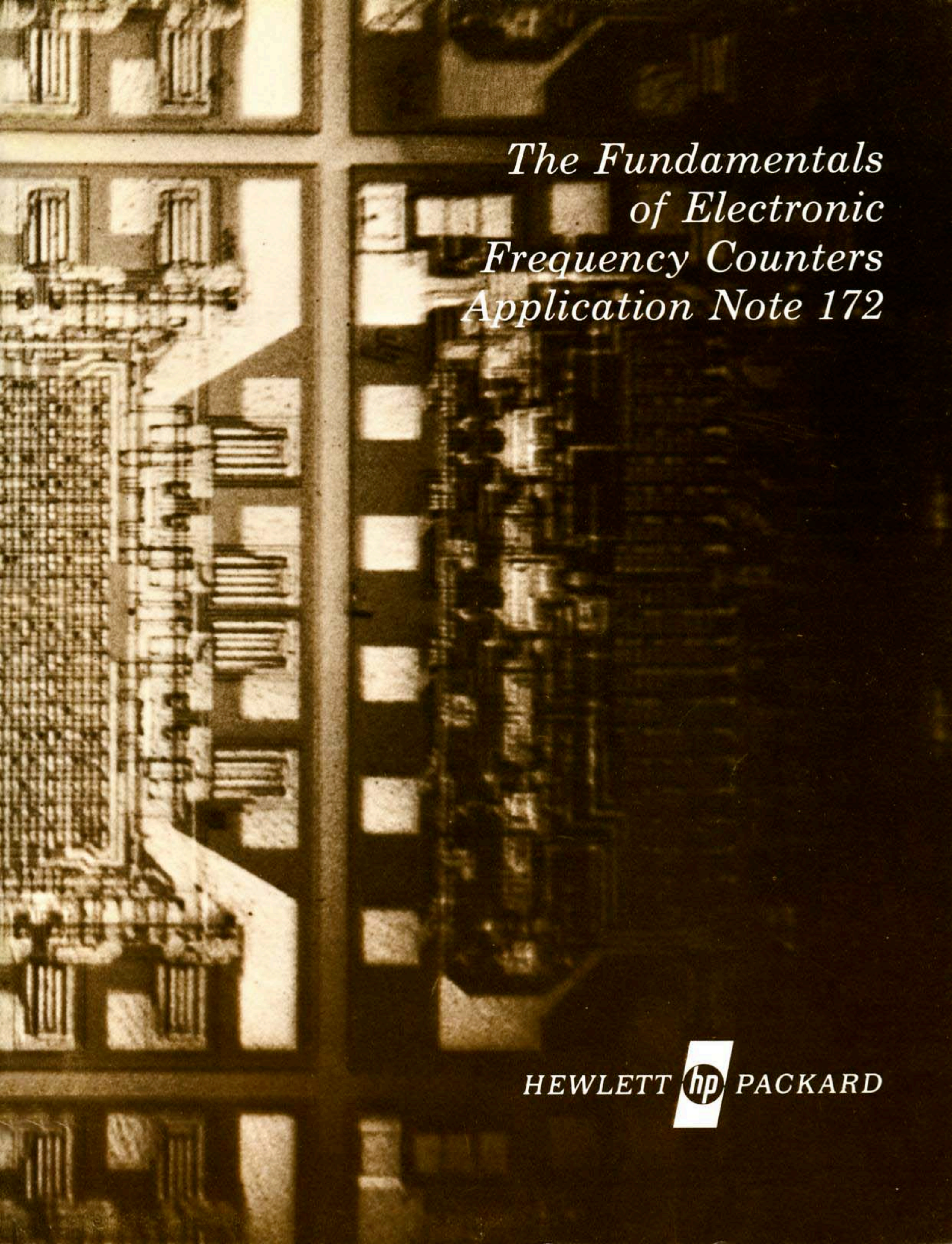
The Fundamentals of Electronic Frequency Counters
Application Note 172



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I. introduction

The digital electronic frequency counter has come a long way since the first versions appeared over two decades ago. Once the luxury of the large metrology labs and some crystal manufacturers, the frequency counter is now common place in laboratories, on production lines, as a service tool and in automatic instrumentation systems. Moreover, counters have become increasingly more versatile and more powerful in the measurements they perform, thereby finding much wider applications. When Hewlett-Packard introduced the 524A in 1952 it was considered a milestone; the counter could measure frequencies up to 10 MHz, or the time between two electrical events to a resolution of one ten millionth of a second, 100 ns. Twenty years later, HP's product line features counters that can measure the frequency of a 10 mV signal at 18 GHz completely automatically, or can resolve time to one ten billionth of a second (100 psec), the same time it takes light to travel one inch!

To a large extent the key to the vast increase in power and flexibility of the electronic counter lies in the component technology evolution. The integrated circuit and the technology of high speed, bipolar, MOS and LSI devices has wrought as great a change in counters as it has in digital communications and computers. Figure 1 shows a 10 MHz LSI MOS chip designed and manufactured at Hewlett-Packard, that contains in its 100 mil^2 area almost one thousand transistors! Compare this to the 524A with its complement of 75 vacuum tubes; small wonder the electronic counter has advanced so far so rapidly.

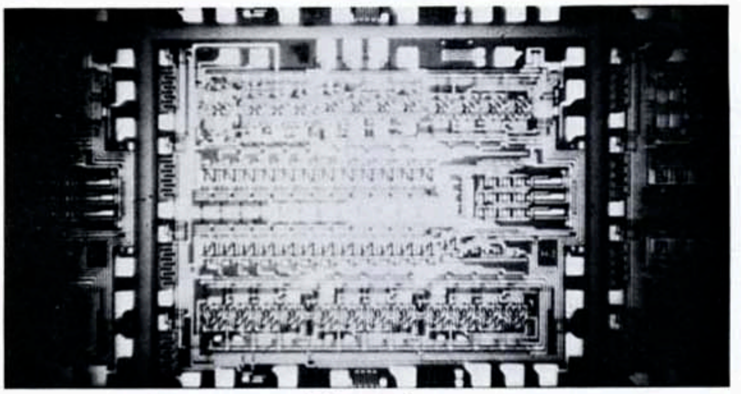


Figure 1

This MOS LSI chip is 0.109 inches square and contains 980 transistors. It operates at a speed of 10 MHz.

With such component power at his disposal, the design engineer has also been able to be more ingenious in reaching his design objectives, thereby offering the user a level of versatility which could not be attained without considerable expense only a few years ago. Keeping this in mind, the primary objective of this book is to introduce the reader to the basic concepts, techniques and considerations employed in electronic counters rather than deal with specific examples, which could mask the underlying principles involved.

II. *fundamentals of electronic counters*

The operation of the conventional digital electronic frequency counter is best understood by describing how the counter performs a frequency measurement. If n is the number of cycles of a signal that occurs in a time period t, the average frequency f of that signal over the time period t is given by

$$f = \frac{n}{t} \tag{1}$$

The conventional counter measures the frequency f by accumulating the number of cycles n of the input signal that occurs over the time period t. The basic counter elements necessary to perform this measurement are shown in Figure 2.

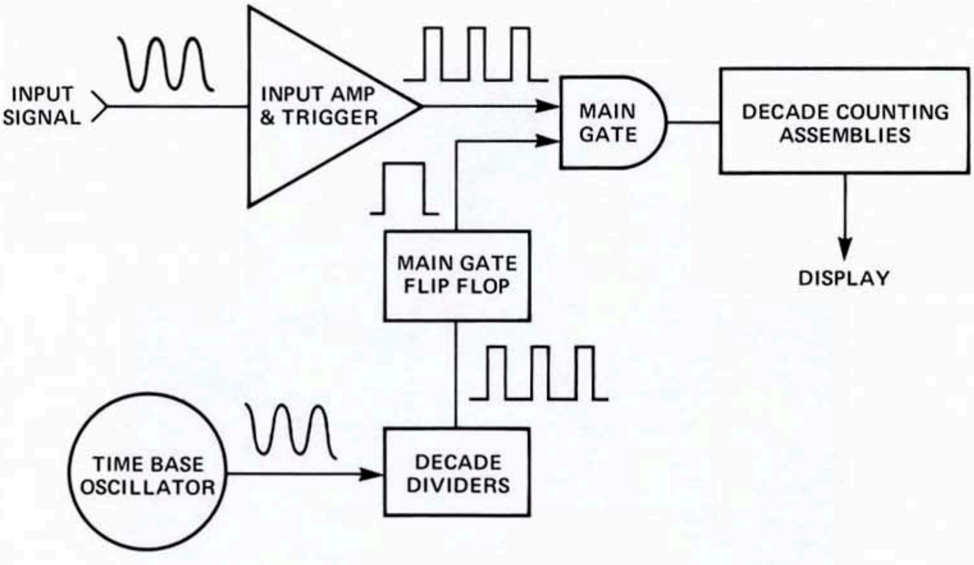


Figure 2

Basic Elements of Conventional Frequency Counter

Input Amplifier and Trigger — essentially conditions the input signal to a form that is compatible with the internal circuitry of the counter. As Figure 2 indicates, the output of the amplifier/trigger is a pulse train where each pulse corresponds to one cycle or event of the input signal.

Time Base Oscillator — is that element of the counter from which the time t of Equation (1) is derived. From Equation (1) it may be seen that the accuracy with which t is determined has a significant effect on the measurement accuracy of the frequency f. Consequently most counters employ crystal oscillators with frequencies of 1, 5 or 10 MHz as the time base element.

Decade Dividers — take the time base oscillator signal as the input and provide as an output a pulse train whose frequency is variable in decade steps. The operator can control this frequency with the gate time switch. The time t of Equation (1) is determined by the period of this pulse train.

Main Gate — is the heart of the counter. When this gate is opened pulses from the amplifier/trigger are allowed to pass through. The opening and closing of the main gate is controlled by the decade divider output to the main gate flip-flop.

Decade Counting Assemblies — totalize the output pulses from the main gate and display this total after the gate is closed. If, for example, the gate is open for precisely one second, the decade counting assemblies (DCA's) display the frequency, in Hertz, of the input signal.

Other basic measurements the conventional counter described above can perform include:

Period

Period, the inverse of frequency, can be measured by the counter by reversing the inputs to the main gate. Now the input signal controls the duration over which the main gate is open and the decade divider output is counted by the DCA's. The duration of the count is, of course, one cycle or period of the input signal (see Figure 3).

Unused decades in the decade divider chain can be used to divide the amplifier/trigger output so that the gate remains open for decade steps of the input period rather than a single period. This is the basis for **multiple period averaging**. Period and period averaging techniques are used to increase measurement accuracy on low frequency measurements as explained further in Sections III and IV.

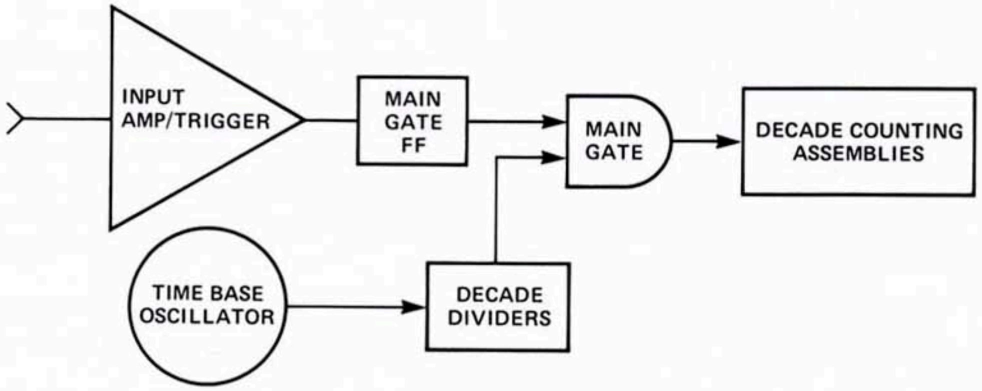


Figure 3
Measuring Period

The roles of the amplifier/trigger and decade divider outputs are reversed in measuring the period. This same configuration also serves for ratio measurements with the second input replacing the time base oscillator.

Ratio

By replacing the time base with a second input of frequency, f2, the same configuration as in Figure 3 can be used to measure the ratio f2/f. For higher resolution the signal at frequency f can be divided in decade steps in a manner identical to multiple period averaging.

Time Interval

Figure 4 shows the configuration for the measurement of time between two events or time interval. The main gate is now opened by the START input and closed by the STOP. The decade divider output is again counted and the display shows the elapsed time between START and STOP signals. The measurement of time interval is considered in more detail in Section V.

