohmic. For this reason, the static capacitance should be compensated for using a parallel inductance:

$$L_p = \frac{1}{\omega_s^2 C_0} \quad (6)$$

in excess of a certain limit.

A rule-of-thumb for this limit is:

$C_0$ -compensation should be provided when  $X_{C_0} < 5 \times R_1$ , or generally in excess of 100 MHz.



**Fig. 7:**  
Locus of a crystal  
with and without  
 $C_0$ -compensation

The result of the compensation is given in Figure 7 in the form of a dashed line. The locus is symmetric to the real axis, however, there exist two parallel resonances above and below  $f_s$ . The attainable total phase deviation is up to  $\pm 90^\circ$ .

A compensating coil having a low Q ( $R_p > 10 R_1$ ) is suitable, and the compensation condition (6) need not be exactly maintained. It is sufficient to use a standard inductance (or a corresponding number of turns wound on a  $10 \text{ k}\Omega$  resistor).

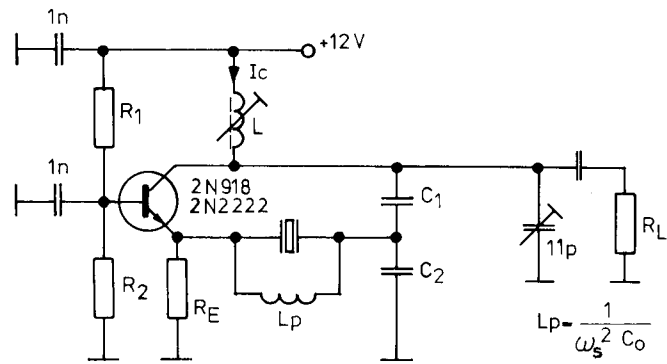
Aperiodic oscillators will not operate reliably with overtone crystals, even when this is stated from time to time (5). A resonant circuit should always be provided in order to avoid oscillation at the fundamental frequency.

When using a Pierce circuit as shown in Figure 4 b, it is possible for the collector capacitor to be replaced by a capacitively detuned circuit. Since overtone crystals are usually aligned in series resonance, this will result in a residual load capacitance for this circuit, which means that only customer-specified crystals will operate satisfactorily.

In order to pull the crystal frequency towards a lower value, an inductance is often connected in series with the crystal. However, it is possible for parasitic oscillations to be excited across this inductance L and the static capacitance of the crystal  $C_0$ , which could be difficult to neutralize (see section 4).

It is therefore better to use a true series-resonant circuit as shown in **Figure 8**. The values of  $C_1$  and  $C_2$  are selected so that a sufficient loop gain results. This is reduced both by the divider  $C_1/C_2$  and by the voltage division across the crystal impedance and the input impedance at the emitter (20).

When selecting a suitable transistor, a rule-of-thumb is, that the transit frequency should be at least ten times that of the oscillator frequency. In addition to this, transistors are to be recommended that have a high DC-gain ( $h_{FE}$ ) at a low base resistance ( $r_{bb'}$ ).



	75MHz	120MHz	150MHz	200MHz
$C_1$ [pF]	8	8	5	3
$C_2$ [pF]	100	50	25	20
$I_c$ [mA]	25	25	5	5
$R_E$ [ $\Omega$ ]	510	390	1.1k	1.1k
$R_L$ [ $\Omega$ ]	470	300	600	600
$L_p$ [ $\mu$ H]	0.25	0.10	0.08	0.05

Fig. 8: Overtone crystal oscillator up to 200 MHz

## 2. LOAD CAPACITANCE; OSCILLATORS WITH PARALLEL AND SERIES RESONANCE

The designations series and parallel resonance are often combined in a confusing manner. In the case of series-resonance crystal oscillators, the crystal will oscillate together with its pulling elements at the low-impedance resonance. Such a case is the example of the Butler oscillator given in Figure 3. However, this does not mean that the oscillator operates at the series-resonance frequency of the crystal. The Butler remains a series-resonance oscillator, even when the crystal is pulled with the aid of a series capacitor, or even when (at higher frequencies) the phase angle of the transistor gain deviates from  $0^\circ$  or  $180^\circ$ .

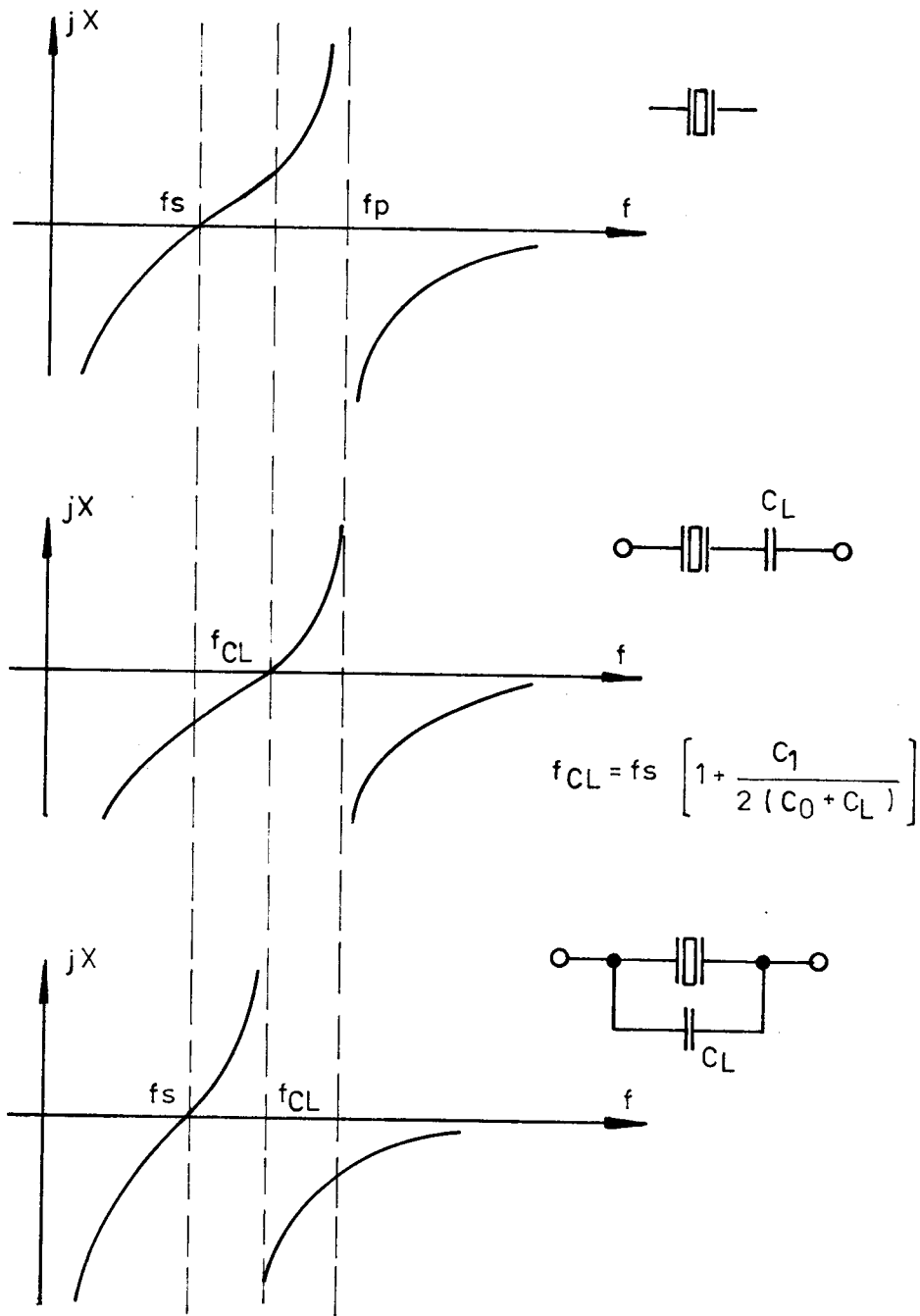
On the other hand, another commonly used definition is not advisable: This states that a series-resonance oscillator is designated by the fact that the oscillator will also oscillate when the crystal is replaced by a resistor. If this were the case, the Butler oscillator given in Figure 3 would not be a series-resonance oscillator; however, would be, if the collector resistor of transistor 1 was replaced by a resonant circuit.

A series load capacitance  $C_L$  will generate a new series resonance at

$$f_{CL} = f_s \left( 1 + \frac{C_1}{2(C_0 + C_L)} \right) \quad (7)$$

In the case of a parallel-resonance oscillator, the oscillator will operate at a high-impedance resonance together with its adjacent (pulling) elements. In the case of the Colpitts oscillator shown in Figure 5,  $C_1$  and  $C_2$  are connected in series across the crystal. In the case of an ideal amplifier stage they will form the load capacitance  $C_L$  and reduce the parallel resonance frequency of the crystal to  $f_{CL}$ . If this value of  $C_L$  is just as great as the series- $C_L$  in the upper case, the pulled series-resonance frequency will be the same as the pulled parallel-resonance in the last example. In both cases, the crystal will operate at a point at which it behaves as a high-Q inductance. This is summarized in **Figure 9**.





**Fig. 9: Crystals with load capacitances**

Crystals that are designated with a load capacitance are usually aligned by the manufacturer in conjunction with a series capacitor. However, since both measurements are equivalent, it is immaterial for the crystal specification whether a parallel or series resonance oscillator is to be used. It is sufficient for a load capacitance to be given. In this case it is advisable, if possible, for standard values of  $C_L$  (e.g. 30 pF) to be used by varying the other capacitances of the oscillator.

### 3. CRYSTAL DISSIPATION IN THE OSCILLATOR

#### 3.1. Typical Values

The crystal dissipation of crystals will exhibit the following values in the various oscillator circuits:

Tube oscillator:	1 to 10 mW, typical 2 mW
TTL-oscillator:	1 to 5 mW
Transistor oscillator:	10 $\mu$ W to 1 mW, typical 100 $\mu$ W
CMOS oscillator:	1 $\mu$ W to 100 $\mu$ W

Since the crystal frequency and resonance impedance are somewhat dependent on the load, a nominal load should be specified especially for low-tolerance crystals. The following limits are advisable physically:

#### **Crystal Drive Level not more than 2 mW**

Higher drive levels will deteriorate the stability, the Q and aging characteristics. In the case of LF-crystals, and very small AT-crystals (cases HC-45/U or HC-35/U), 2 mW will be too much. Since reactive power = Q x effective power, a reactive power of 200 W will be periodically present at the reactances of the crystal at a drive level of 2 mW and a Q of 100 000 !

#### **Crystal Drive Level not less than 1 $\mu$ W**

Too low a drive level could cause difficulties in commencing oscillation, since – physically speaking – a certain minimum amount of energy is required for commencing oscillation. This varies, as a result of unavoidable fluctuations in the quality of the transition crystal/electrode (in the submicroscopic range), and other damping influences. This can cause problems with certain CMOS and other low-power oscillators.

#### 3.2. Practical Determination of the Crystal Drive Level

Since the transistor parameters are only low-signal magnitudes, they are only valid as long as the transistor operates in class A. In the case of a self-limiting oscillator, the transistor operates up into the non-linear saturation range. Therefore, it is virtually impossible to calculate the expected drive level of the crystal.

In order to determine the actual drive level of the crystal in a measuring setup, either the RF-current to the crystal or the voltage difference across the crystal is measured with the aid of a thermistor, oscilloscope or RF-voltmeter. If the equivalent data of the crystal ( $C_0$ ,  $C_1$ ,  $R_1$ ) are known, it is possible for the phase angle to be calculated from the oscillator frequency. Finally, the actual power can be determined from this. This is often very much lower than would result without consideration of the phase.

