# Errata

**Document Title:** Direct-Type Frequency Synthesizers: Theory, Performance and Use (AN 96)

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## HP References in this Application Note

This application note may contain references to HP or Hewlett-Packard. Please note that Hewlett-Packard's former test and measurement, semiconductor products and chemical analysis businesses are now part of Agilent Technologies. We have made no changes to this application note copy. The HP XXXX referred to in this document is now the Agilent XXXX. For example, model number HP8648A is now model number Agilent 8648A.

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# HP DIRECT-TYPE FREQUENCY SYNTHESIZERS
## Theory, Performance and Use



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# SECTION I

# INTRODUCTION

## Definition of a Synthesizer

A frequency synthesizer is an instrument which translates the frequency stability of a single frequency to any one of many other possible frequencies usually over a broad spectrum. Its single frequency generator is usually of frequency standard quality; thus, the synthesizer may be called a frequency standard capable of furnishing a large number of frequencies. The two basic approaches to frequency synthesis are known as Direct and Indirect. Direct synthesis performs a series of arithmetic operations (multiplication, division, and mixing) on the signal from the frequency standard to achieve a desired output frequency. The indirect method uses tunable oscillators, phase-locked to harmonics of a standard frequency, to derive a desired output frequency.

One advantage of HP direct synthesis type of instruments is that they can be switched (frequency changed) much faster than indirect synthesis where the switching speed is governed by the phase-locked circuitry. With direct synthesis the limitations on switching speeds are set by the time constants in the filtering circuits on the control lines to the switches and by the circuit bandwidths. Direct synthesis is also fail-safe in that if any part of the circuitry becomes inoperative the output signal is not present.

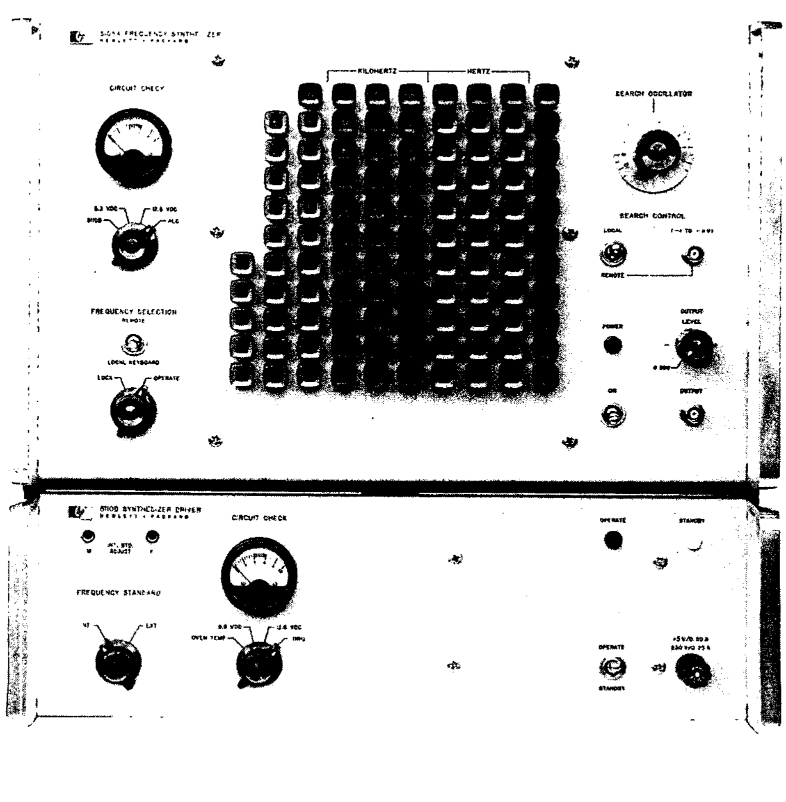
On the other hand, the indirect method of synthesis can be less costly for an equal frequency range as less circuitry is normally required in the synthesis process. It then follows that for an equivalent cost the indirect synthesizer can include modulation and attenuation functions that the direct synthesizer does not.

Hewlett-Packard direct synthesizers are particularly noted for the following:

1. Complete guaranteed specifications on tested noise and stability characteristics for all instruments -- spurious signals, phase noise, a.m. noise, rms fractional frequency deviation, long term stability, and harmonic signals.
2. Frequency switching speeds of 4 $\mu$s for 2 digits and typically 20 $\mu$s for a change of many digits.
3. Rugged and modular construction designed for reliable, long life operation.
4. Customized modifications for special requirements often to achieve a cost savings.

This application note describes these characteristics in detail and how they affect application of precision frequency synthesizers. Also, the basic operation of HP direct frequency synthesizers is covered as well as a brief description of typical applications.