Scale

Some counters also provide a scaling mode where an input can be divided in decade steps to provide a low frequency output which is coherent with the input.

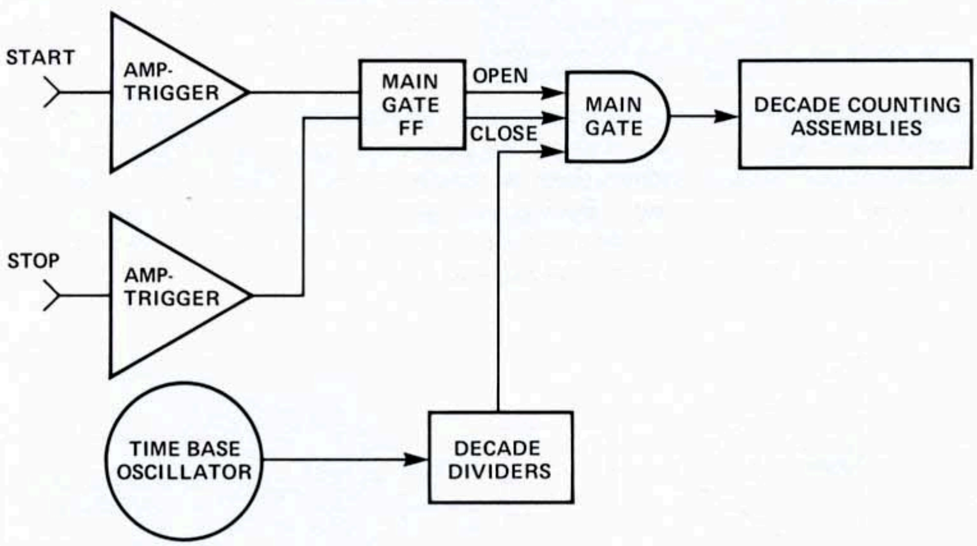


Figure 4

Basic Elements of a Time Interval Counter

III. *more about the basic frequency counter*

A. *input considerations*

The major elements of the input circuitry are shown in Figure 5 and consist of attenuator, amplifier and Schmitt trigger. The Schmitt trigger is necessary to convert the analog output of the input amplifier into a digital form compatible with the counter's decade counting assemblies.

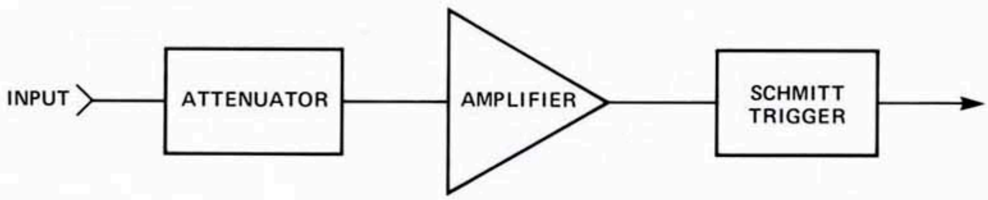


Figure 5

Major Elements of a Counter's Input Circuitry

1. Sensitivity

The sensitivity of a counter is defined as the minimum specified input signal that can be counted. Sensitivity is usually specified in terms of the RMS value of a sinusoidal input. For pulse type inputs, therefore, the sensitivity is $2\sqrt{2}$ of the specified value.

The amplifier gain and the voltage difference between the Schmitt trigger hysteresis levels determine the counter's sensitivity. At first glance it might be thought that the more sensitive the counter input, the better. This is not so, however, since the conventional counter has a broadband input and with a highly sensitive front end, noise can cause false triggering. Optimum sensitivity is largely dependent on input impedance since the higher the impedance the more susceptible to noise and false counts the counter becomes. For 1 MΩ inputs, 100 mV is considered a typical sensitivity while for 50Ω inputs sensitivities to 10 mV are practical.

Inasmuch as the input to a counter looks like the input to a Schmitt trigger we refer to the hysteresis levels of the counter's input. The separation between these levels is the peak-peak sensitivity of the counter. To effect one count in the counter's decade counting assemblies, the input must cross both the upper and lower hysteresis levels. This is summarized by Figure 6.

2. AC-DC Coupling

As Figure 7 shows, AC coupling of the input is almost always provided to enable signals with a DC content to be counted.

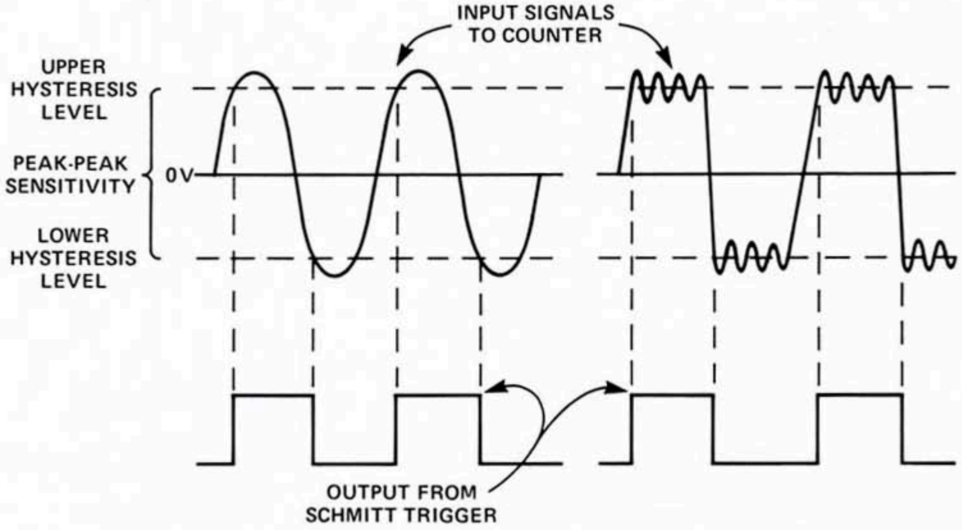


Figure 6
Input Characteristics

To effect a count the signal must cross through both the upper and lower hysteresis levels. Thus in (b), the "ringing" on the input signal shown does not cause a count.

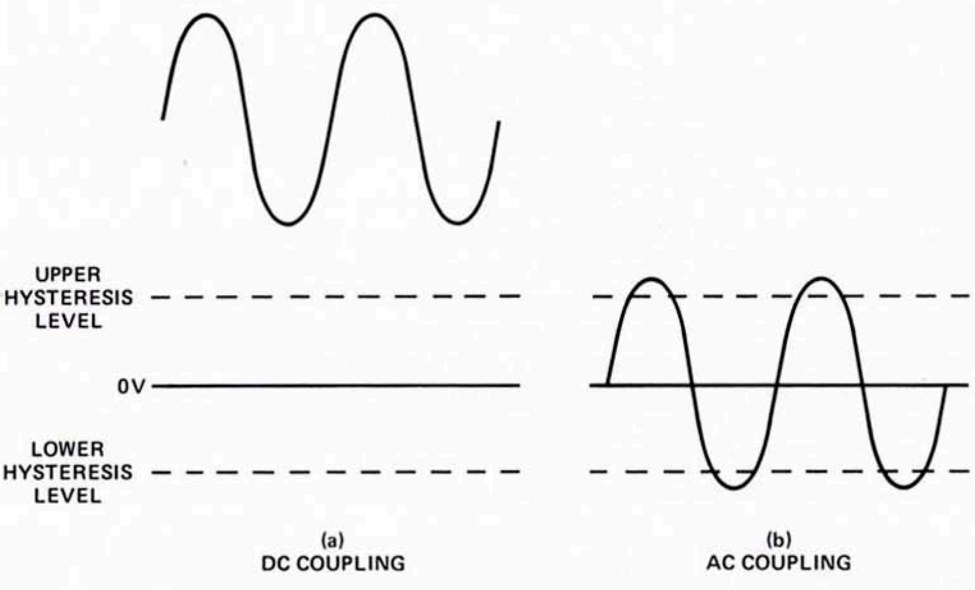


Figure 7
AC-DC Coupling

An input signal with the DC content shown in (a) would not be counted unless AC coupling, as shown in (b), was used to remove the signal's DC content.

3. Trigger Level

In the case of pulse inputs AC coupling is of little value if the duty cycle is low. Moreover, AC coupling should not be used on variable duty cycle signals since the operator has little idea where his signal levels are in relation to ground at the amplifier input. The function of the trigger level control is to shift the hysteresis levels above or below ground to enable positive or negative pulse trains respectively, to be counted. This is summarized in Figure 8.

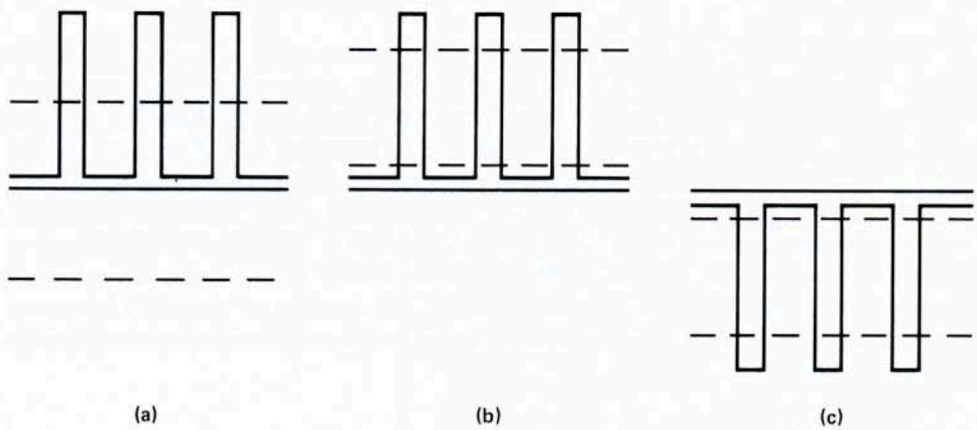


Figure 8
Trigger Level Control

The signal (a) will not be counted. Using the trigger level control to shift the hysteresis levels above ground (b), enables a count. For negative pulse trains (c), the hysteresis levels can be moved below ground.

Many counters provide a three position switch level control with the "preset" position corresponding to Figure 8(a), a position normally labelled "+" corresponding to Figure 8(b) and "-" for the Figure 8(c) case. The more sophisticated counters provide a continuously adjustable trigger level control, adjustable over the whole dynamic range of the input. This more flexible arrangement ensures that any signal within the dynamic range of the input and of an amplitude consistent with the counter's sensitivity can be counted. It should be noted, however, that the majority of inputs can be handled with the simplified 3 position level switch.

4. Dynamic Range

The dynamic range of the input is defined as the input amplifiers linear range of operation. Clearly, it is not important for the input amplifier of a frequency counter to be absolutely linear as it is in an oscilloscope for example (this is **not** the case for time interval, see Section IX). Moreover, with a well designed amplifier exceeding the dynamic range will not cause false counts, however, input impedance could drop and saturation effects may cause the "pulse pair resolution" or the amplifier speed of response to decrease. Of course, all amplifiers have a damage level and protection is usually provided. Conventional protection often fails, however, where high speed transients (e.g., at turn-on of a transmitter) and low impedance 50Ω inputs are involved. To this end, several of the Hewlett-Packard counters (5327 family and 5303B) employ high speed fuses in addition to the conventional protection to further protect wideband 50Ω amplifiers.

5. Attenuators

It is, nevertheless, not good practice to exceed the dynamic range of the input. To avoid this on larger level signals attenuators are provided. The more sophisticated inputs with wide dynamic range usually employ step attenuators with attenuation positions such as X1, X10, and X100. (The actual attenuation values used depend on the dynamic range of the input. Thus a two position setting of X1 and X20 is indicative of a wider dynamic range than the example above.) Another variation is a variable attenuation scheme. This is mandatory for low dynamic range inputs but it also provides the additional advantage of variably attenuating noisy signals to minimize the noise while maintaining maximum signal amplitude.

6. Input Impedance

For frequencies up to around 100 MHz, a 1 MΩ input impedance is usually preferred. With this impedance level the majority of sources connected to the input are not loaded and the inherent shunt capacity of about 20 pF has little effect. As noted earlier, for noise considerations, sensitivities of 50 mV to 100 mV are preferred. Beyond about 100 MHz, however, the inherent shunt capacity of high impedance inputs rapidly reduces input impedance. For this reason 50Ω impedance levels, which can be provided with low shunt capacity, are preferred. Sensitivities of 10 mV to 25 mV are considered optimum with 50Ω inputs. A sensitivity of 1 mV for example, is possible, of course, however the user must pay a premium for this and noise problems can occur.