### 4. PULLABILITY AND MODULATION OF CRYSTAL OSCILLATORS

#### 4.1. Pullability

The pullability of the oscillator frequency is dependent both on the pulling reactances of the circuit and the equivalent data of the crystal.

#### 4.1.1. Pulling Reactances

The reactive impedance characteristic of a crystal is given in **Figure 10a**, whose frequency is pulled using an inductance, or a series-resonant circuit in series with the crystal. Generally speaking, the series resonance will be pulled to the following with the aid of a connected reactive impedance  $X_V$ :

$$f_X \approx f_s \left[ 1 + \frac{C_1}{2 \left( C_0 - \frac{1}{\omega_s X_V} \right)} \right] \quad (8)$$

A series capacitor  $C_V$  will increase the frequency to

$$f_C \approx f_s \left( 1 + \frac{C_1}{2 (C_0 + C_V)} \right) \quad (9)$$

A series inductance  $L_V$  will reduce the frequency to

$$f_{L_1} \approx f_s \left[ 1 - \frac{C_1}{2 \left( \frac{1}{\omega_s^2 L_V} - C_0 \right)} \right] \quad (10)$$

If a series resonant circuit is connected, it is possible for the series-resonant frequency to be tuned up and down:

$$f_{LC_1} \approx f_s \left[ 1 + \frac{C_1}{2 \left( C_0 - \frac{1}{\omega_s^2 L_V - \frac{1}{C_V}} \right)} \right] \quad (11)$$

These equations are valid at sufficient accuracy for the pulling range up to 1000 ppm ( $10^{-3}$ ).

If an inductance is provided as pulling element, an additional series-resonance point ( $f_{L_2}$  or  $f_{LC_2}$ ) will appear. At this position, the connected inductive reactive impedance will be in resonance with the capacitance  $C_0$  of the crystal. This frequency can, it is true, be relatively far from the main resonance, however, it is possible for the oscillator to jump to this parasitic resonance.

The reactance characteristics of the circuit when provided with an additional parallel compensation of  $C_0$ , are given in **Figure 10 b**. This will lead to two parallel resonance positions at

$$f_{p_{1,2}} \approx f_s \left( 1 \pm \frac{1}{2} \sqrt{\frac{C_1}{C_0}} \right) \quad (12)$$

which are several hundred kHz below and above the main resonance frequency. The parallel inductance will increase the pulling range and is usually designed so that it is in resonance with the static crystal capacitance  $C_0$ :

$$L_p = \frac{1}{\omega_s^2 C_0} \quad (13)$$

The pulling equations result in the same manner as for equations 8 to 11, however, without the term  $C_0$ . In detail:

A series capacitor  $C_V$  will increase the series resonance to

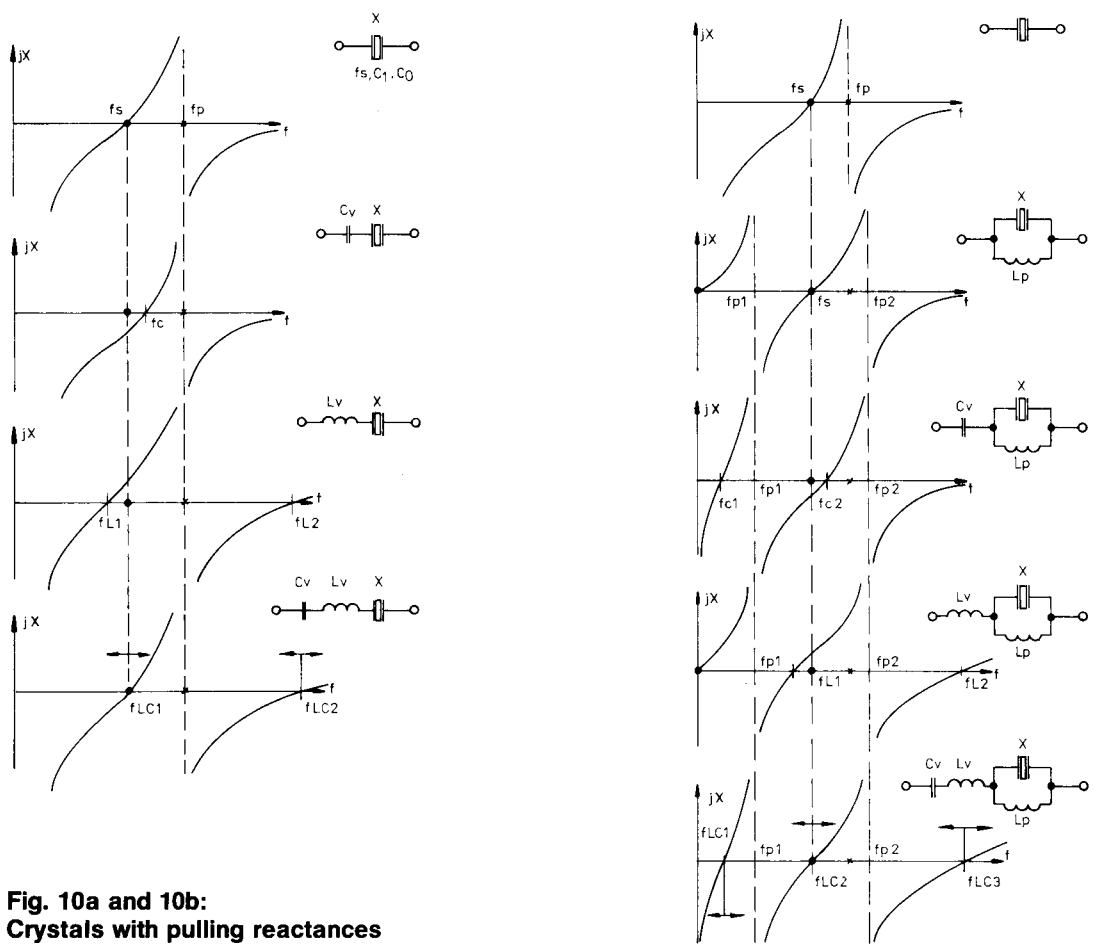
$$f_{C_2} \approx f_s \left( 1 + \frac{C_1}{2C_V} \right) \quad (14)$$

however, an additional series-resonance point  $f_{C_1}$  will appear below  $f_{p_1}$ .

A series inductance  $L_V$  will reduce the main resonance to

$$f_{L_1} \approx f_s \left( 1 - \frac{C_1}{2} \omega_s^2 L_V \right) \quad (15)$$

whereby a further series-resonance position  $f_{L_2}$  will appear in excess of  $f_{p_2}$ .



**Fig. 10a and 10b:**  
**Crystals with pulling reactances**

When pulling the frequency using a series circuit  $L_V$ ,  $C_V$ , a new frequency will result as:

$$f_{LC_2} \approx f_s \left[ 1 - \frac{C_1}{2} \left( \omega^2 L_V - \frac{1}{C_V} \right) \right] \quad (16)$$

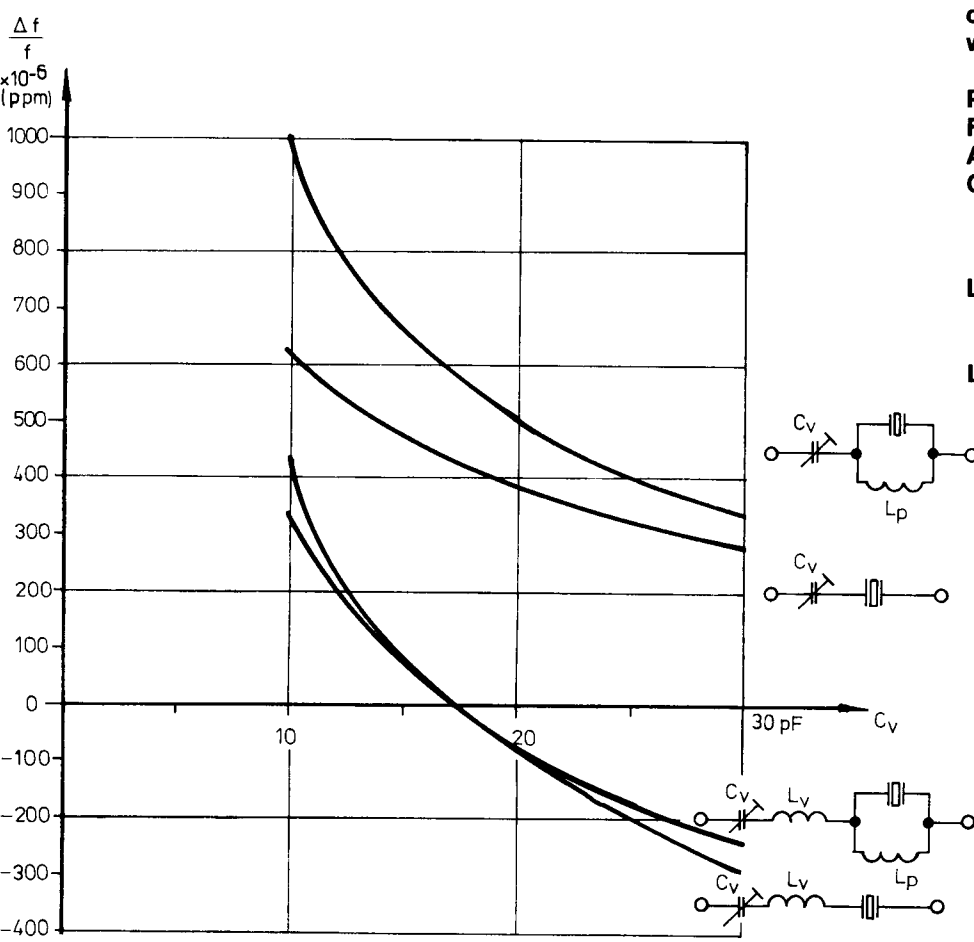
which can be below or above the main resonance.

In this case, two further series resonances  $f_{LC_1}$  and  $f_{LC_3}$  appear.

All additional resonance points are not crystal-controlled, but are parasitic resonances between the pulling link and the detuned circuit comprising  $C_0$  and  $L_p$ , which differs from the main resonance in a capacitive or inductive direction.

In the case of oscillators with a large pulling range, these additional resonances can cause considerable problems in the form of a jumping of the oscillator frequency. Since the parasitic resonances are often of lower impedance than the pulled main resonance, it is often difficult to avoid such jumping of the frequency.

**Figure 11** provides a quantitative evaluation of the pulling behaviour of the different types of circuit. It is based on a fundamental-wave crystal of 10 MHz with  $C_0 = 6$  pF and  $C_1 = 20$  fF with and without  $C_0$ -compensation. A series capacitor is compared to a series-resonant circuit  $L_v, C_v$ , when used as pulling element. In this case,  $C_v$  was varied between 10 pF and 30 pF, and  $L_v$  was selected so that the series-resonant circuit was tuned to the crystal frequency  $f_s$  at a mean  $\bar{C}_v = \sqrt{10 \times 30 \text{ pF}}$ . This resulted in a pulling range that was approximately symmetrically spread around  $f_s$ .



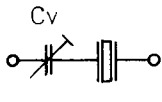
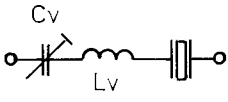
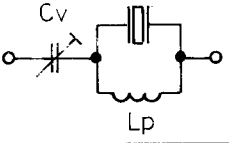
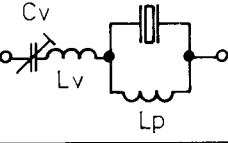
**Fig. 11:**  
**Pulling characteristic**  
**of different circuits**  
**with  $C_v = 10 - 30$  pF.**

**Parameters:**  
**Fundamental mode**  
**AT-crystal with**  
 **$C_0 = 6$  pF;  $C_1 = 20$  fF.**

$$L_v = \frac{1}{\omega_s^2 \sqrt{10 \text{ pF} \cdot 30 \text{ pF}}}$$

$$L_p = \frac{1}{\omega_s^2 C_0}$$

These curves were extended by a complex circuit analysis by which the crystal losses ( $Q_q = 39\,800$ ) and the coil losses ( $Q_{L_v} = Q_{L_p} = 100$ ) were taken into consideration. The results of this are summarized in **Table 4**.

				
Pulling range $\frac{\Delta f}{f}$ (ppm)	347	622	665	666
$R_1'$ (10 pF)	51.2 $\Omega$	40.6 $\Omega$	29.6 $\Omega$	30.9 $\Omega$
Impedance transformation $R_1'$ (30 pF)	28.8 $\Omega$	23.8 $\Omega$	21.1 $\Omega$	29.8 $\Omega$
Effective Q	$Q' = 39\,800$	$Q' = 30\,800 - 24\,400$	$Q' = 27\,000 - 37\,700$	$Q' = 26\,000 - 27\,000$
<p>Crystal: AT-Fundamental, <math>C_0 = 6</math> pF; <math>C_1 = 20</math> fF; <math>R_1 = 20</math> <math>\Omega</math> (<math>Q_X = 39\,800</math>)</p> <p>Pulling elements: <math>L_V = \frac{1}{\omega_0^2 \sqrt{10 \text{ pF} \cdot 30 \text{ pF}}}</math>; <math>L_P = \frac{1}{\omega_0^2 \cdot C_0}</math>; <math>Q_{L_V} = Q_{L_P} = 100</math>  (at 10 MHz: <math>L_V = 14,6</math> <math>\mu\text{H}</math>; <math>L_P = 42,2</math> <math>\mu\text{H}</math>)</p>				

**Table 4: Pulling behaviour of various types of pulling circuits for  $C_V = 10$  to 30 pF**

## 4.1.2. Remarks on the Pulling Circuits

### 4.1.2.1. Crystal without $C_0$ -Compensation

If only a pulling capacitor  $C_V$  is used, this will result in the smallest pulling range, and all frequencies will be below the crystal frequency. The overall Q will remain practically constant over the whole range (same crystal Q), however, the series-resonance impedance will be transformed up to the following value:

$$R_1' = R_1 \left(1 + \frac{C_0}{C_L}\right)^2 \quad (17)$$

At very low load capacitances, it is possible for  $R_1'$  to attain very high values (e.g. at 5 pF,  $R_1' = 96.8$   $\Omega$ ). This can cause the oscillator to cease operation, or cause problems in commencing oscillation. Since the pullability increases considerably at such low capacitances (see equ. 9), this can cause instability, or non-reproducibility of the oscillator frequency. This fault is often to be found in the data sheets of integrated circuits. For instance, an effective load capacitance of approximately 6 pF (!) is given in the original data sheet of the well-known mixer SO 42 P. Similar load values are given in conjunction with the  $\mu\text{P}$ -clock oscillator of the 8080-system.

In a series circuit, the pulling range will be considerably larger and can be set to be symmetrical to the crystal series resonance. The transformed impedance is less than in the first case. The overall Q will, however, deteriorate greatly due to the inductance. As can be seen in **Figure 12**, it is possible for the pulling range to be greatly extended asymmetrically towards lower frequencies, if  $L_V$  is increased, however, the overall Q will quickly disappear. It will be seen in the lower curve of Figure 12 that the pulling range is  $1106 \times 10^{-6}$ , however, the Q is reduced to 12 500 ! This is a general rule for any pulling circuit equipped with inductances.

**4.1.2.2. Crystals with C<sub>0</sub>-Compensation**

The pulling range will be very large if only one pulling capacitor is used in addition to the C<sub>0</sub>-compensation, and the transformed dissipation resistance R<sub>1</sub>' will be considerably more favorable than with other circuits. The overall Q also remains relatively high. It is only of disadvantage for certain applications that the oscillator frequency will always be in excess of the crystal frequency.

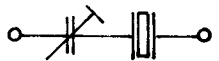
When using a series-circuit in addition to the C<sub>0</sub>-compensation, the pulling range will not be larger than in the previous example inspite of the larger number of components; however, it will be symmetric to the crystal frequency. It is true that R<sub>1</sub>' is relatively constant, however, the overall Q is very low due to the two inductances.

To summarize, it should be noted that larger pulling ranges are possible with C<sub>0</sub>-compensation than without. However, the effect of the compensation is reduced to a minimum when pulling with a series-resonant circuit. In order to achieve a large pulling range, it is usually best to use a circuit with C<sub>0</sub>-compensation and pulling capacitor (upper curve in Figure 11).

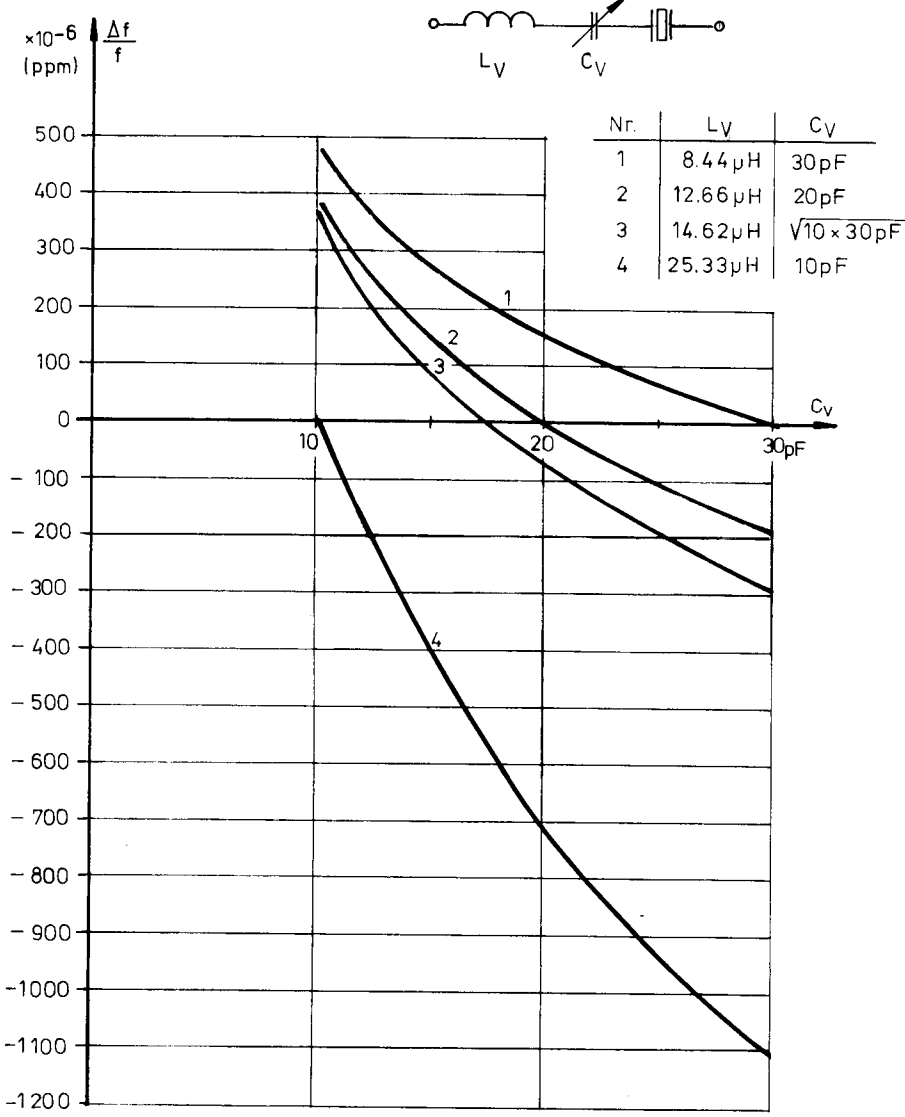
**4.1.3. Effect of the Motional Parameters of the Crystal**

As can be seen in equations (8) to (16), the pullability will be better, the greater the dynamic capacitance C<sub>1</sub> of the crystal. This can be influenced within certain limits by the design of the crystal, however, this will cause an increase of the static capacitance C<sub>0</sub> of the crystal, as well as an increase of the intensity of unwanted, anharmonic modes. For this reason, the realization of extreme pulling demands in practice should be found in close cooperation with the crystal manufacturer.

The pullability of overtone crystals is reduced by factor 1/n<sup>2</sup> in the same manner as the motional capacitance C<sub>1</sub>. **Table 5** gives the values for the simplest pulling circuit using only a pulling capacitor without C<sub>0</sub>-compensation. A crystal oscillator of an oscillator chain which is to be pulled in frequency should therefore be equipped with a crystal having a high fundamental-wave frequency (20 to 30 MHz), whereas the lowest possible overtone should be used in the case of overtone crystals.

 C <sub>series</sub> = 10 - 30 pF	Funda- mental	3rd overtone	5th overtone	7th overtone	9th overtone
	Crystal data: C <sub>0</sub> C <sub>1</sub>	6 pF 20 fF	5 pF 2 fF	5 pF 0.7 fF	5 pF 0.35 fF
Pulling range $\frac{\Delta f}{f}$	347 · 10 <sup>-6</sup>	38.1 · 10 <sup>-6</sup>	13.3 · 10 <sup>-6</sup>	6.7 · 10 <sup>-6</sup>	4.7 · 10 <sup>-6</sup>

**Table 5: Pullability as a function of overtone**



**Fig. 12:**  
**Pulling with series-resonance circuits**  
 $C_V = 10 - 30 \text{ pF}$   
**Parameter:**  
 $C_0 = 6 \text{ pF}$   
 $C_1 = 20 \text{ fF}$

### 4.2. Modulation of Crystal Oscillators

The pulling curves given in **Figures 11 and 12** are more or less non-linear. When pulling with varactor diodes, the non-linearity of the capacitance-voltage characteristic will have an opposite effect, which means that this will improve the modulation characteristic.

As an example, the modulation characteristic is given in **Figure 13** when using a varactor diode type BB 109 in conjunction with the four pulling circuits given in section 4.1.

The capacitance-voltage characteristic of the diode is given in **Figure 13a**.

Even the simplest pulling circuit results in a practically straight modulation characteristic in this (special) case. Even the circuit equipped with a series-circuit will provide good linearity. On the other hand, the  $C_0$ -compensated pulling circuit will be inferior to the appropriate uncompensated circuit – especially at larger frequency-deviation levels.