7. Automatic Gain Control

Automatic Gain Control (AGC) may be thought of as an automatically adjustable sensitivity control. The gain of the amplifier-attenuator section of the input (see Figure 5) is automatically set by the magnitude of the input signal.

A tradeoff exists between the speed of response of the automatic gain control and the minimum frequency signal that can be counted. For this reason the lower frequency limit for AGC inputs is usually around 50 Hz. AGC inputs, therefore, are useful primarily for frequency measurements only.

AGC provides a certain amount of operator ease since the sensitivity control is eliminated. A second advantage of AGC is where the input signal amplitude is varying. Figure 9 shows an example of this. The output of a magnetic transducer is shown as the frequency as the rotating member reduces from 3300 Hz to 500 Hz. The signal level decreases from 800 mV to 200 mV and the noise decreases from 300 mV to 50 mV. If the sensitivity were set to count the lower level signal, any attempt to count the higher level signal at 3300 Hz would result in false counts due to the 300 mV noise level. AGC eliminates this problem since the noise shown on the high level signal is attenuated, along with the signal, to a level where it does not cause false triggering. This assumes, of course, that the trigger level is appropriately set in the first place.

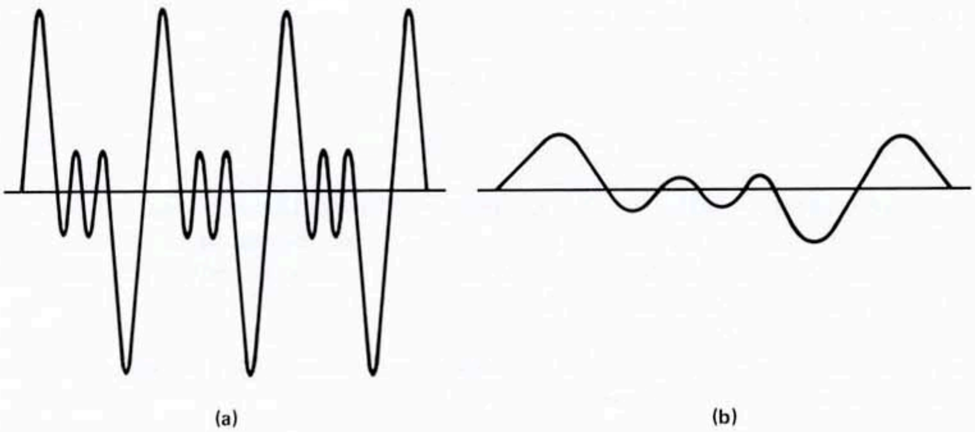


Figure 9

Output of a magnetic transducer at 3300 Hz (a) and 500 Hz (b). Without AGC it would be impossible to measure this changing frequency since a sensitivity setting to measure the lower frequency signal would result in erroneous counts due to noise at the higher frequencies.

B. oscillator characteristics

At the heart of the counter is the time base oscillator. Almost all counters employ the quartz crystal as the oscillating element. There are three basic types of crystal oscillators used in counters with the main difference between them being the steps taken to minimize errors caused by change in oscillator frequency with temperature.

The basic types of crystal oscillators are:

1. Room Temperature. Such oscillators will typically vary by about $\pm 5 \times 10^{-6}$ from center frequency with 0°C to 50°C temperature change.

2. Temperature Compensated. Aften called TCXO's, these oscillators include temperature compensating elements which provide typically five times improvement over the uncompensated type. Thus specifications of $\pm 1 \times 10^{-6}$ over 0°C to 50°C* temperature range are not uncommon.

3. Oven Controlled. These oscillators go one step further and house the crystal in an oven to minimize temperature changes. The simplest is the on-off type, which like the normal house thermostat, operates with the element being turned fully on or fully off. The more sophisticated proportional oven provides a heating that is proportional to the temperature differential between the inside and outside of the oven. This is the preferred method since it minimizes cycling of the oven's internal temperature and consequently minimizes changes in oscillator frequency. The highest quality oscillators employ double proportional control ovens, where one proportional oven is inside a second. Inner oven temperature can be controlled to .01% with this method, which minimizes changes in oscillator frequency due to transients in the ambient temperature in which the oscillator is working. A well designed single oven oscillator will provide a frequency stability of 5×10^{-8} over the 0°C to 50°C temperature range, while for the same range a double oven oscillator can provide up to ten times improvement. These types of oscillators are often specified in terms of rate of frequency change with temperature; e.g., 5×10^{-11} per °C from 0°C to 50°C.

Apart from temperature effects there are other factors which effect oscillator frequency. These are:

a. Line Voltage. The amount of change in oscillator frequency with power line voltage depends on the degree of regulation that the oscillator power supply possesses. A well designed supply for a high quality oven oscillator will provide frequency stability in the order of 5×10^{-10} for 10% changes in line voltage. On the other hand, a room temperature oscillator provides stabilities in the order of 1×10^{-7} for the same line voltage change. Regulation better than this is essentially unnecessary since the inevitable temperature variations will mask line voltage effects.

b. Aging Rate. This phenomenon, which is also called long term stability, is exhibited by all crystals and refers to the slow cumulative drift in oscillator frequency with time. The magnitude of the aging rate is essentially dependent on the quality of the manufactured crystal, with the best crystals (used in double oven oscillators) providing aging rates of better than 5×10^{-10} per day (1.5×10^{-8} per month). On the other hand, a room temperature oscillator is typically in the order of 3×10^{-7} per month. Per month aging rates are usually quoted for these types of oscillators since temperature changes of only a degree or so can change oscillator frequency more than a day's aging. It usually takes 24 hours or more after turn on for the oven oscillator to achieve its specified stability. For this

*Frequency change with temperature is typically non linear thus one cannot specify a rate of change of frequency with temperature.

reason most counters employing oven oscillators provide a feature whereby the oscillator is always on if the power line is connected, irrespective of whether the counter is turned on or not. Keeping the counter connected to the power line, therefore, avoids the warm-up process.

c. Short Term Stability. More properly referred to as fractional frequency deviation, this term refers to the inevitable noise (random frequency and phase fluctuations) generated by the oscillator. Since this noise is spectrally related, any specification of short term stability must include the averaging or measurement time involved. The effect of this noise usually varies inversely with measurement time and to be meaningful the specification should include a measurement time that is commonly used with the counter. Thus specifications involving ten minutes of measurement time, for example, are totally meaningless. With quoted averaging time the specification of short term stability essentially specifies the uncertainty due to noise in the oscillator frequency over the quoted time. Short terms of better than 5×10^{-11} for one second averaging can be achieved with oven oscillators. This specification is rarely quoted for room temperature oscillators since temperature effects again predominate.

The total oscillator error is the sum of the individual errors described above. This error may or may not be significant for a given oscillator depending on the application involved. Suffice to say that careful selection of the appropriate oscillator for a given application is highly desirable.

The chart below summarizes the oscillator characteristics described, utilizing typical specifications of well designed oscillators.

	Room Temperature	Temperature Compensated	Single Prop Oven	Double Prop Oven
Aging Rate	$<3 \times 10^{-7}$/mo	$<1 \times 10^{-7}$/mo	$<3 \times 10^{-9}$/ 24 hrs.	$<5 \times 10^{-10}$/ 24 hrs.
Short Term			$<2 \times 10^{-10}$ (RMS)/1 sec average	$<5 \times 10^{-11}$ (RMS)/1 sec average
Temperature	$<5 \times 10^{-6}$, 0°C - 50°C	$<5 \times 10^{-7}$, 0°C - 50°C	$<2 \times 10^{-10}$/°C 0°C - 50°C	$<5 \times 10^{-11}$/°C 0°C - 50°C
Line Voltage	$<1 \times 10^{-7}$/10% line change	$<5 \times 10^{-8}$/10% line change	$<5 \times 10^{-10}$/10% line change	$<1 \times 10^{-10}$/10% line change

Figure 10

Summary of Crystal Oscillator Characteristics

C. sources of measurement error

The major sources of measurement error are the ±1 count ambiguity, the time base error and trigger error. These are discussed in turn below.

1. ±1 Count Ambiguity

Since the signal input to the main gate of the counter and the clock input are not coherent, an inherent ±1 count ambiguity exists in the count accumulated in the decade counting assemblies. This is illustrated by Figure 11.

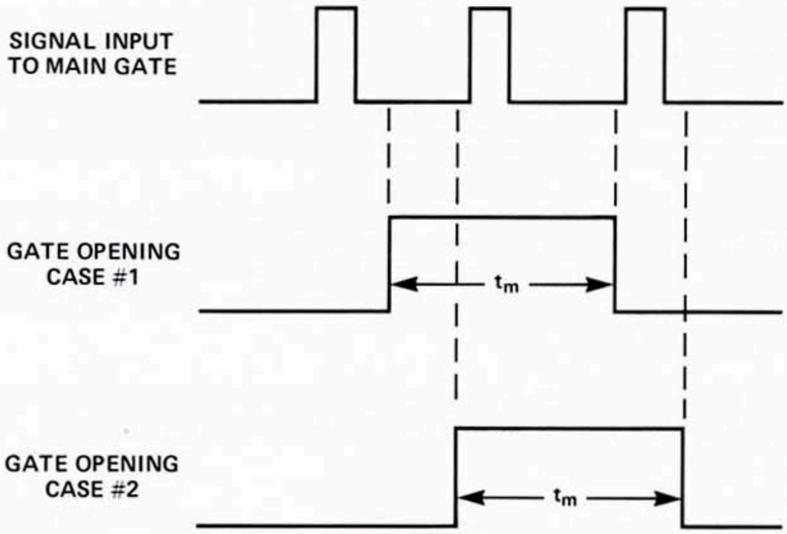


Figure 11

The main gate is open for the same time t_m in both cases. Incoherence between the clock and the input signal can cause two valid counts which for this example are 1 for case #1 and 2 for case #2.

The error caused by this ambiguity is in absolute terms, ±1 of the accumulated count. For a frequency measurement the signal counted is the input signal of frequency f_{in}. Thus the relative error is given by

±1 count error, relative frequency measurement error

$$\frac{\Delta f}{f} = \frac{\pm 1}{f_{in}} \tag{2}$$

For period measurement, on the other hand, the signal counted is the internal time base clock of period t_C. Hence the relative error becomes

±1 count error; relative period measurement error

$$\frac{\Delta T}{T} = \frac{\pm t_C}{T_{in}} \qquad (3)$$

a. Main Gate Requirements

The ±1 count error described above assumes the main gate itself does not contribute any error. As with any gate, however, the main gate does exhibit propagation delays and takes finite times to both switch on and off. Any differential between the times taken for the main gate to switch on and off show up as uncertainties in the length of time the gate is open. This uncertainty in turn translates into a measurement error that increases the ±1 count. However, provided this uncertainty is substantially less than the period of the highest frequency counted, this error is not appreciable. A 500 MHz signal has a period of 2 nsec and for this error to be insignificant uncertainty in the gate must be substantially less than 1 nsec. This implies that 1 GHz devices are needed in the gate circuitry, and this is a major reason why one sees few counters about today claiming true 500 MHz operation.

2. Time Base Error

Any error in the time base oscillator directly translates itself into a measurement error. Thus if the total of all the errors described in Section B amount to 1×10^{-6}, the total error contributed by the time base in the measurement of a 10 MHz signal is $1 \times 10^{-6} \times 10^{7} = 10$ Hz. Similarly, for the measurement of a 100 msec period, the error would be $1 \times 10^{-6} \times 10^{-1} = 1 \times 10^{-7}$ or 100 nsec.

3. Trigger Error

Noise on the input signal will cause uncertainties in the point at which the Schmitt trigger switches. Provided the noise is not large enough to cause false triggering (i.e., produce more pulses out of the Schmitt trigger than input cycles to it) no error is introduced in a frequency measurement. This is so since this error is absorbed by the ±1 count. For period measurements, however, this uncertainty produces like error in the time the gate is open since it is this signal that controls the gate. It can be shown that with essentially low frequency noise and a signal to noise ratio of 40 dB, the resultant worst case trigger error is 3.2×10^{-3}. Thus the trigger error in the measurement of the period of a 1 kHz signal is $3.2 \times 10^{-3} \times 10^{-3} = 3.2\ \mu$sec worst case. For 60 dB S/N worst case error is 3.2×10^{-4} while for a 20 dB S/N signal it is 3.2×10^{-2}. Figure 12 summarizes this. It should also be noted that pulse input, with their faster slew rate, produces less trigger error than a sinusoidal input with the same level of noise.