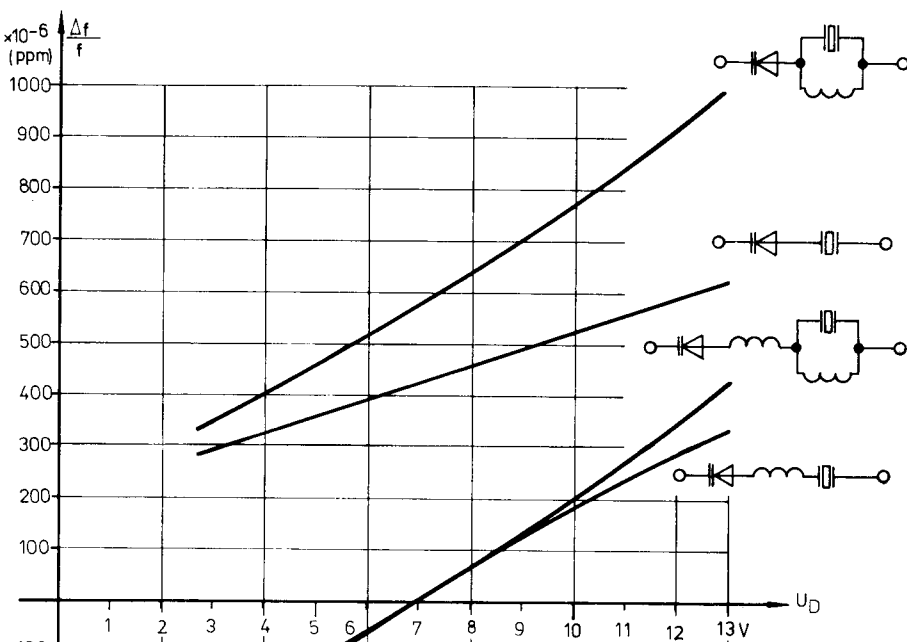


It is assumed during this calculation that the RF-voltage across the varactor diode is considerably less than the DC-voltage, which is usually not the case. This is to be shown in **Figure 14** as an example using the simplest pulling circuit. At resonance, the crystal will possess an inductive reactive impedance that coincides with the reactive impedance of the load capacitance of the opposite sign.  $C_0$  is not taken into consideration here. If the total voltage  $U_0$  is present across the crystal and varactor diode, the following voltage values will be present across the reactance of the crystal and the load capacitance, which is assumed to be loss-less:

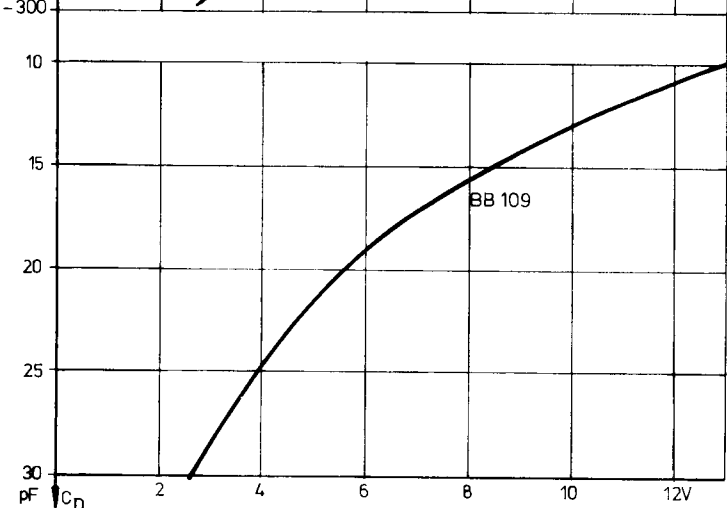
$$U_{L_1} = U_{C_D} = Q_X \times U_0 \quad (18)$$

where  $Q_X$  is the Q of the crystal. The voltage  $U_0$  is present across  $R_1$  and will determine the drive level of the crystal. At a drive level of  $P_q$ , a voltage will be built up across the varactor diode due to the resonance peak.

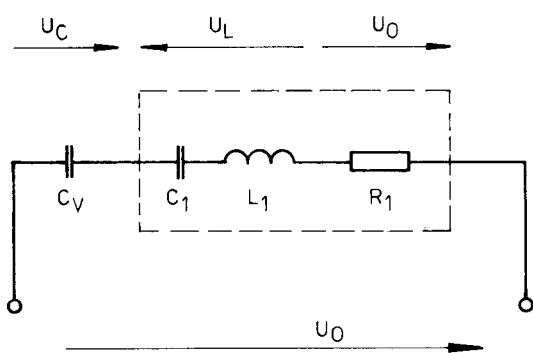
$$U_{C_D} = Q_X \sqrt{P_q \cdot R_1'} = \frac{1}{\omega_0 C_1} \sqrt{\frac{P_q}{R_1}} \quad (19)$$



**Fig. 13:**  
Modulation behaviour  
with varactor diodes  
BB 109



**Fig. 13a:**  
Diode characteristic  
 $C_D = f(U_D)$



**Fig. 14:**  
Resonance peak  
across the load  
capacitance

Example:  $P_q = 2 \text{ mW}$ ,  $R_1 = 20 \Omega$ ,  $f_0 = 10 \text{ MHz}$ ,  $C_D = 10 \text{ pF}$

In this realistic case, an AC-voltage will result across the varactor diode of  $U_{C_D} = 15.9 \text{ V}$ , which corresponds to a peak-to-peak voltage of  $45 \text{ V}$ , which is superimposed on the DC-voltage. In the case of the diode BB 109, a bias voltage of approximately  $13 \text{ V}$  is necessary for  $C_D = 10 \text{ pF}$ . Even then, the whole diode characteristic will be swept in time with the RF-voltage! This will be improved by the anti-phase (series) connection of two diodes, but will always be problematic!

## 5. FREQUENCY STABILITY OF CRYSTAL OSCILLATORS

### 5.1. Long-term Stability

The long-term stability is dependent on the aging characteristics of the external components, especially on the  $Q$  of the resonant circuits and the damping effect of the transistors on the  $Q$  of the crystal. It is also dependent, of course, on the aging of the crystal, which differs according to the type of crystal and its drive level in the oscillator, and will amount to a typical value of  $1$  to  $3 \times 10^{-6}/\text{year}$  during the first year. Since the aging is reduced logarithmically as a factor of time, it is possible to reduce this by previously aging the crystal, if possible at the manufacturer, at a temperature of between  $85^\circ\text{C}$  and  $125^\circ\text{C}$ .

The drive level of the crystal should be as low as possible for an oscillator that is to have a good long-term stability ( $1$  to  $20 \mu\text{W}$ ). Due to their better temperature characteristics, AT-crystals are preferable. When very stable crystal oscillators are required, relatively low-frequency overtone AT-crystals should be used due to their higher  $Q$  and higher  $L_1/C_1$  ratio. In this case, crystals operating at their third or fifth overtone of  $5$  or  $10 \text{ MHz}$  are used.

### 5.2. Short-term Stability

The short-term stability of crystal oscillators was only of interest in the past for high-precision oscillators such as secondary frequency and time standards. In recent times, however, this has become more and more important due to the widespread use of synthesizers in HF, and especially in VHF and UHF receivers, as well as for oscillator chains for the microwave frequencies.

The noise-content at the output of a crystal oscillator

$$U(t) = (U_0 + \varepsilon(t)) \sin(\omega_0 t + \varphi(t)) \quad (20)$$

will have a mean amplitude  $U_0$ , which will vary in the order of a noise component  $\varepsilon(t)$ , and an overall phase with the center frequency  $\omega_0$ , which has a noise component  $\varphi(t)$ . Since a phase variation  $\frac{d\varphi}{dt}$  as a function of time is correlated with a frequency, this means that:

$$f(t) = f_0 + \frac{1}{2\pi} \cdot \frac{d\varphi}{dt} \quad (21)$$

The oscillator signal will thus be modulated by the phase noise, and will possess noise sidebands, which will be visible on a sensitive, selective spectrum analyzer.

In the receive mixer, the input signals are mixed with the oscillator signal and its noise sidebands. This means that a noise-signal will be present in the passbands in addition to the selected input signal. This noise-component can be so large that it is able to block the receiver (6), (7).

Simple methods of measuring phase noise were described in (8) and (9). Further details were given in (10). The measuring and evaluation method described by the IEC in (11) has found international recognition.

Details regarding the noise-behaviour of crystals are given in (12) and (13). The following aspects should be considered during the design of short-term, stable crystal oscillators:

In contrast to extremely long-term, stable crystal oscillators, the drive level to the crystal should be relatively high (100 to 500  $\mu\text{W}$ ) for this application.

The Q of the crystal will be dampened in any oscillator; in the case of single-stage, self-limiting oscillators, the effective Q will amount to only 15 % to 20 % of the Q of the crystal. Usually, series-resonance oscillators are more favorable than parallel-resonance oscillators.

In the case of bipolar transistors, the noise will be mainly dependent on the base-emitter path. In this case, the noise of PNP-transistors will be lower than a complementary NPN-transistor. MOSFETs have a very high noise level, where 1/f-noise dominates at low frequencies, and thermal noise of the drain-source path at higher frequencies. Junction-FETs possess lower noise levels in comparison to bipolar transistors and MOSFETs. For this reason, a high-current power FET such as type CP 643, or P 8000 (14) is recommended for low-noise crystal oscillators.

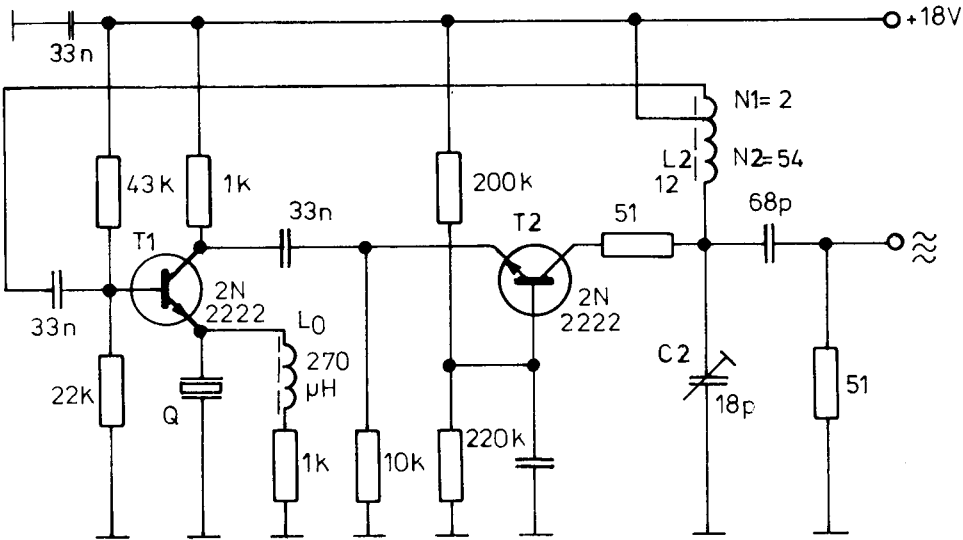
If bipolar transistors are to be used, then one should select types with the highest possible DC-gain ( $h_{FE}$ ), but with very low base resistance ( $r_{bb'}$ ), in other words, typical VHF transistors, which should then be used at the lowest collector current.

However, the short-term stability can be improved more using the following means than by the above methods (even including especially designed crystal):

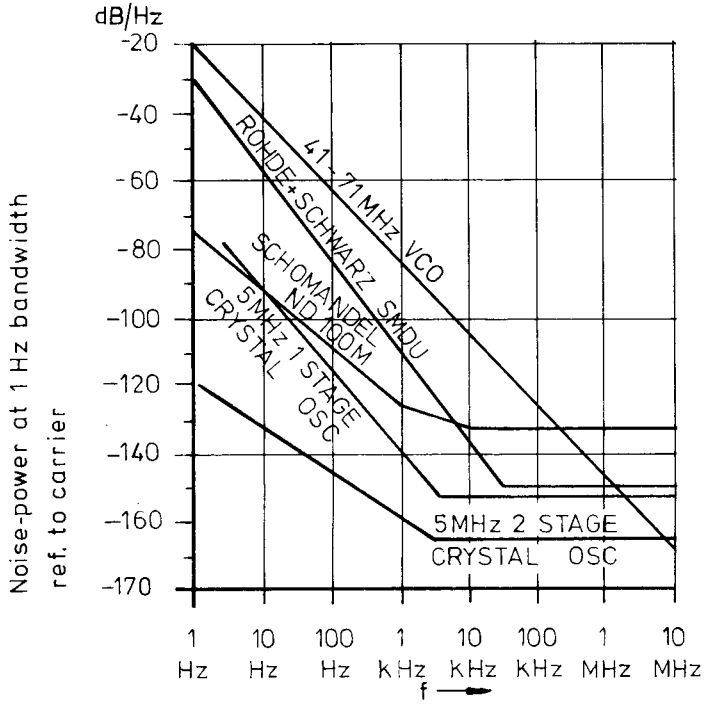
Single-stage crystal oscillators should be avoided. They possess a very high phase noise, since the transistor is driven into limiting. In this case, the collector base-voltage will be virtually zero during parts of the cycle, and the base-emitter threshold voltage (silicon: 0.6 V) will be exceeded. The transistor impedance that is »seen« by the crystal, will fluctuate in time with the RF-signal, which will generate strong noise sidebands on the oscillator signal.