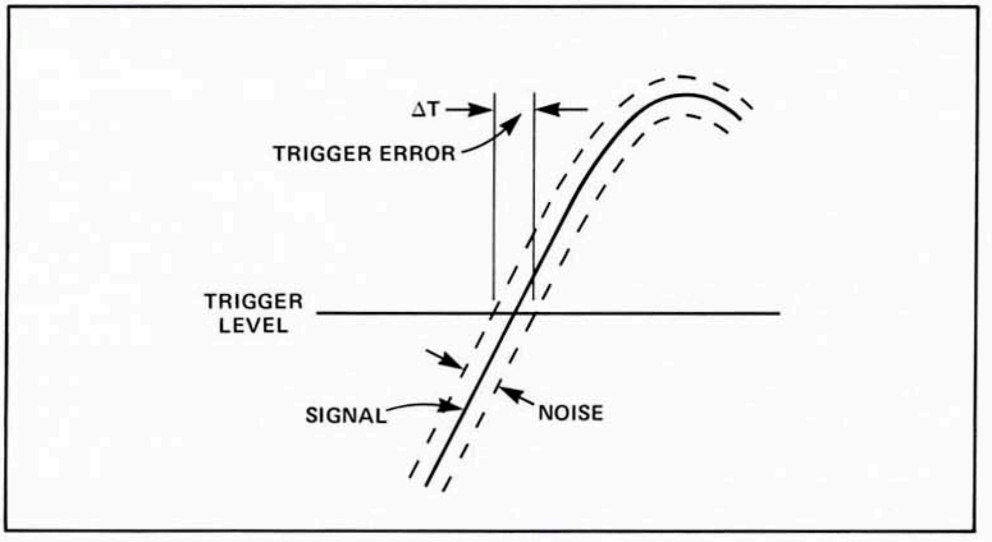


Figure 12
Trigger Error

Noise on the input signal causes an uncertainty ΔT in the point at which triggering occurs. In a period measurement this causes an error known as trigger error since it is this signal that controls the main gate. Trigger error is not a factor with frequency measurements.

In summary then, the accuracy statement for frequency measurement is given by

frequency measurement error = ±1 count ± time base error (4)

As noted earlier, all the factors that contribute to time base errors should be taken into account. For a period measurement the accuracy statement becomes

period measurement error = ±1 count ± trigger error ± time base error (5)

The error in (5) can be reduced by **period averaging**. This simply means that instead of measuring one period, n periods are measured (n is usually in decade steps for counters that offer it). The accuracy statement now becomes

multiple period averaging; measurement error =

$$\frac{\pm 1 \text{ counter} \pm \text{trigger error}}{n} \pm \text{time base error} \quad (6)$$

where n is the number of periods averaged.

It should be noted that in both Equations (5) and (6) the ±1 count refers to the counted clock or time base while in (4) the ±1 count is that of the input signal.

At first glance, when one compares Equation (4) with Equations (5) and (6), the accuracy of frequency measurements look to be better than that of period measurements, This is, however, not so. Ignoring noise, the frequency content of a continuous wave signal is contained in one cycle of the signal and the resolution one obtains is dependent on the frequency of the counted clock for a period measurement. For a frequency measurement one cycle of the input will produce one count. Thus for a given measurement time, period measurement is always more accurate than frequency measurement as illustrated by Figure 13.

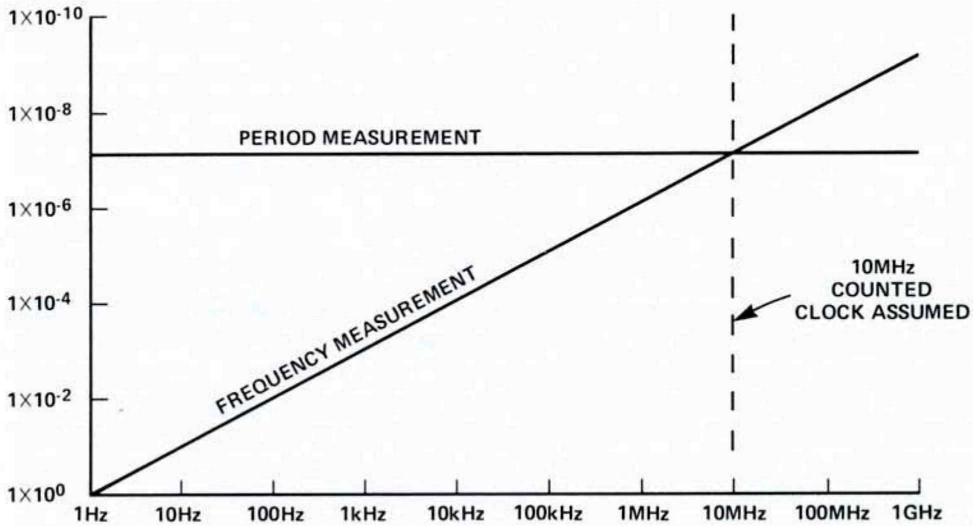


Figure 13

Comparing the accuracy of period and frequency measurements. This assumes a noiseless input signal and shows that for a given measurement time period measurement is always more accurate than frequency measurement for all frequencies less than that of the counted clock. This statement holds true for inputs that have noise associated with them although the period measurement accuracy shown here is degraded.

D. prescaling — increasing the frequency response

It was noted in Section C that apart from the input amplifier trigger, the elements that determine the counter's upper frequency count are the speed of the main gate and decade counting assemblies. By dividing the input signal prior to application to the main gate, the speed limitations of these two elements can be reduced and the upper frequency limit of the counter increased. This is known as prescaling. If the prescale factor is n, the main gate must remain open n times longer to ensure the correct count in the DCA's. Thus prescaling is a tradeoff, the frequency response is increased by a factor of n but so is the measurement time. In addition, less expensive gate and DCA circuitry is needed, at the expense of an additional divider.

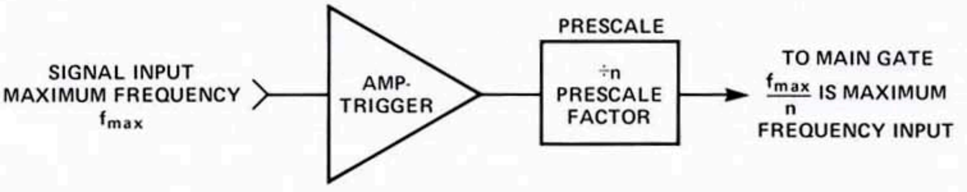


Figure 14

The input to a prescaling counter with prescale factor of n.

At the time of writing, the state-of-the-art of component technology limits the direct count frequency response to about 500 MHz. Prescaled 500 MHz counters are available at substantially lower cost than their direct count counterparts since only the front end need operate at this speed. Prescaling is widely used and is perfectly satisfactory for the digital measurement of average frequency — the majority of frequency measurements for which counters are used. There are, however, some limitations namely, for a given measurement time resolution is n times less than that of a direct counter (n is prescale factor), short measurement times (e.g., 1 μsec) are not available and the prescaled counter cannot totalize at rates of the upper frequency limits indicated.

E. normalizing and preset counters

The **normalizing counter** accepts a signal of frequency f and displays a result y where y = af and "a" is a numerical constant. This type of counter finds wide use in industrial applications for measurements such as flow and RPM. The input signal to the counter is derived from a transducer which measures the physical parameter y and produces a signal f, i.e., the constant "a" is the transducer's transfer function.

A simple example is that of a flow meter which outputs 100 pulses per gallon. If the flow rate being measured is 100 gallons per minute, the number of output pulses per minute is 10,000. The output frequency as measured by a conventional counter is 166.66 Hz. In a normalizing counter the constant "a" would be set to 60/100 or 0.6 and the readout would then be directly in gallons per minute. The normalizing factor, 0.6 in this case, is set into the counter via thumbwheel switches that are normally located on the front panel of the counter.

With the availability of relatively inexpensive integrated circuit memories the equation y = af is now solved by digital processing means. For this reason such counters are sometimes referred to as "computing counters". (It should be noted that these computing counters should not be confused with the Hewlett-Packard 5360A Computing Counter.) Prior to the ready availability of IC's, the normalizing function was performed by varying the gate time by non decade steps via specially designed time base divider assemblies.

Preset counters provide an electrical output when the display exceeds the number that is preset into the counter. As with the normalizing counter, the preset number can usually be set into the counter via thumbwheel switches available on the front panel. Preset counters are also extensively used in industrial applications where the electrical outputs can be used to control external equipment. Typical examples include batch counting and limit sensing for engine RPM measurements.

Since both preset and normalizing counters find use in the same applications area, it is not uncommon to find counters with both capabilities, normalization and presettability, available.

IV. *period measuring frequency counters*

The availability of complex integrated circuits has given birth to a new class of counters; the period measuring frequency counter. The measurement of period rather than frequency offers several distinct advantages to be discussed below. The overriding disadvantage to date has been that the answer was displayed in terms of period rather than frequency. It is now a relatively simple matter to mathematically process the period information in the DCA's and display directly the frequency information.

The primary advantage of period counting is summarized in Figure 13. The resolution of a period measurement is independent of the frequency measured while for frequency counting, resolution is directly proportional to the frequency. This is particularly advantageous at low frequencies; for example, a 10 Hz signal can be measured to no better than 10% in one second in a frequency counter, no matter how good it is. Conversely, with a period counter and a 10 MHz clock, the same 10 Hz signal can be measured to a resolution of 1×10^{-7} in one second! Thus period counting has an inherent advantage in that the full resolving capability of the instrument is always utilized whereas for frequency counting the same is true at the top end of the frequency range only.

A second advantage of period counting is that it potentially provides the user with the ability to control the main gate in real time. This is so since the signal itself controls the gate. Conversely, in frequency counting the gate is controlled by the time base and the operator has little, if any, control of when the gate opens; all he knows is that at some undetermined point in time the gate will open and accumulate counts that occur from the input, and then close again a precise interval of time after opening, to display the average frequency of the input signal over the time the gate was open.

The ability to control the gate in real time, we refer to as **arming**. A period counting counter can be externally armed as shown schematically in Figure 15. While arming is not needed for most applications it can greatly simplify some difficult measurement problems.

Use of external arming to measure **pulsed RF** is shown in Figure 16.

Of course, arming with such counters can be done automatically and many period measuring frequency counters offer only the automatic mode. With automatic arming the measurement in Figure 16 would start with the first input cycle of the pulsed RF signal.

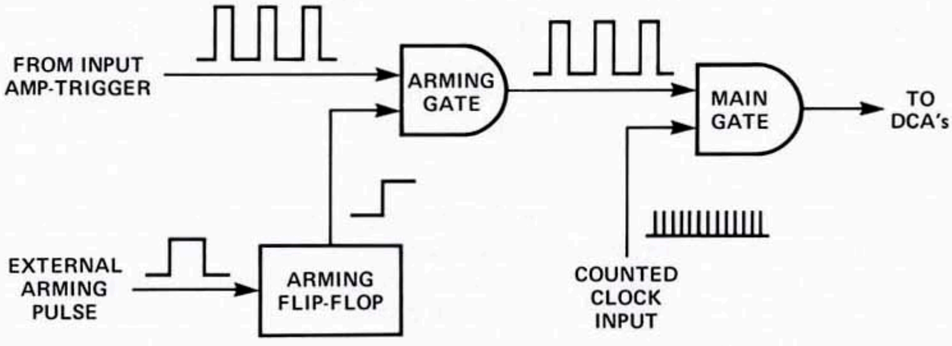


Figure 15

Externally arming a period measuring counter. The measurement starts with the first input cycle that occurs after arming.

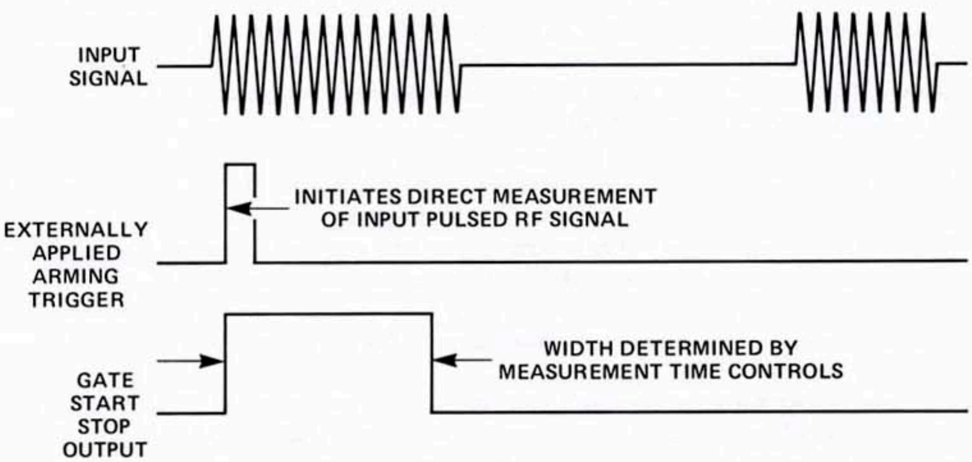


Figure 16

Measuring the frequency of a pulsed RF signal with a period measuring frequency counter via external arming.

The inherent high resolving power of period counting plus the ability to initiate a measurement at any point in real time via external arming gives rise to the concept of **instantaneous frequency measurement**. This allows meaningful measurements on **frequency agile**, **pulse compression** and **Doppler radar** systems. An example is shown in Figure 17 below.

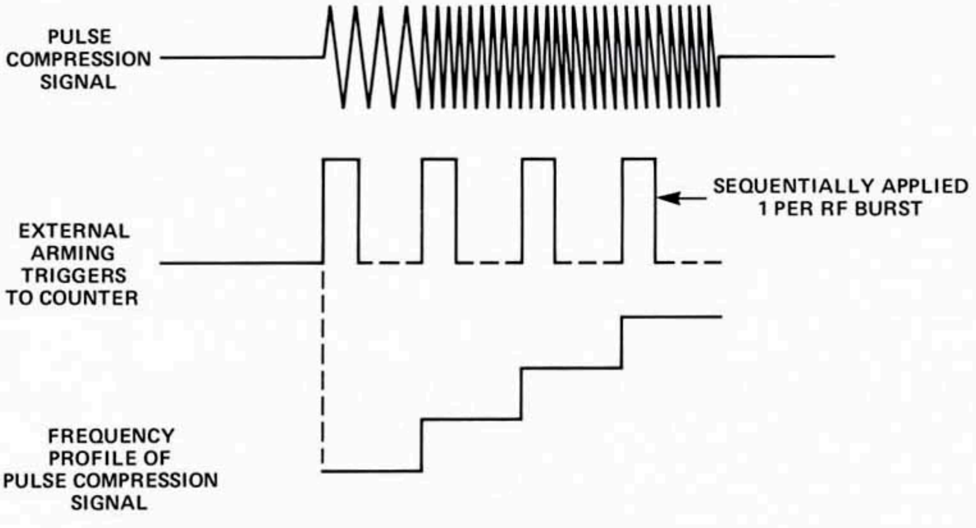


Figure 17
Characterizing a pulse compression system via external triggering of a period measuring frequency counter.

It is theoretically possible to perform the measurements described in Figures 16 and 17 with a conventional frequency counter. If the time base dividers are all preset to 9, the first input clock cycle after arming will open the gate. If the clock period is t_c, the uncertainty in the gate opening is t_c, which can be minimized by using a high frequency clock. However, the conventional frequency counter still suffers from the fact that its resolving power is a function of the input frequency, which is impractical for the measurements described above. In addition, extra divider circuits are needed in the time base divider chain.

In summary, period measurement has the advantage that it utilizes the full resolving capability of the counter over its entire frequency range. In addition, the real time measurement capability of period counting allows measurements on pulsed RF systems and the characterization of such systems via the concept of instantaneous frequency. Other applications for period measuring frequency counters include low frequency measurements (e.g., power line frequency) and the metrology lab where high accuracy can be obtained in conveniently short measurement times. The disadvantage of this type of counter is the additional cost—thus if all one needs is the digital measurement of average frequency, the conventional frequency is perfectly sufficient.

V. time interval

A. introduction

Time interval, the measurement of the time between two events, can be accomplished via the block diagram shown in Figure 4. The main gate is now controlled by two independent inputs, the START input opening the gate and the STOP input closing it. Clock pulses are accumulated for the duration the gate is open and this accumulated count represents the time interval between START and STOP. This is shown by Figure 18.

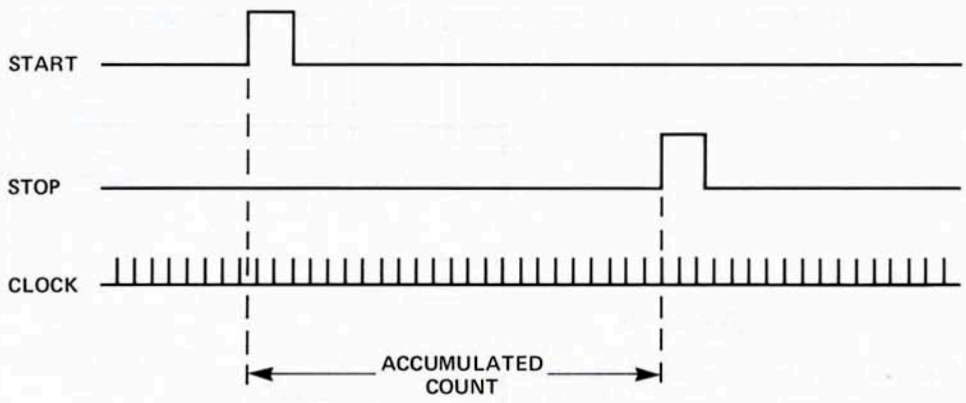


Figure 18

In a time interval measurement, clock pulses are accumulated for the duration the main gate is open. The gate is opened by one event, START and closed by the other, STOP.

The resolution of the measurement is determined by the frequency of the counted clock (e.g., a 10 MHz clock provides 100 nsec resolution). Clearly, the elements within the time interval counter (input amplifier, main gate, DCA's) must operate at speeds consistent with the clock frequency, for otherwise the instrument's resolution would be meaningless. Present state-of-the-art limits resolution to about 10 nsec, although special techniques, to be described later, can be utilized offering substantially improved resolution.

Clock frequencies of 1, 10, 100 MHz, etc. are preferred since the accumulated count, with the appropriate placement of decimal point, gives a direct readout of the time interval. This explains why the conventional time interval counter is at present limited to 10 nsec, a clock frequency of 100 MHz. 1 GHz is beyond reach and a clock frequency of 200 MHz would require some arithmetic processing of the accumulated count in the DCA's to enable time to be displayed directly.

B. input considerations

If the signal inputs to the time interval counter were the clean, sharp pulses depicted by Figure 18, there would be little to consider as far as the input circuitry is concerned. In fact, some special purpose time interval counters are designed solely for use with this type of input, with trigger level permanently set or adjustable via a screwdriver.

In the more general purpose cases, however, time interval is a two dimensional problem. The dimensionality of the time interval measurement is illustrated by the simple example of Figure 19, measuring signal rise time.

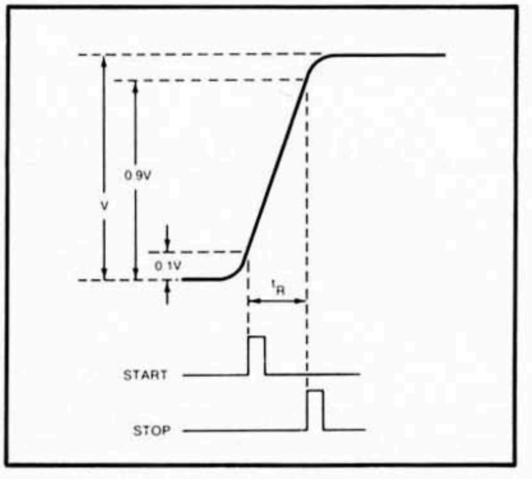


Figure 19

Measuring the rise time t_R by adjusting the trigger levels to the 10% and 90% points of the input amplitude.

The time interval meter must generate a START signal at the 10% amplitude point of the input signal and generate a STOP signal at the 90% point. Clearly this is different from the frequency or period measuring case where the input triggers at the same point on the waveform from cycle to cycle of the input. Inherent in the time interval measurement, therefore, is the dual dimensionality, amplitude and time. It is this dimensionality that places much more stringent requirements on the input amplifier—triggers than those necessary for the measurement of frequency or period.

To take care of the amplitude problem most time interval meters include adjustable trigger level controls for both input channels. With the trigger level set at a certain voltage, V1, the channel produces an output pulse, which is applied to the main gate when the input signal reaches that voltage level, V1. To enable triggering at any point on the waveform, the trigger level is usually made to be adjustable over the entire dynamic range of the input amplifiers. The input amplifiers themselves must, of course, be linear to minimize any distortion effect on the input signal; and to provide full flexibility, a wider dynamic range than a frequency measuring input is needed. Many applications require triggering on the negative going slope of the input signal as well as the positive slope (e.g., pulse width or fall time measurements) and slope controls are added to facilitate this. Input impedance is generally 1 MΩ although for measurement on high speed signals 50Ω is preferred to minimize capacitive loading and reflections due to impedance mismatch in 50Ω systems. Finally, and most obviously, two independent inputs are needed, one for the START channel and one for STOP with provision made so that the two channels may be commoned right at the input. This enables measurements such as risetime to be máde. These then are the essential differences between the inputs of time interval and frequency counters—differences that place far more stringent requirements on inputs designed for the time interval meter.

C. trigger level

Figure 19 emphasizes the importance of accurately setting trigger level, for any error in this setting directly translates itself into a measurement error. Figure 20 illustrates schematically how the trigger level is set.

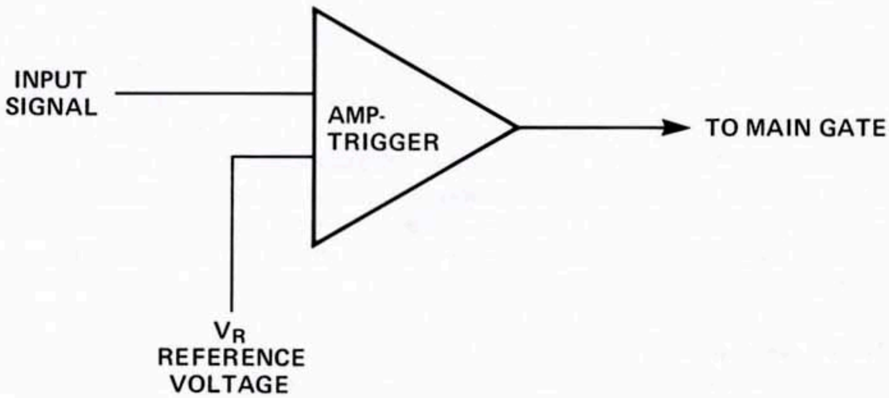


Figure 20

Trigger level is set by varying the reference voltage V_R applied to the second input of the input trigger.

Assuming an idealized trigger circuit setting the reference voltage $V_R = V1$ will cause the trigger to fire when the input signal voltage $V_{in} = V1$. In actual fact the trigger voltage V_t is given by

$$V_t = V_R + \delta + h \qquad (7)$$

where V_R = reference voltage
δ = inherent finite mismatch between elements of the trigger circuit and
h = hysteresis level

There are a variety of ways in which V_t can be determined. For the moment we assume the V_t is determined by measuring V_R (this is, in fact, a technique that is used and is one of the better ones). Equation (7) indicates that simply measuring V_R will cause an error in the amount of $\delta + h$.

The quantity h can be determined simply by measuring the hysteresis level, the minimum signal at which triggering can occur. In fact, with the better designed counters, the quantity h is "bucked out" of the voltage V_R that is available to the user and no calculation is therefore needed. Removal of the hysteresis effect is via potentiometers that are calibrated at the factory and should be regularly checked during the instrument's life.

The magnitude of the quantity δ depends on how well balanced the input amplifier-trigger is. For slower speed inputs (e.g., 10 MHz bandwidth) it is relatively easy to design a well balanced input, however, for high speed inputs finite mismatch becomes a factor. A well designed high speed input will typically have mismatches of no more than 10% of the hysteresis level. Whether this is a factor or not depends on the accuracy required. However, it should be noted that this mismatch can increase with age and temperature variations.

A simple method to determine the relationship between V_t and V_R is to input a slow square wave of accurately known amplitude. Since the peak voltage of the square wave is known, setting the level so the counter is just triggering establishes V_t and the corresponding V_R can be measured. This can be repeated for various amplitude values over the range, V_t, of interest. An oscilloscope may be used to determine the square wave amplitude, or if this absolute accuracy is insufficient, a variable voltage DC standard can be used with an oscilloscope on a comparison basis.