This means that the limiting function of an amplifier stage connected to a crystal oscillator must be avoided. However, an amplitude control loop is unfavorable, since this could generate additional phase noise.

The best means of improving the short-term stability is to use a strong RF-feedback. A well-proved circuit was introduced in 1972 by M. M. Driscoll (15), which uses a third overtone 5 MHz crystal. Since then, several circuits based on this have been published that possess very good short-term characteristics up to 100 MHz (16) to (19). The basic circuit is given in **Figure 15**. It comprises a two-stage, three-pole oscillator with the resonant circuit  $L_2/C_2$ . A cascade circuit is used as amplifier (low internal feedback!), in which the first transistor is provided with feedback in the emitter circuit by the crystal (compensated with  $L_0$ ). Transistor  $T_1$  operates stably in class A ( $I_C = 5 \text{ mA}$ ). Transistor  $T_2$  is isolated from the crystal, and operates at a quiescent current of only 0,8 mA. This means that this stage is firstly limited and will determine the oscillation amplitude. The higher the series-resonance resistance  $R_1$  (for a given  $Q$ !) of the crystal, the better will be the short-term stability, since this will increase the negative feedback of  $T_1$ .



**Fig. 15: Crystal oscillator (5 MHz/third overtone) with exceptional short-term stability**



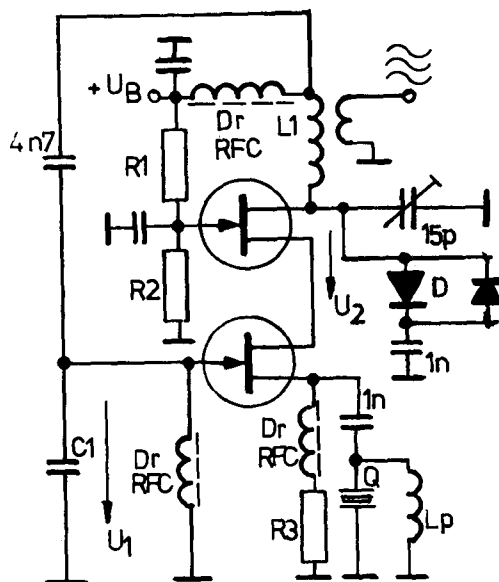
**Fig. 16a: Sideband noise of various signal generators in comparison to the 5 MHz oscillator shown in Fig. 15 according to (21)**

The crystal dissipation amounts to 85  $\mu\text{W}$  in this oscillator, the RF output level will be in the order of 4 dBm, and the effective Q will amount to approximately 50 % of the crystal Q. **Figure 16** shows the results of measurements of phase noise, that were given by U.L. Rohde in (21).

The amplitude limiting can also be achieved when biased Schottky diodes are connected in anti-phase at the output of  $T_2$  (due to the low  $1/f$ -noise of these diodes).

A low-noise oscillator which has been designed for use with a 96 MHz crystal according to this principle is given in **Figure 16 b**. Power-FETs type P 8000 are used and adjusted for stable class A ( $R_1$ ,  $R_2$ ,  $R_3$ ). The circuit is provided with a low feedback with the aid of  $C_1$  by selecting a relatively high capacitance value. The value of  $L_p$  is calculated according to equation 13, with  $C_0 = 5$  pF. Inductance  $L_1$  should have approximately 0,25  $\mu\text{H}$ , and diodes  $D$  used for amplitude limiting are Schottky diodes such as HP 2800.

**Fig. 16b:**  
Recommended circuit  
for a short-term stable  
VHF-crystal oscillator  
(96 MHz)



For alignment, the limiting diodes should be disconnected and the crystal short-circuited. With the aid of the trimmer, the self-excited frequency is aligned to approximately 96 MHz. After connecting the crystal, compensation coil  $L_p$  is aligned to 96 MHz with the aid of a dipper with the oscillator switched off. The oscillator must commence crystal-controlled oscillation immediately on connecting the operating voltage. The RF-amplitude will drop to approximately half the value of the self-limiting oscillator on connecting the diode ( $U_{limit}$ ). More details on this circuit were published recently by the author (28).

It is also possible to use the limiting characteristics of a subsequent differential amplifier instead of this (19).

Stable short-term crystal oscillators up to 100 MHz can be constructed according to this cascade principle with the crystal in the emitter circuit. A further reduction of the noise sidebands is obtained by placing a simple crystal filter after the crystal oscillator, as is shown in **Figure 17**.  $L_2$  is wound bifilar on a toroid core and aligned to the center frequency with the aid of  $C_2$ . The capacitance  $C_3$  should be selected so that it is equal to the static capacitance  $C_0$  of the crystal. The terminating impedance (22) is:

$$R_T = \frac{1}{2\pi f_0 C_0} \quad (22)$$

where the bandwidth is dependent on the ratio  $C_1 : C_0$  :

$$b_{ges} = f_0 \frac{C_1}{2C_0} \quad (23)$$

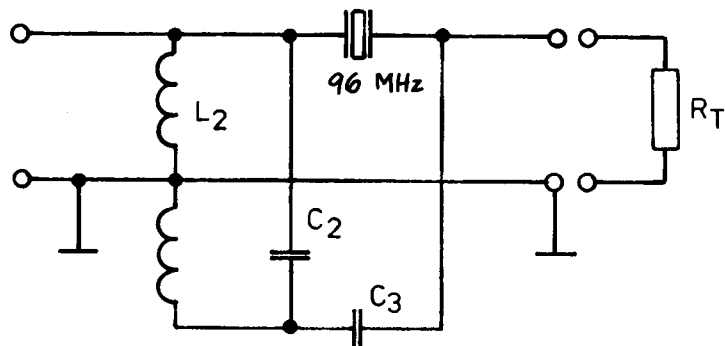


Fig. 17:  
Simple crystal filter  
for reducing  
phase noise

## 6. DIODE SWITCHING OF SEVERAL CRYSTALS

Poor practices in this respect are even to be found in professional equipment. A »worst case« -circuit is given in **Figure 18 a**, which possesses all typical faults:

The switching diodes have RF-voltage at both connections, and the blocking voltage is 0 V. This means that the diodes are opened in time with the RF-voltage and the interaction over the junction capacitance of the diodes can then cause:

- A pulling of the frequency due to the adjacent crystal
- Jumping of the oscillator frequency during the pulling process
- Finally a non-selected crystal having a higher Q, or a lower-impedance spurious resonance of any of the crystals can determine the frequency.

Since these effects often only occur at certain temperatures or operating voltages, it is often difficult to localize such faults. In some cases, the crystals are simply short-circuited by the parallel-connected diode, which may not have any effect if the diode impedance is far less than the impedance of the crystal.

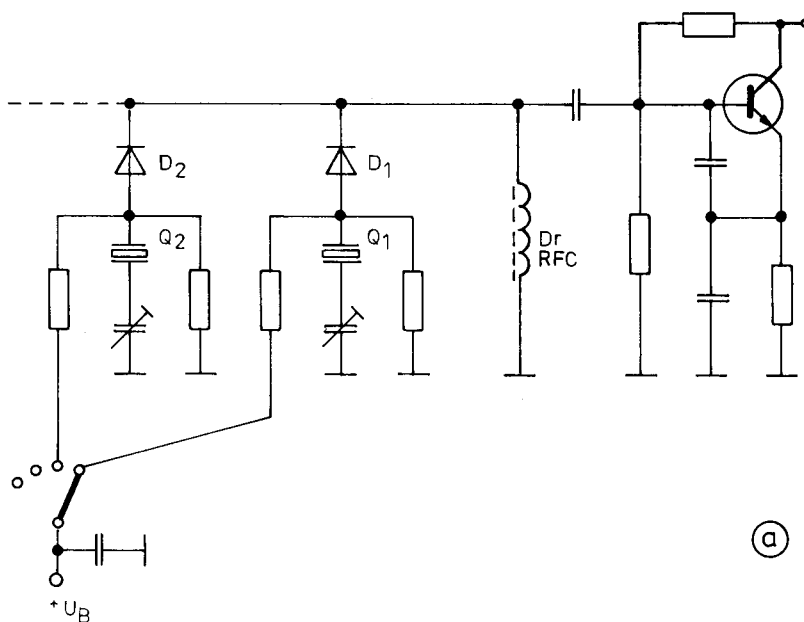


Fig. 18a: Worst-case method of diode switching

A favorable circuit is given in **Figure 18 b**. The diodes are »cold« with respect to RF-voltages at one pole and are biased with half the operating voltage. When diode  $D_1$  is opened,  $D_2$  will be blocked with half the operating voltage ( $U_B/2$ ). A further improvement of the isolation is obtained when an additional short-circuit diode is connected in parallel to the crystal (dashed). The disadvantage is, of course, the large number of components. Suitable, low-capacitance, fast switching diodes are, for instance, types BAY 67, 1 N 4148, or 1 N 4151.

The most reliable method of switching crystals is to use separate crystal oscillators !

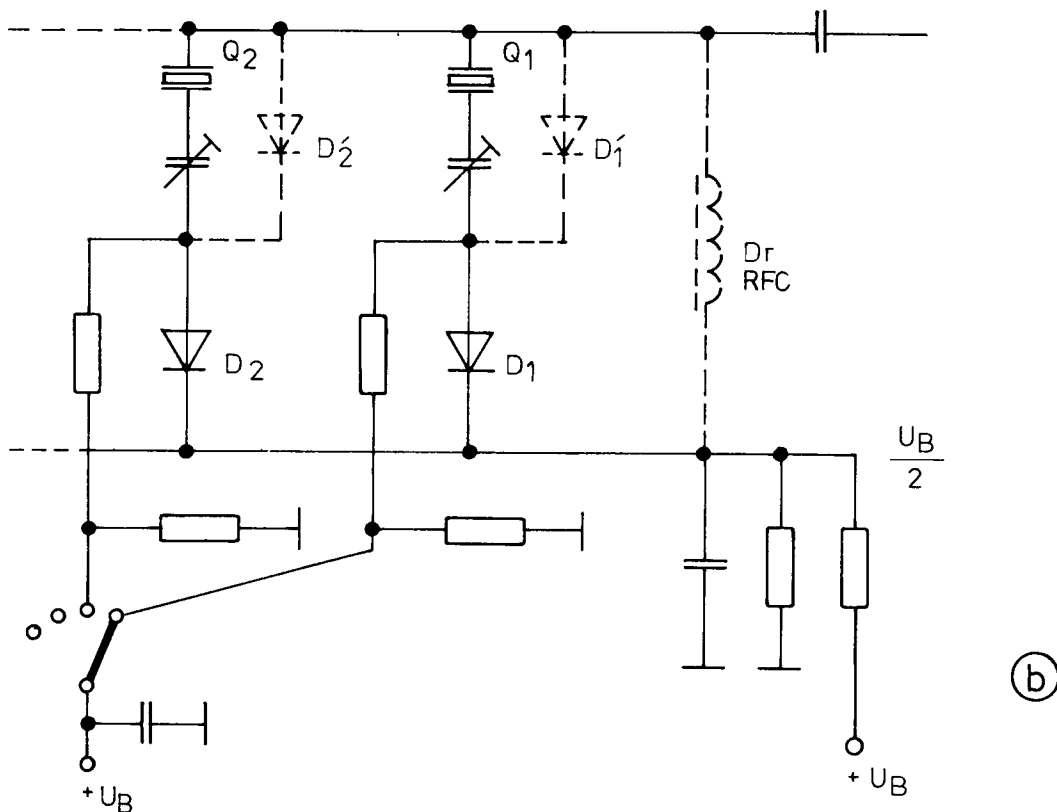


Fig. 18b: Improved diode switching circuit for crystal oscillators

## 7. MODERN CRYSTAL OSCILLATORS USING INTEGRATED CIRCUITS

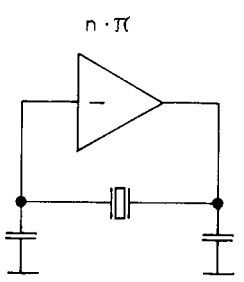
### 7.1. TTL and CMOS Gate Circuits

These oscillators are usually used where they are least suitable: in frequency and period counters, clocks and other measuring equipment.