1. High Speed Inputs

For high speed inputs an additional error in the point at which triggering occurs becomes a factor. It takes a finite charge for a trigger circuit to fire. Charge starts to accumulate when the input voltage crosses the trigger level set at V_t and when sufficient charge has accumulated, the trigger fires. In the meantime, however, the input signal is now at V_t^1 which is effectively where triggering occurs. This is summarized by Equation (8).

$$V_t^1 = V_t + \Delta V \tag{8}$$

where V_t = trigger point voltage

ΔV = error due to finite charge needed to cause the trigger to fire and,

V_t^1 = actual voltage level at which triggering occurs

The charge necessary to cause triggering is a function of the amplifier-trigger design and is related to both the sensitivity and bandwidth of the circuit. Each amplifier-trigger type may, however, be characterized by a constant K that is a measure of the amount of charge necessary and

$$K = vt \text{ mV} \cdot \text{nsec} \quad (9)$$

where K is a constant for an amplifier type.

The relationships described by Equations (8) and (9) are summarized by Figure 21 below.

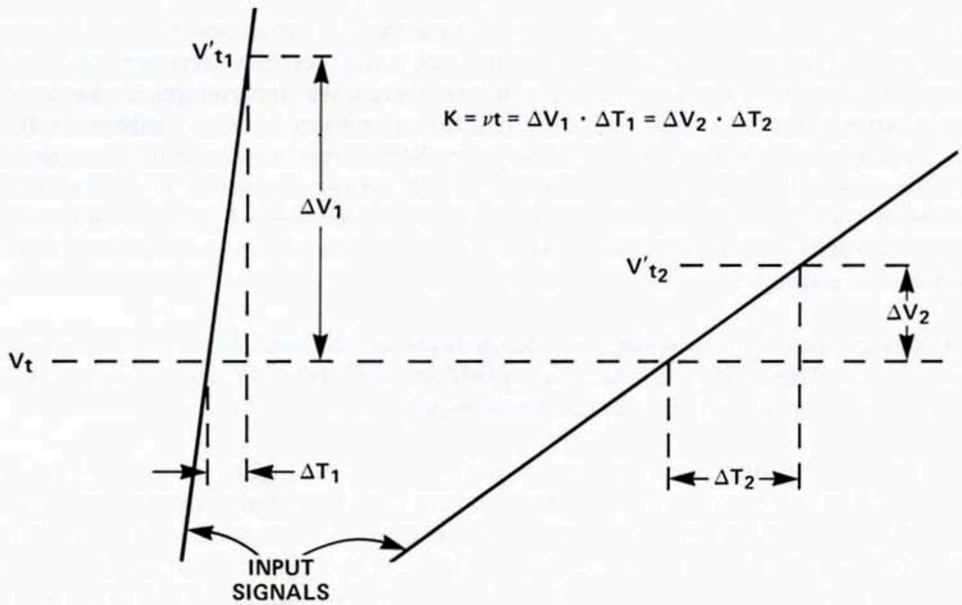


Figure 21

Trigger error caused by charging requirements of input amplifier-trigger. The magnitude of the trigger error, ΔV, at left is larger than for the slower input at right, however, the same charge was necessary in both cases.

A typically respectable value for K is 100 mV nsec (although state-of-the-art design can provide a figure almost ten times better). Given this value for K and a measurement of a 1 nsec risetime 2V pulse, the trigger error ΔV is given by solving the following equations for v.

$$\text{input signal slew rate } \frac{v}{t} = 2000 \text{ mV/nsec} \quad (10)$$

$$\text{amp-trigger constant } K = vt = 100 \text{ mV nsec} \quad (11)$$

Eliminating one of the variables in turn gives trigger error $\Delta V = v =$ 447 mV! and, in terms of time error $\Delta T = t = .22$ nsec. This error, ΔT, directly translates into a measurement error.

The examples given above would rate as worst case since the input is extremely fast, 2000 mV/nsec; nevertheless, it does serve to illustrate how large this error can become with fast inputs, and if errors of this magnitude cannot be tolerated, they can be reduced to negligible proportions by taking into account the considerations above.

2. Measuring Trigger Level

In days gone by when resolution of less than 1 μsec were all that was required, trigger level determination was satisfactorily accomplished by the oscilloscope intensification scheme. Signals derived from the start and stop channels were routed through the time interval meter to the Z axis modulation of an oscilloscope. With the input signal displayed on the oscilloscope, the points at which triggering occurred were evidenced by intensified dots. With today's resolutions of 10 nsec or better, the inherent delays of this method cause it to be inadequate for high speed signals.

A second popular technique is to provide calibrated front panel trigger controls. The drift and balance problems referred to earlier make this technique usable for the slower inputs only.

A third technique, that is somewhat better than either of the above, is to monitor the counter's main gate output and the signal being measured simultaneously on an oscilloscope. Changing the trigger levels varies the position of the leading and trailing edges of the gate output with respect to the signal being measured, and it is these edges of the gate that indicate where the START and STOP channel trigger levels are set. Obviously, this method is only usable on repetitive signals. There are also inherent delays in this system both within the instrument and the external cabling carrying all the signals that make it difficult to use for signal speeds of better than 10 nsec.

Since it is independent of signal speeds, the best method to measure trigger level is to actually measure the DC voltage V_R at which the trigger is set, as described above. In counters that use this technique the DC voltages are available at the instrument panel and can be measured with a DVM. In fact two counters, the Hewlett-Packard 5326B an 5327B, go one step further and include an internal DVM that can be used for DC voltage measurement in addition to measuring the trigger level voltages.

D. *measurement accuracy*

The accuracy statement for time interval usually reads as

±1 count ± trigger error ± time base stability

The same comments apply for time base stability in time interval and frequency measurements. In addition, however, it should be noted that for the measurement of predominantly short time intervals a high stability time base is not needed. For example, measuring 1 μsec to a resolution of 1 nsec requires a time base with accuracy of better than 1×10^{-3}, well within the accuracy of a TCXO. Note too that the short term stability needs to be better than the accuracy limiting ±1 count.

The inherent ±1 count error refers to one count of the clock frequency. Thus the higher the clock frequency the smaller this error. As stated earlier, 10 nsec (100 MHz), represents state-of-the-art at this time. Higher frequencies, such as 200 MHz, could be used but the instrument would then require digital processing capability to display the time interval measured.

Trigger error is rarely a factor since the inputs are usually pulses that have a faster slew rate across the trigger and hence produce less trigger error. Figure 22 summarizes this — given that the sinusoidal signal can be expressed as

$$e = e_m \sin \omega t$$

then the maximum slew rate is given by $\frac{(de)}{df} \max = \omega e_m$

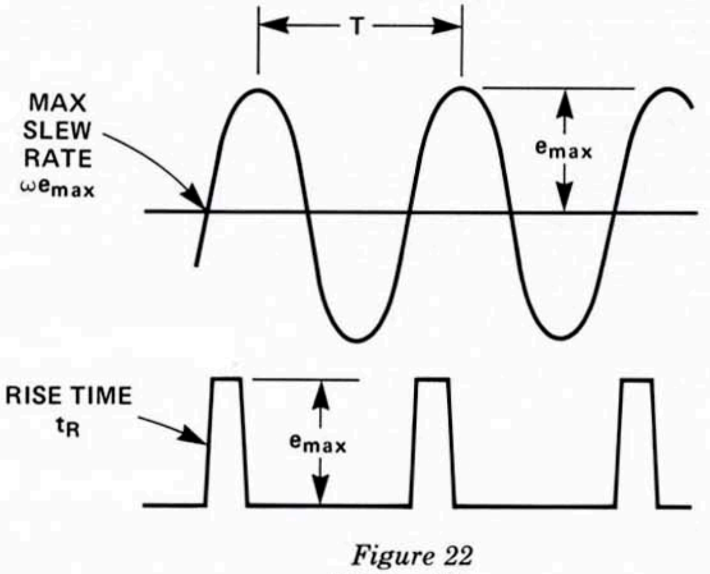


Figure 22
Trigger Error

This error is substantially reduced for a pulse type input over a sinusoidal input of same peak amplitude and having the same noise level. The reduction can be approximated to $2\pi t_R/T$.

Compare this to the slew rate of a pulse of risetime t_R and the same peak amplitude e_{max}. The slew rate is given by

$$\frac{de}{dt} \approx \frac{e_{max}}{t_R}$$

Thus the reduction in trigger error for a pulse input, assuming both signals have the same noise, is given by

$$\text{trigger error reduction} = \frac{2\pi t_R}{T} \tag{12}$$

Not included in the usual accuracy statement, but nevertheless extremely important, is trigger level settability. The importance of this is that errors in poorly set trigger levels can swamp any and all of the factors described above.

E. increasing the accuracy and resolution

There are several techniques that may be used to increase measurement accuracy and resolution over the 10 nsec limit of the conventional instrument. It should be noted that no matter what technique is used there is no easy way out in design and if a technique claims 1 nsec accuracy, for example, 1 nsec speed circuits must be employed. What the techniques do offer, however, is limiting the number of 1 nsec circuits needed over those required utilizing the conventional measurement method. Two of the techniques that can be used to increase accuracy and resolution are described below.

1. Time Interval Averaging

This technique is based on the fact that if the ±1 count error is truly random it can be reduced by averaging a number of measurements. The words "truly random" are significant. For time interval averaging to work the time interval must (1) be repetitive and (2) have a repetition frequency which is asynchronous to the instrument's clock.

Under these conditions the resolution of the measurement is:

$$\frac{\pm 1 \text{ count}}{\sqrt{N}}$$

where N = number of time intervals averaged.

With averaging, resolution of a time interval measurement is limited only by the noise inherent in the instrument. A typical figure of 50 picoseconds resolution can be obtained with good low noise design. This is not the whole story, however, since the averaging described to date suffers one severe limitation; namely, the minimum measurable time interval remains at the period of the clock. This limitation may be removed by circuits known as synchronizers.

The synchronizers operate as in Figure 23. The top waveshape shows a repetitive time interval which is asynchronous to the square wave clock. When these signals are applied to the main gate, an output similar to the third waveform results. Note that much of this output results in transitions of shorter duration than the clock pulses. DCA's designed to count at the clock frequency dislike accepting pulses of shorter duration than the clock. The counts accumulated in the DCA's will therefore approximate those shown in the fourth trace—the exact number of counts is indeterminant since the number of short duration pulses actually counted by the DCA's cannot be known. Since the time interval to be measured is slightly greater than the clock period, the fourth waveshape shows that the average answer will be in error, having been biased, usually low, because of the DCA's requirement of having a full clock pulse to be counted.

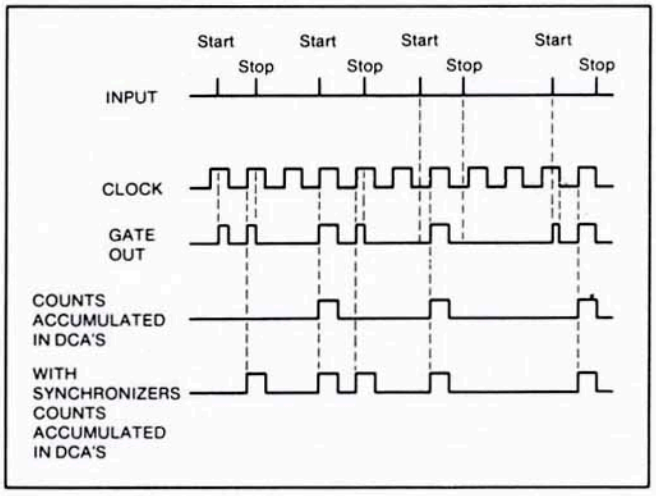


Figure 23

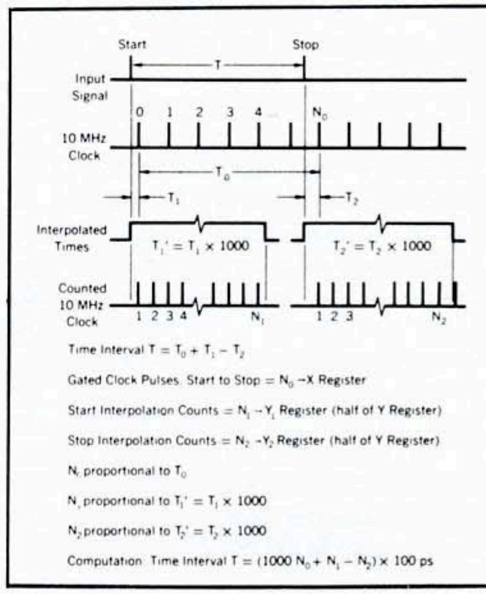
Synchronizer operation with time interval averaging

This problem is alleviated by the synchronizers which are designed to detect leading edges of the clock pulses that occur while the gate is open. The waveshape applied to the DCA's, when synchronizers are used, is shown by the fifth waveform. The leading edges are detected and reconstructed, such that the pulses applied to the DCA's are of the same duration as the clock.

Synchronizers are a necessary part of time interval averaging; without them the averaged answer is biased. In addition, it may easily be seen that with synchronizers involved, time intervals of much less than the period of the clock can be measured. This technique is only as good as the synchronizers, however, high speed synchronizers can enable intervals as small as 150 picoseconds to be measured, even though the clock frequency could be, for example, 100 nsec.

Figure 24

Time intervals measured by interpolation



2. Interpolation

In interpolation the inherent ±1 count ambiguity is measured and thereby removed. See Figure 24. The time interval T can be written as

$$T = T_0 + T_1 - T_2$$

where T_0 is the time indicated by counting the basic clock frequency and T_1, T_2 are the inherent time ambiguities between the clock and the start and stop pulses respectively.

The start interpolator charges a capacitor for the time T_1 and then discharges it for a duration 1000 times longer. During the discharge time the clock is again counted resulting in N_1 counts. The stop interpolator performs in exactly the same manner, resulting in N_2 counts. Coincidentally, the T_0 is measured in the conventional manner resulting in N_0 counts. It is easily seen that the time T is represented by the simple formula

$$T = 1000\ N_0 + N_1 - N_2$$

The resolution of the measurement has been increased 1000 times by interpolation. The system behaves exactly as if the counted clock were 1000 times faster. There is no limitation on the input, and events which occur only once can readily be measured. Interpolation does require arithmetic capability in the instrument; however, this can be put to good use in many ways. One is that it allows zero time interval (coincidence) to be measured and even negative time interval. Thus, not only magnitude but sign or which event occurred first can be determined.

The HP 5360A Computing Counter System utilizes exactly this scheme. The counted clock is 10 MHz but the instrument behaves exactly as if it were 10 GHz, providing 100 picosecond resolution.

VI. *microwave frequency measurements*

The counting techniques discussed in earlier sections of this book are limited in practice to an upper frequency limit of about 500 MHz at this time. Two techniques are available that enable microwave frequencies to be measured; heterodyne conversion and the transfer oscillator. Both of these processes essentially translate the microwave frequency to be measured down to within the frequency range that can be handled by the conventional frequency counter.

A. *heterodyne conversion*

Heterodyne conversion is the most accurate method of measuring high frequency or microwave signals. In a given measurement time it provides the same absolute resolution as the conventional direct counting frequency counter.

Heterodyne converters simply down convert the unknown frequency, f_x, by mixing with an accurately known frequency, f_a, such that the difference, f_d, is within the counter's range. See Figure 25. The frequency f_a is selected by first multiplying the time base to a convenient frequency, f_1, (usually the maximum direct frequency of the counter), and then passing this signal through a harmonic generator. The appropriate harmonic $Nf_1 = f_a$ (N is an integer) is selected by the tuning cavity and passed to the mixer. The cavity is operated from a front panel control calibrated to read the frequency f_a directly. The difference frequency $(f_x - f_a) = f_d$ is amplified and measured by the counter. To the counter reading the operator adds the front panel control setting f_a to obtain the final answer f_x. The tuning meter of Figure 25 indicates when the unknown frequency has been located.

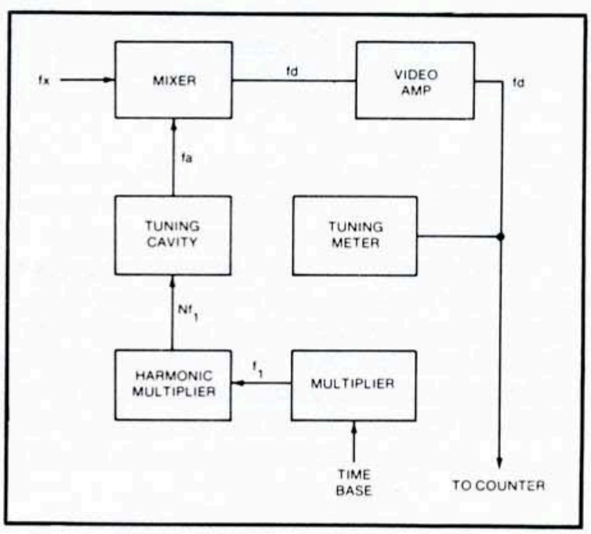


Figure 25

Basic operation of a heterodyne converter

This conversion process does not introduce any measurement error, provided the harmonic multipliers do not introduce noise in the multiplied signal Nf_1 that could be detected by the counter. Since noise increases as the square of the harmonic number N, a low noise oscillator and "quiet" multipliers are needed if full accuracy is to be maintained.

The heterodyne converter also has the advantage that it can handle wideband FM signals. So long as the signal, f_x, does not deviate to an extent that the down converted frequency, f_d, falls out of the IF band, the FM signal can be measured. Clearly too, this measurement is independent of deviation rate.

Pulsed RF signals can also be measured provided, of course, that the counter itself has this basic measurement capability. To fully handle pulsed RF, the input to the mixer (f_x input) should be balanced. A single ended input will produce a DC transient, or level shift, on the signal at frequency f_d that will prevent measurement on the first few μsecs of the pulse.

The process described by Figure 25 is manual in that the mechanical tuning cavity must be manually tuned to the correct harmonic number N. The operator reads the frequency f_d indicated on the counter and adds to it the tuning setting, Nf_1. While this is a relatively simple task, component technology has arrived to the point where the whole process can be automated. The YIG tuned filter is one technique and the switchable filter another for providing automatic tuning. The YIG filter is tuned by varying the current supplied to it and readily lends itself to the automation process. The YIG is usable through 18 GHz but it suffers from a limited frequency range (1 decade maximum), it requires considerable power and the sensitivity of counters using YIG's is much more temperature dependent. The major drawback, however, is speed—it takes approximately one second to sweep a YIG over its full frequency range, the "acquisition time" of counters using YIG's is therefore relatively slow.

With the advances in thin film technology, the switchable filter has become a practical possibility. The switchable filter has none of the drawbacks of the YIG, moreover, it can be extremely fast in acquiring the signal (less than 1 msec typical). The disadvantage is that the number of filter sections is proportional to frequency range; thus with a 500 MHz IF it would require eight filter sections to cover 4 GHz, easily possible on a practical basis, however, 18 GHz coverage would require 36 filter sections and would be prohibitively expensive.

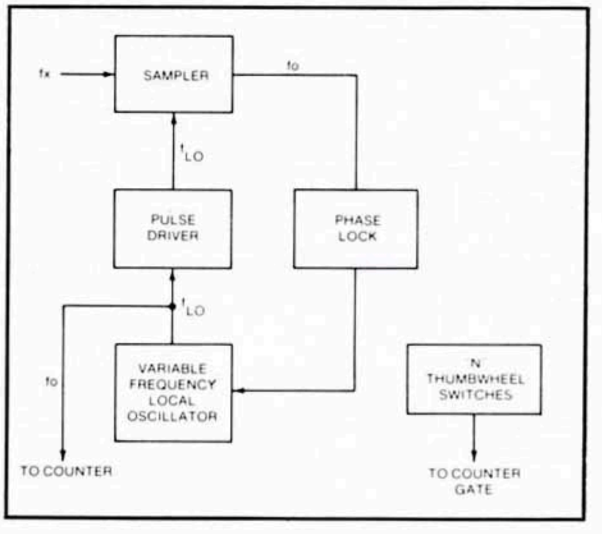
B. *transfer oscillator*

In contrast to the heterodyne converter, the transfer oscillator measures the frequency f_x by comparing it to the harmonic of an internal local oscillator of known frequency. The frequency information is "transferred" to the local oscillator.

The block diagram of the transfer oscillator is shown in Figure 26. To operate, the user simply tunes the local oscillator for zero beat (i.e., $f_x = Nf_{L0}$) and closes the phase lock loop. The local oscillator frequency is then measured by the counter and a direct reading of f_x can be obtained by extending the counter gate by N.

Figure 26

Basic Operation of a Transfer Oscillator



If f_x is completely unknown, so too is N, however, the latter can easily be determined by tuning the local oscillator to the next zero beat then

$$Nf_{L01} = (N - 1)\, f_{L02}$$

where

$$f_{L02} > f_{L01} \tag{13}$$

thus

$$N = \frac{f_{L02}}{f_{L02} - f_{L01}}$$

By opening the phase lock loop the transfer oscillator can also measure fm deviation and pulsed RF. The technique again is to zero beat a harmonic of the local oscillator with the input signal. Zero beat is detected on an oscilloscope so the minimum duty cycle pulsed signal that can be measured is determined by the CRT persistence. It is beyond the scope of this book to treat in any detail, the many measurements the versatile transfer oscillator can perform. The reader is referred to Hewlett-Packard Application Note 141, "AM, FM Measurements with the Transfer Oscillator" for more details.

The key to transfer oscillator performance is the harmonic mixer. A sampler is shown in Figure 26 and, thanks again to thin film hybrid technology, extremely broadband operation can be obtained with this device. Frequency measurements from essentially DC to greater than 20 GHz are now possible.

The manual process described above can also be automated as shown in the block diagram in Figure 27. The upper sampler and phase lock loop, phase locks the signal f_1 such that

$$f_x = Nf_1 - f_{IF} \tag{14}$$

The quadrature detector ensures that lock occurs at $Nf_1 - f_{IF}$ not $Nf_1 + f_{IF}$.

The lower sampler determines N. By offsetting f_1 a known amount f_0 gives

$$f_2 = f_1 \pm f_0 \tag{15}$$

and it is this signal that is used to drive the lower sampler whose output frequency, f_{IF2}, is given by

$$f_{IF2} = Nf_2 - f_x \tag{16}$$

thus

$$f_{IF2} = f_{IF} \pm Nf_0$$

The output of the second sampler at f_{IF2} is mixed with f_{IF} to obtain Nf_0. N is then determined in a ratio counter with Nf_0 and f_0 as inputs. Once determined, N is then used to extend the time base while f_1 is being measured. By offsetting the display by f_{IF}, Equation (14) is solved and the unknown frequency f_x displayed.

With good sampler design and a narrow bandwidth phase lock loop, extremely high sensitivity can be achieved. The front end looks like a narrow band pass filter, in contrast to the heterodyne converter that has a bandwidth equal to the IF, and it is this difference that, potentially at least, offers much higher sensitivity with the transfer oscillator technique.

The Hewlett-Packard 5340A Microwave Counter is an excellent example of an automatic microwave counter utilizing the transfer oscillator technique. Any signal from 10 Hz to better than 18 GHz can be measured automatically via a single input. Sensitivity is specified at -35 dBm, some 25 dB better than similar instruments utilizing different techniques.

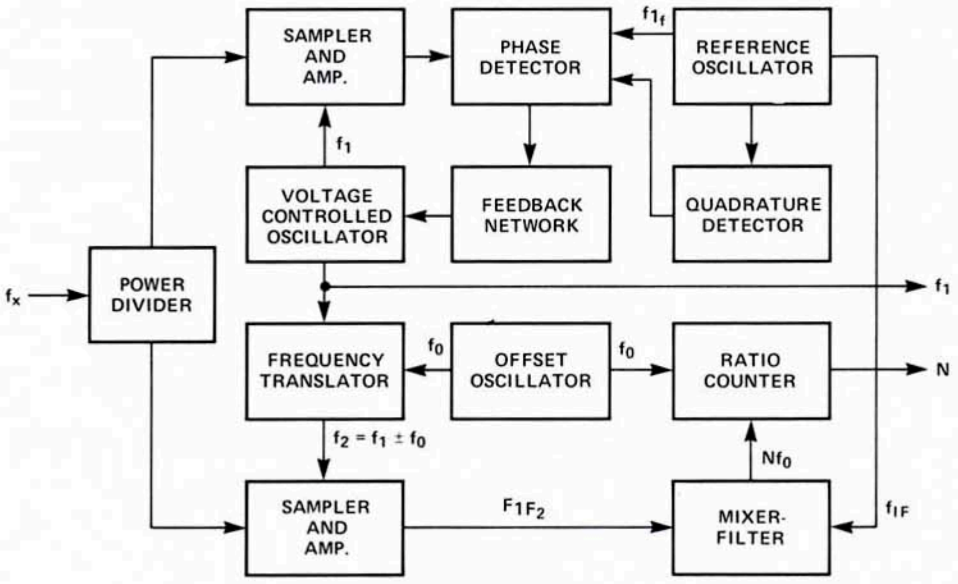


Figure 27
Automatic Transfer Oscillator Block Diagram

The automatic transfer oscillator relies on its operation on phase locking a harmonic of a known signal to the unknown. Inherently, therefore, such a technique cannot be used to measure pulsed RF. Conversely, the automatic heterodyne converter can measure pulsed RF automatically, since the signal can be detected in the IF when the tuning element is at the appropriate harmonic of the time base reference.