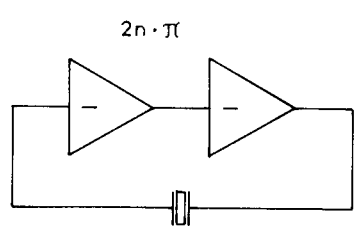
The well-known basic circuits for parallel and series-resonance oscillators using TTL and CMOS gates are given in **Figure 19** (23), (24), (25). The feedback resistors are necessary for linearizing the gates; if these were deleted, the oscillator would have difficulties in commencing oscillation, if at all.

TTL and CMOS gates are digital components and are optimized to switch cleanly between two limit values. It is true that they can be linearized using feedback, however, this linear behaviour is not reproducible and most certainly does not represent an optimum.

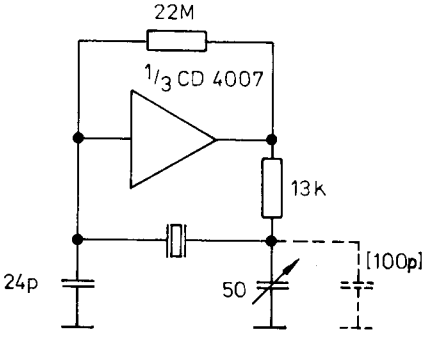
Let us consider, for instance, the amplitude and phase response of two series-connected, linearized Schottky-NAND gates SN 74 S 00 as a function of frequency (**Figure 20**). The theoretical phase shift of  $0^\circ$  will only be achieved in the vicinity of zero Hz (!), and in excess of this will achieve any phase angle between  $0^\circ$  and  $160^\circ$ , and will virtually represent an ideal inverter (26) at the upper frequency limit.



„Parallel-resonance“

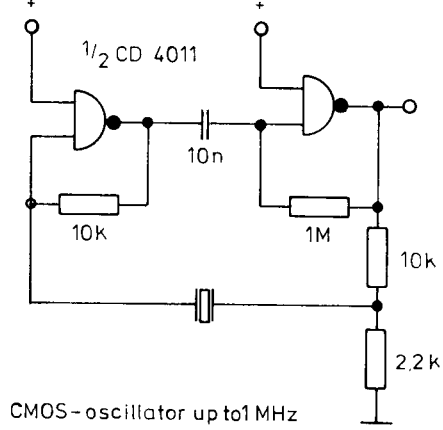


„Series-resonance“



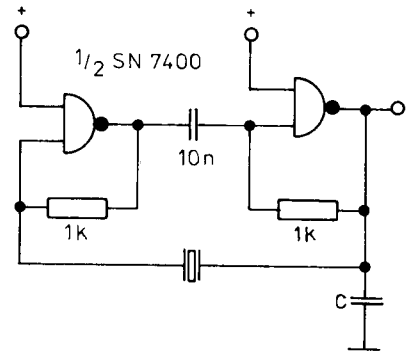
[ ] only at  $f < 1\text{MHz}$

CMOS- oscillator up to 3MHz



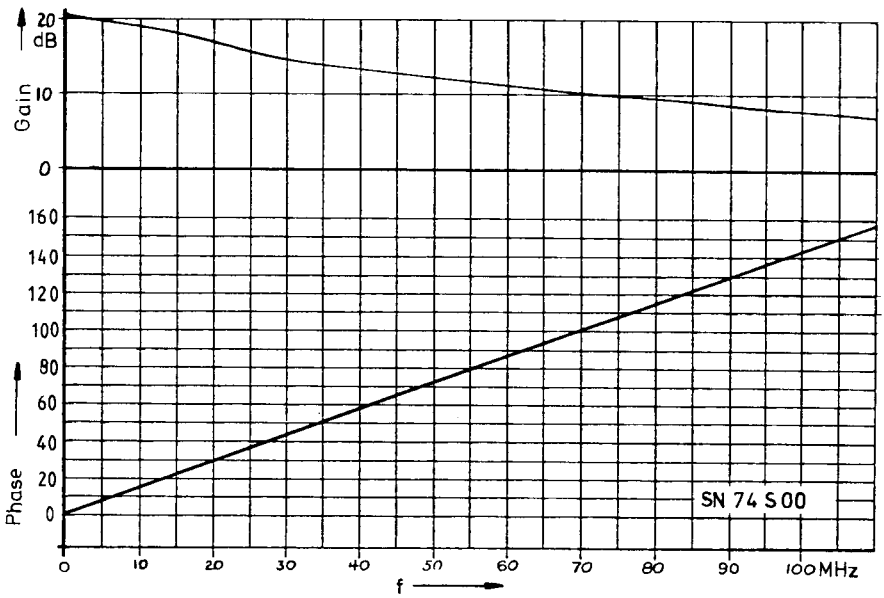
CMOS- oscillator up to 1MHz

f [MHz]	0,2	0,5	1	2	5
C	33 nF	1.2 nF	680 pF	330 pF	120 pF



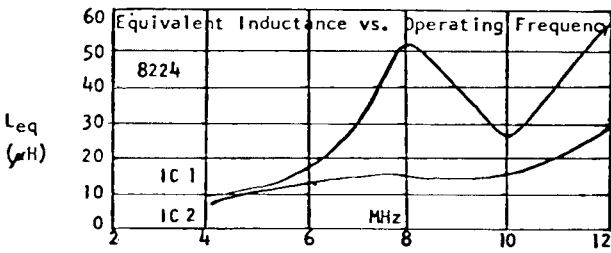
TTL- oscillator up to 5 MHz

**Fig. 19:**  
Oscillators with logic gates



**Fig. 20:** Amplitude and phase response of a linearized SN 74 S 00





8224 IC1

$f_s$ (MHz)	$\Delta f$ (Hz)	$\Delta f$ (PPM)	$L_{eq}$ ( $\mu h$ )	* $C_L$ To Cancel $L_{eq}$ (pf)
2	OPERATES ON 3RD. OVERTONE			
4	-156	-39	6.8	233
8	-2958	-370	13.8	29
10	-6776	-678	18.1	14
12	-12200	-1017	28.2	6
16	-59318	-3637	—	—
18	NOT CRYSTAL CONTROLLED			

8224 IC2

$f_s$ (MHz)	$\Delta f$ (Hz)	$\Delta f$ (PPM)	$L_{eq}$ ( $\mu h$ )	$C_L$ To Cancel $L_{eq}$ (pf)
2	OPERATES ON 3RD. OVERTONE			
4	-214	-53	9.4	170
8	-7455	-932	50.8	8
10	-8506	-851	25.2	1
12	-16503	-1376	54.8	3
16	-171947	-10541	—	—
18	NOT CRYSTAL CONTROLLED			

**Fig. 21:**  
Equivalent gate inductance  $L_{eq}$  and frequency error  $\Delta f$  with various samples of clock IC type 8224 (for microprocessor 8080)

Holmbeck (27) examined several samples of the clock IC 8224 of the microprocessor system 8080. The measured values of two samples are given in **Figure 21**. Generally speaking, the inverters have an inductive behaviour, in other words, react in a similar manner as when an inductance  $L_{eq}$  were connected in series; a strong fluctuation was found between the individual samples. The frequency response of this equivalent inductance is shown in the upper diagram for two samples of the same integrated circuit. The effects of this on a crystal oscillator can be seen in the two tables:

- At low frequencies (2 MHz) both oscillators will operate stably at the third overtone instead of the fundamental mode;
- The frequency shift with respect to series resonance will become greater and greater on increasing frequency and will attain thousands of ppm, with fluctuations of also thousands of ppm due to the spread from IC to IC.
- Sometimes, a capacitor is connected in series in order to compensate for the effect of  $L_{eq}$ . The values of such a capacitor are given in the last column. If, for instance, 14 pF are required by one IC at 10 MHz, it may be that only 1 pF (!) will be required with another IC. At frequencies in excess of 12 MHz, a capacitor will no longer be able to help.
- Finally, at 18 MHz, both crystal oscillators are no longer crystal-controlled. The cause of this is the second resonance position formed by  $L_{eq}$  and  $C_0$  of the crystal (see section 4). If this is far above the crystal frequency, this excitation can only be avoided when a small capacitor (lowpass) is connected in parallel to the feedback resistor of the gate.
- Since the loop gain is very high, it is also possible for spurious parasitic resonances to be present together with the parasitic external capacitances.

- In the case of LF-crystals, it is often easier to excite higher-frequency modes than the fundamental.
- Finally, it is also possible for the drive level of the crystal to become so high that the crystal will operate in an unstable manner.

This information is also valid for circuits equipped with integrated dividers such as CD 4060, MC 14521, MC 14410, E 1115, as well as the clock IC MC 6875 (for  $\mu$ P 6800), etc.

## 7.2. Survey On Integrated Circuits as Crystal Oscillators

### 7.2.1. ICs with Digital Outputs

The TTL-VCO circuits such as 74324, 74325, 74326, 74327, as well as their Schottky and low-power Schottky derivations fall under the same category as the gate oscillators. They operate more or less by chance when a crystal is provided instead of the frequency-determining capacitor.

Measurements made by the author gave the following results:

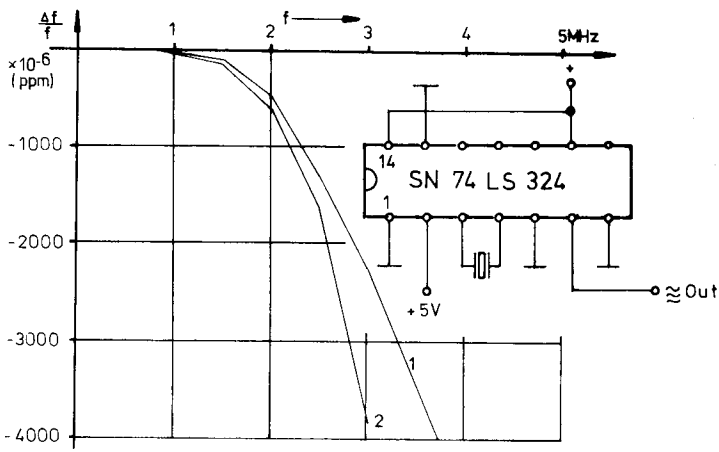
**Figures 22 a and 22 b** show the frequency error (difference between the oscillator frequency and the series-resonance frequency of the crystal) of several samples of types SN 74 LS 324 and SN 74 LS 325. The first ones oscillated wildly in excess of 3 or 4 MHz, and oscillated several 1000 ppm below the crystal frequency in excess of 2 MHz. IC type SN 74 LS 325 stopped oscillation, according to the sample in question, at 5, 6 or 8 MHz, and showed similar frequency errors, but at a higher frequency. Type SN 74 LS 326 only operated in a crystal-controlled manner at 1 MHz, only between 1 and 2 MHz, or only at 2 to 2.5 MHz.