The individual advantages of the two methods are summarized in Figure 28. It should be recognized, of course, that these advantages are potential only and the degree to which these advantages are realized is dependent on the quality of design.

Heterodyne Converter	**Transfer Oscillator**
High resolution, accuracy	Wide bandwidth
Good FM tolerance	High sensitivity
Can measure pulsed RF[1]	Can measure pulsed RF, FM deviation, AM modulation[2]

[1]Assumes counter measuring the IF has pulsed RF capability.
[2]This capability applies to manual mode only with phase lock loop open.

Figure 28
The relative merits of heterodyne converters and transfer oscillators.

VII. *some examples of component technology*

The rapid evolution in component technology has had as profound an effect on the electronic counter as it has on digital computers and communications. Thanks to this evolution, the measurement power and flexibility of the counter has increased enormously. Moreover, for a given performance, the price has declined substantially. The purpose of this chapter is to briefly acquaint the reader with samples of today's state-of-the-art technology, technology that has made the counter of today possible.

The **integrated circuit** has had the greatest effect and has made possible counters that are smaller, lighter, more reliable and less expensive. The **MOS** process provides extremely high component density which, in turn, enables complex functions to be performed by the one IC chip. Figure 29 is an example of such high density (LSI) MOS chip.

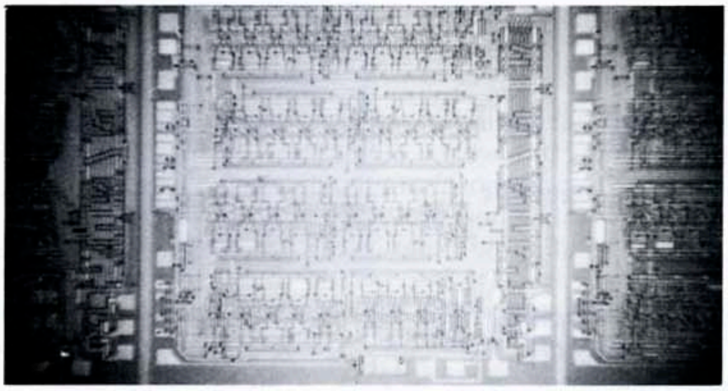


Figure 29

A 10 MHz MOS/LSI six digit counter circuit. The chip is 0.096 in. by 0.118 in. and contains 930 transistors.

MOS is limited in speed to about 10 MHz and finds major use in counting circuits to that speed and also for digital processing and control. In fact, much of the control that was once performed by mechanical switches and wafer contacts can now be performed electronically by MOS memories. This, in turn, has enabled remote programming of the instrument to be implemented in a more flexible and complete manner than was previously possible.

Bipolar technology is capable of much higher speeds than MOS. Figure 30 is an example of a state-of-the-art bipolar IC.

Figure 31 shows a linear bipolar circuit that is state-of-the-art in every way. The bandwidth of the amplifier extends from DC to beyond 600 MHz. The initial design and optimization of this amplifier was performed by a computer model.

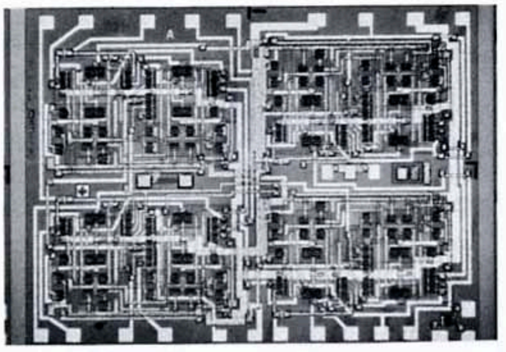
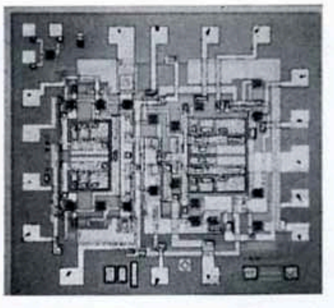


Figure 30

A 250 MHz bipolar IC chip. This circuit is a quinary, accepting a pulse train of frequency up to 250 MHz and dividing by a factor of five.

Figure 31



A linear bipolar amplifier chip. Frequency response is DC to >600 MHz, voltage gain is 8, noise figure 26 dB and the input impedance is 10 kΩ with shunt capacity of less than 1 pF. Thus this amplifier can be used in either 50Ω or high impedance environments.

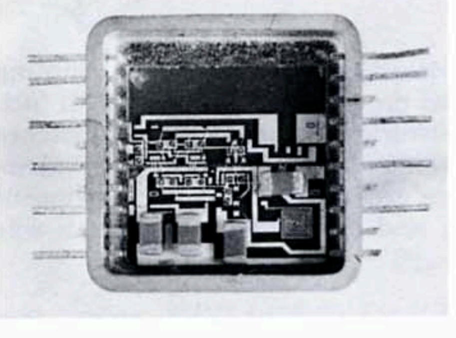


Figure 32

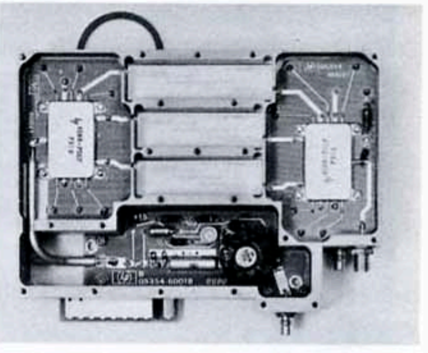
1 GHz thin film binary. Actual size is approximately one half inch square.

As noted earlier, the monolithic processes described are limited at this time to around 500 MHz. For frequencies above this limit the monolithic process gives way to **hybrid** technology. The process of placing discrete devices on a thin film substrate is known as the hybrid process. The speeds at which hybrids operate depends on the devices themselves, but frequency response through 20 GHz can be achieved. The substrates can be either alumina, sapphire or quartz. The cost of the substrates increases in the order stated and so too does the frequency response. In general, alumina substrate can be used up to 5 GHz, sapphire to around 13 GHz and quartz for higher frequencies.

The thin film chip shown in Figure 32 utilizes sapphire as the substrate and consists of a master-slave flip-flop that acts as a binary. Bandwidth is in excess of 1 GHz and sinusoidal or pulse inputs can be accepted.

The 4 GHz switchable filter shown in Figure 33 measures approximately 3 inches by 5 inches. At top left and top right are two single pole, six throw PIN switches mounted on sapphire substrates. Between these switches are six bandpass filters (only three are shown here, the other three are on the under side). These filters are half wave resonator coupled, mounted on alumina substrate. The filters are 500 MHz wide and the input to the filter assembly is a 500 MHz comb generated by applying a 500 MHz signal to the power amplifier—spectrum generator assembly located at the bottom in Figure 33. Switching speed is extremely fast; any filter section can be selected in less than 20 μsec.

Figure 33



4 GHz switchable filter. See text above for description.

A DC to 20 GHz sampler (Figure 34) utilizes beam lead chip sampling diodes on a thin film quartz substrate. Used as the front end of an automatic transfer oscillator microwave counter (the Hewlett-Packard 5340A) these samplers allow extremely high sensitivity in the order of -35 dBm to be achieved.

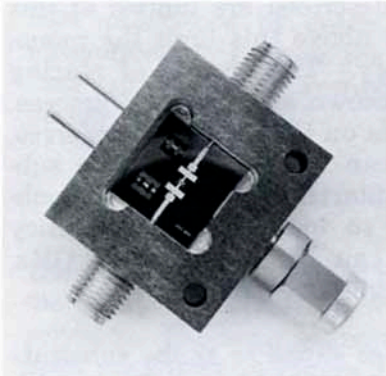


Figure 34

20 GHz sampler relative conversion efficiency varies by less than 1 dB over the 20 GHz spectrum.

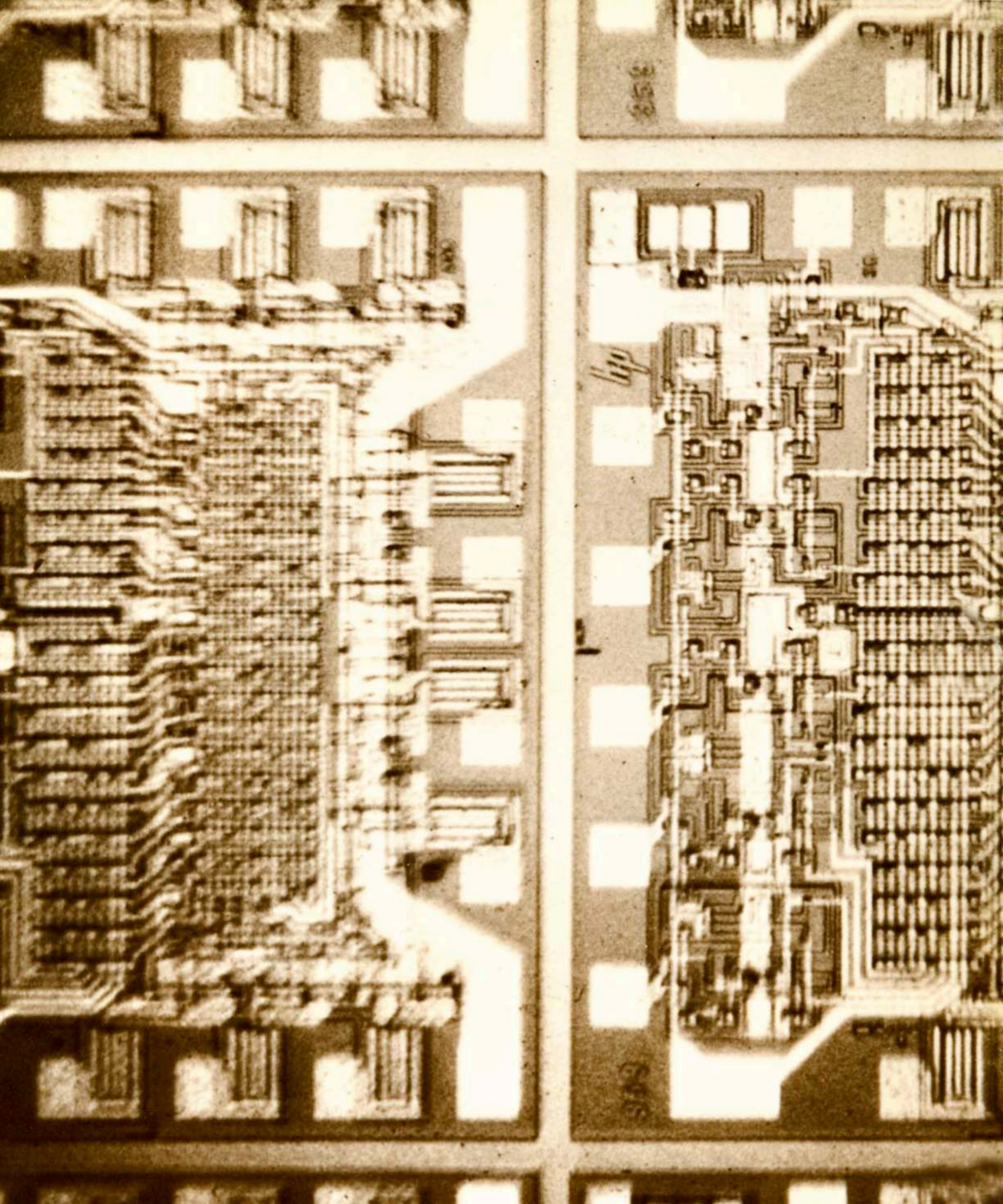
Often overlooked because of the rapid evolution in IC technology, is the crystal oscillator manufacturing process. From the raw crystal and about fourteen cutting, grinding and polishing processes later comes the finished product.

Figure 35

Approximately 14 separate processes are needed to convert the raw quartz crystal to the finished product.

The crystal shown here is the heart of a state-of-the-art 5 MHz oscillator which when enclosed in a proportional oven exhibits aging rates of $<5 \times 10^{-10}$ per day and temperature stability of $<3 \times 10^{-9}$ over the range 0°C to 50°C.

Further advances in component technology, particularly in the area of monolithic integrated circuits, can be expected. The result will be a continuation of the trends already experienced since IC's were first used in counters; that is, further price reductions for the same performance or improved performance for the same cost—in short, a steadily improving price/performance ratio.



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