During these measurements, the high »range« input (pin 2 of 324) was directly connected to + 5 V, and »frequency control« inputs (pin 13 of 324, pins 6 and 11 of 325, pins 9 and 10 of 326) were grounded. According to subsequent information from the manufacturer, better results should be possible if only + 4 V is present at these inputs instead of + 5 V and a bias voltage of + 1 V instead of 0 V.

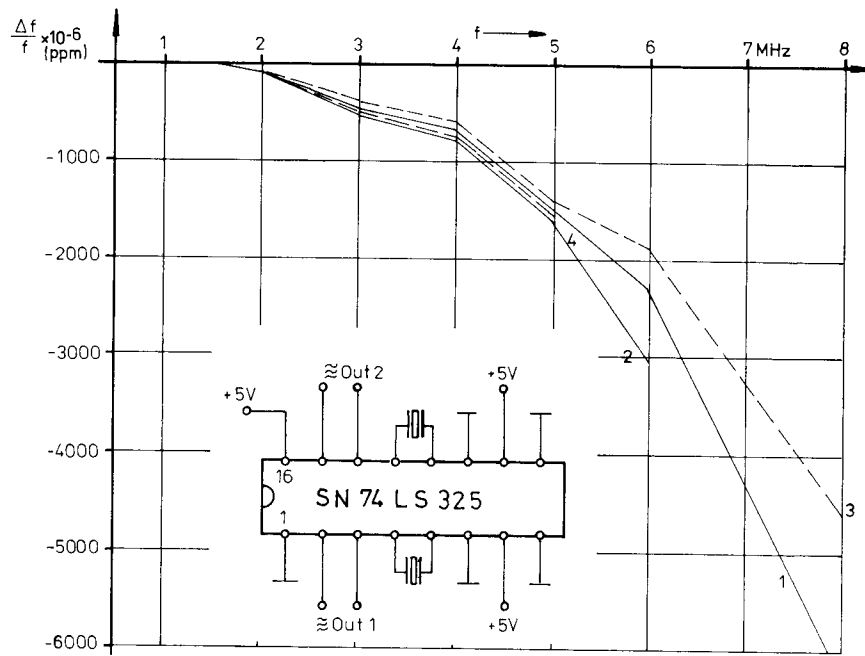
Two integrated circuits which are especially designed as crystal oscillators are now to be introduced:

**Plessey SP 705 B:** This is a Butler oscillator (1 to 10 MHz) with anti-phase outputs of  $f/2$  and  $f/4$ . According to the data sheet, the frequency error at 10 MHz is typically  $-50 \times 10^{-6}$ . However, the spread of this value is not given.

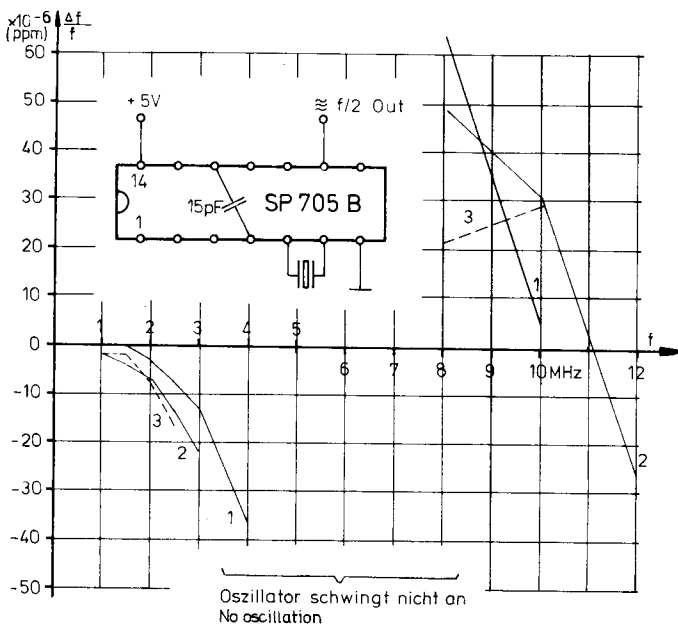
The results measured by the author are given in **Figure 23**: None of the oscillators operated below 1 MHz. From 1 MHz the well-known inductive behaviour (reduction of frequency) was noticed until oscillation ceased, according to the sample, in excess of 2.5 MHz, 3 MHz, or 4 MHz. In excess of 8 MHz, oscillation commenced again with a positive frequency error, and ceased again in excess of 10 MHz or 12 MHz. Furthermore, the built-in frequency divider did not work reliably. In the case of integrated circuits No. 2 and 3, the full, undivided frequency appeared at the output in the lower frequency range !



**Fig. 22a:**  
**Frequency error**  
**of an oscillator**  
**equipped with**  
**SN 74 LS 324 (TI)**



**Fig. 22b:**  
**Frequency error**  
**with SN 74 LS 325 (TI)**

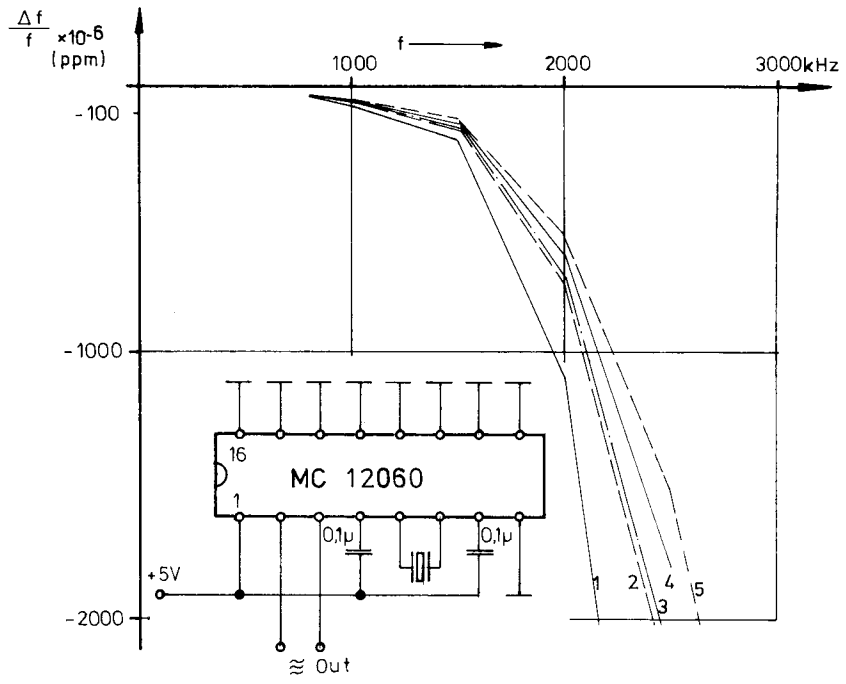


**Fig. 23:**  
**Frequency error**  
**with SP 705 B**  
**(Plessey)**

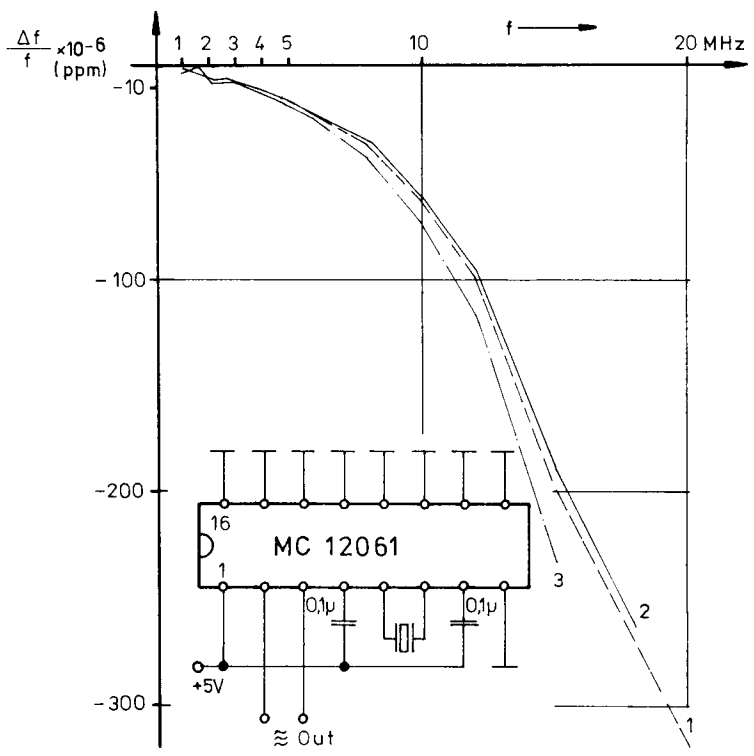
**Motorola MC 12060/12061** : IC type 12060 is designed for the frequency range from 0.1 MHz to 2 MHz, and type MC 12061 from 2 to 20 MHz. Both types are in the form of Butler circuits with built-in AGC. They supply two anti-phase sinewave, as well as TTL and ECL signals.

Measuring results:

**MC 12060:** All samples oscillated reliably up to in excess of 2 MHz, but provided considerable frequency errors of between  $-550 \times 10^{-6}$  and  $-1100 \times 10^{-6}$  at 2 MHz (**Figure 24**). At 1.5 MHz, these frequency errors are still between  $-110 \times 10^{-6}$  and  $-200 \times 10^{-6}$ . This large spread limits the applications of this IC to a large degree if a reproducible oscillator frequency is required.



**Fig. 24:**  
Frequency error  
with MC 12060  
(Motorola)



**Fig. 25:**  
Frequency error  
with MC 12061  
(Motorola)

**MC 12061:** Up to 15 MHz, all oscillators operate with frequency errors of up to  $-235 \times 10^{-6}$  (**Figure 25**), which is very good in comparison to other oscillator ICs. The frequency error of the individual oscillators in comparison with another is approximately  $45 \times 10^{-6}$  (at 15 MHz) which can satisfy moderate demands. IC No. 2 stopped oscillation in excess of 18 MHz, and IC No. 1 in excess of 20 MHz. All samples operated excellently at 1 MHz (frequency error only  $-1$  to  $-3 \times 10^{-6}$ ).

## 7.2.2. Analog ICs

### **National Semiconductors LM 175/275/375**

According to the data sheet, this type is a differential amplifier that can be used in series and parallel-resonance in a frequency range of 800 kHz to 20 MHz. The IC requires a large number of external components. As is also given in the data sheet, it is not possible for an aperiodic oscillator to be constructed! It is always necessary to provide a resonant circuit. On the other hand, no resonant circuit should be provided in the buffer stage (e.g. for harmonic suppression), since the stage could otherwise break into oscillation. This was also confirmed by Harrison in (4).

Only the series-resonance circuit was tested according to the data sheet. The crystal was fed back from the output to the non-inverting input of the differential amplifier, and an additional capacitor was inserted to ground. The inverting input is provided with a neutralizing voltage via a capacitive divider in order to ensure that the oscillator will not oscillate wildly together with the static capacitance of the crystal.

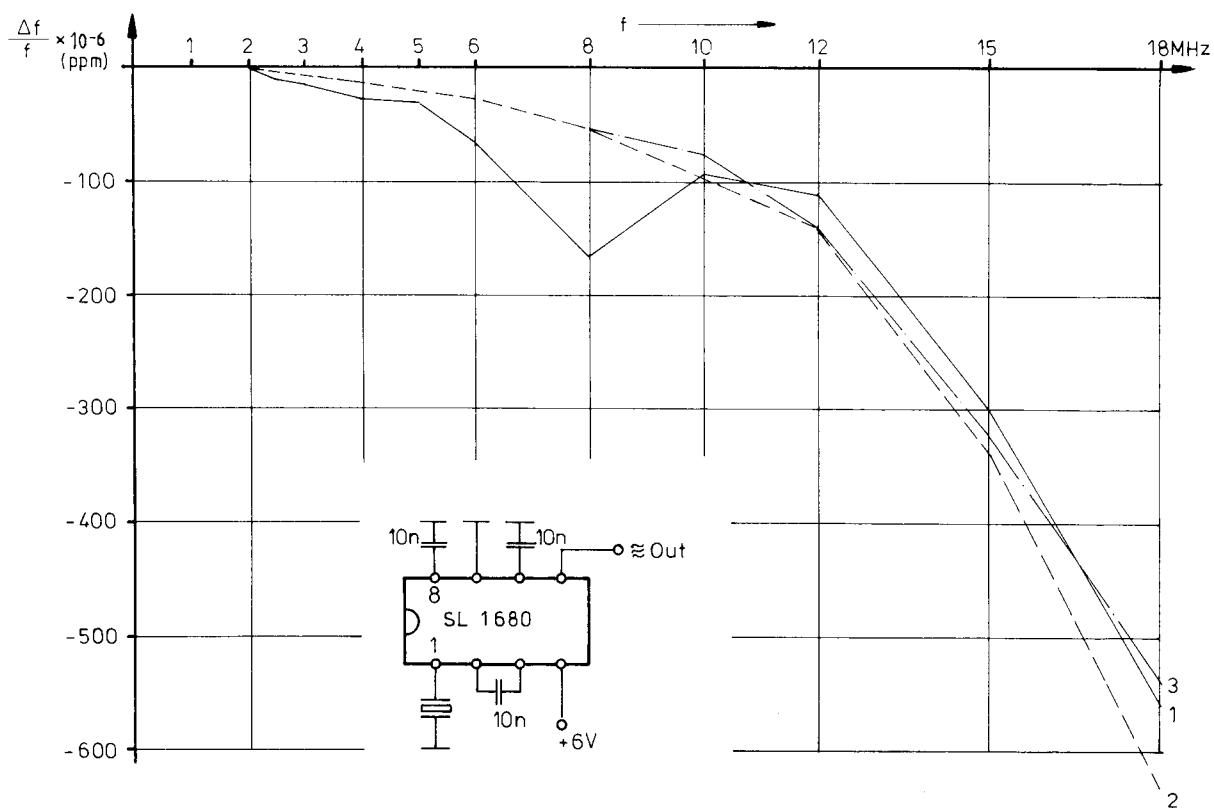
The oscillator can be pulled by several hundred ppm with the aid of the output circuit. If the circuit is aligned for maximum output voltage, a frequency error of between  $-760$  ppm and  $+210$  ppm with respect to the crystal frequency will be obtained when using crystals of between 800 kHz and 20 MHz. These frequency errors can always be corrected by detuning the output circuit. The required detuning was usually in the vicinity of the upper limit of stable commencement of oscillation, but this could possibly be improved by altering the compensation capacitors.

The frequency shift on varying the operating voltage between 5 V and 24 V amounted up to  $\pm 5$  ppm. The DTL/TTL logic divider did not operate satisfactorily with any of the ten samples tested.

### **Plessey 680/1680**

This is a series-resonance oscillator (emitter crystal) with AGC for the frequency range of 100 kHz to 100 MHz (an older data sheet listed: 150 MHz). The drive level of the crystal is very low at  $0.5 \mu\text{W}$ . The feedback link is broken to allow insertion of a resonant circuit in the case of overtone oscillators. A disadvantage of the series-resonance oscillator: it is true, that the crystal is grounded at one side, but if the frequency is to be pulled slightly, it will be necessary to isolate the crystal, or pulling trimmer from ground. Both are inadvisable, and can be avoided when using a parallel-resonance oscillator. In practice, it was also found that approximately 20 % of the integrated circuits were not suitable since they were very temperature dependent. In addition to this, the data sheets list differing pin connections (!!).

The measuring results are shown in **Figure 26**. The samples operated reliably up to 18 MHz. The frequency error with respect to the series-resonance frequency increased continuously and attained values between  $-540 \times 10^{-6}$  and  $-630 \times 10^{-6}$  at 18 MHz. At higher frequencies, especially in conjunction with overtone crystals, a resonant circuit should be inserted between pin 2 and pin 3 so that an impedance matching of the stages can be made using a suitable coil tap.



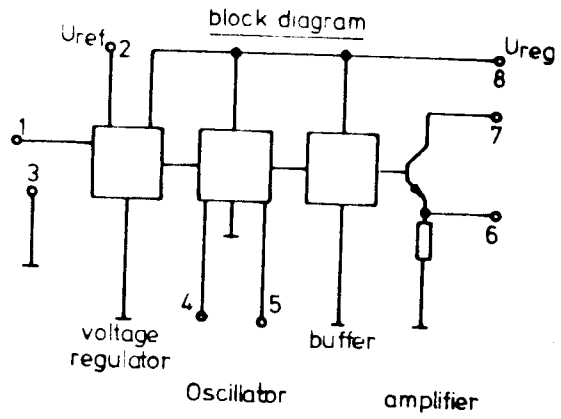
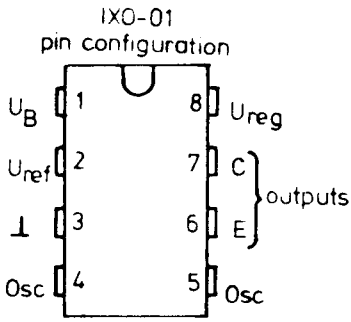
**Fig. 26: Frequency error with SL 1680 (Plessey)**

**KVG IXO-01**

Recently, KVG offers a special crystal oscillator IC of their own manufacture (**Figure 27**). This is a parallel-resonance oscillator using a Darlington-Colpitts oscillator similar to that discussed in section 2. The IC also has an extensive voltage-stabilizer circuit, which is also externally accessible, as well as a multi-stage buffer amplifier. This IC is accommodated in an eight-pin Cerdip-case. Two alternative outputs can be used: an open collector to which a resonant circuit can be connected (an output voltage of more than 5 V (peak-to-peak) across 2 kΩ can be gained, or a low-impedance emitter output ( $R_L \geq 200 \Omega$ ). The oscillator operates in the whole fundamental range up to 30 MHz (even without external circuit), and up to 60 MHz with 3rd overtone crystal.

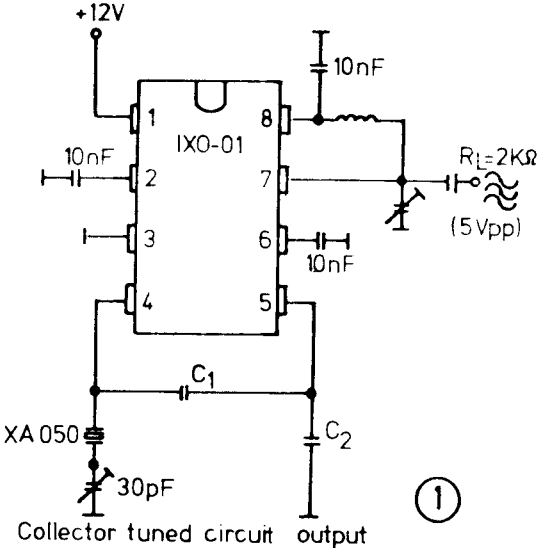
The external circuitry of the oscillator can be made in a similar manner as given in Figure 5 by placing a capacitance of approximately 40 pF in series with the crystal. In the case of overtone crystals,  $C_2$  is replaced by a resonant circuit. Further details can be taken from the original data sheet.

Suitable crystals for the whole frequency range are available (specification XA-050 at KVG) especially for this oscillator. This means that all problems in specifying the crystal are avoided.

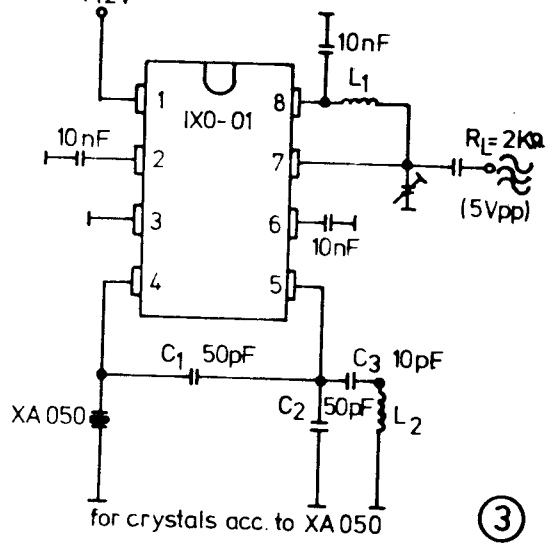


Application notes

Fundamental mode oscillators 1MHz....30MHz



overtone oscillators up to 60MHz



Frequency < 5MHz  $X_{C_1} = X_{C_2} = 200\Omega$

Frequency > 5MHz  $X_{C_1} = X_{C_2} = 100\Omega$

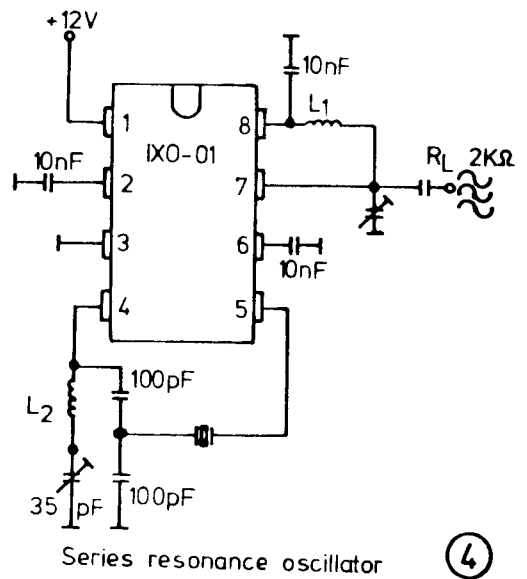
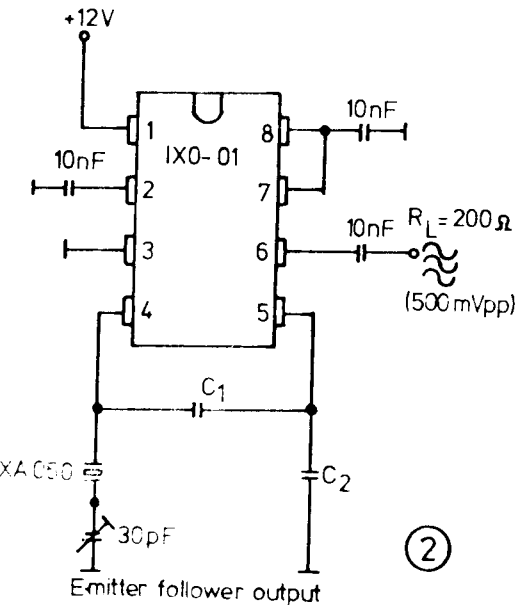


Fig. 27

		Test conditions
Supply voltage	9.5 V ... 20 V	
Current drain	< 10 mA	U <sub>B</sub> = 12 V
External load of stabilized voltage U <sub>reg</sub> (about 7.5 V at pin 8)	< 20 mA	U <sub>B</sub> = 12 V
Stability of U <sub>reg</sub>	$\frac{\Delta U}{U} < 0.5 \text{ ‰ per mA}$	U <sub>B</sub> = 12 V
Temperature coefficient of U <sub>reg</sub> (pin 8)	< 0.5 mV per deg C	U <sub>B</sub> = 12 V
RF output voltage	> 5 V p-to-p	10 MHz, circuit no. 1
Frequency stability due to supply voltage change	$\frac{\Delta f}{f} < 5 \times 10^{-8}/\text{Volt}$	U <sub>B</sub> = 9.5 V ... 20 V
Operating temperature range	- 40°C ... + 85°C	

**Table 6: Extract of the data sheet of the IXO-01 crystal oscillator IC**

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