Figure 1-1. 5105B/5110B Frequency Synthesizer, 100 kHz – 500 MHz



# SECTION II

# PRINCIPLES OF OPERATION

## 5110B Synthesizer Driver

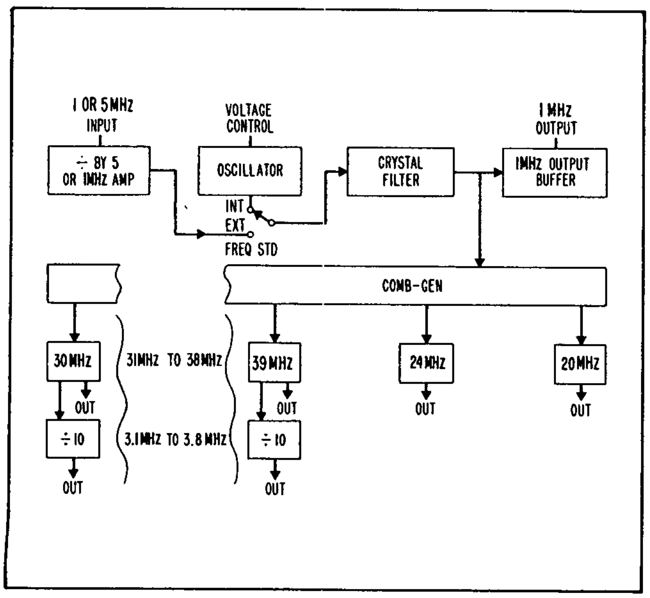
The HP 5100B and 5105A Synthesizers have top frequencies of 50 MHz and 500 MHz, respectively. They are comprised of switching matrices and mixers to process the input signals derived from the 5110B Synthesizer Driver. This arrangement is used not only to keep down the size and weight of the individual equipment cabinet but also to allow a possible cost savings. A single 5110B Driver can furnish signals for up to four synthesizers.

The direct method of synthesis used by the 5110B and 5105A requires 22 spectrally pure signals. These signals generated in the 5110B Synthesizer Driver are derived from either an internal 1 MHz frequency standard which has a maximum aging rate of ± 3 parts in $10^9$ per 24 hours or an external 1 MHz or 5 MHz standard.

An external 1 MHz signal is accepted without alteration, whereas if an external 5 MHz signal is applied, a divider circuit reduces the frequency to 1 MHz. The internal frequency standard assembly contains a precision 1 MHz quartz crystal resonator housed in a proportional oven. A useful feature of the internal frequency standard is the provision for voltage control which permits the frequency to be pulled 5 parts in $10^8$ with an externally applied voltage.

Figure 2-1 shows the technique for generating the 22 signals in the 5110B. The selected frequency standard is filtered by the crystal filter to eliminate broadband noise. This filtered 1 MHz signal is amplified

Figure 2-1. 5110B Block Diagram




and furnished to both an output jack for external use such as synchronizing other equipment and to a spectrum generator comprised of a step-recovery diode and a group of filters.

The pulse output of the step recovery diode circuit passes through a 24 to 39 MHz bandpass filter and an amplifier. It is then fed to active filters comprised of synchronously tuned transistor amplifier stages which select the 20 MHz, 24 MHz and 30 through 39 MHz components. (This filtering is very effective, with adjacent 1 MHz signals 105 dB down.) The 30 to 39 MHz components are also applied to decade frequency dividers yielding the low frequency spectrum output of 3.0 to 3.9 MHz in 100 kHz steps.

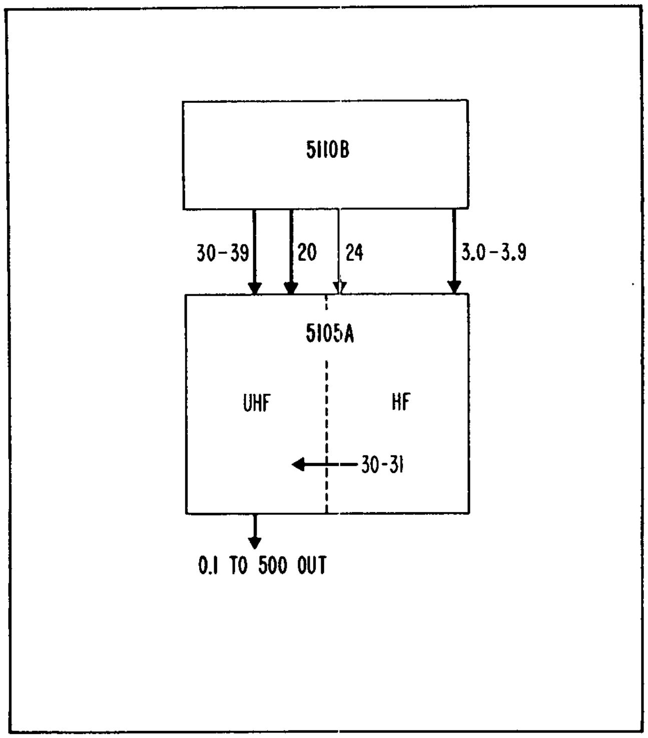
In summary, the 5110B Synthesizer Driver provides to either the 5100B or 5105A Frequency Synthesizer, by means of rear panel BNC connectors, very stable and clean signals of 20 MHz, 24 MHz, 3.0 through 3.9 MHz in 100 kHz steps, and 30 through 39 MHz in 1 MHz steps. (The 20 MHz signal is used in the 5105A only.) One 5110B can drive more than one or combinations of 5100B's and 5105A's up to four. Options for these instruments specify how many can be driven. If an optional unit is driving less than its full complement of synthesizers, the unused outputs should be terminated in 50 ohms. 10510A terminations are available for this purpose.

## HF Section

The 5100B or 5105A Frequency Synthesizer combines by direct synthesis the 22 signals provided by the 5110B Synthesizer Driver. The 5102A and 5103A utilize the same hf design but with an internal driver and without a uhf section. Figure 2-2 shows the 5100B and 5105A Synthesizers separated into an hf section and a uhf section. The block diagram of the hf section is shown in Figure 2-3. This section utilizes the 24 MHz and 3.0 through 3.9 MHz signals from the 5110B. The latter ten signals are fed into a Diode Switch Matrix controlled by dc voltages. Control is provided either by front panel pushbuttons or remotely through the rear panel connectors.

The basic diode switch circuit is shown in Figure 2-4. These switches are embedded in an aluminum casting to achieve maximum shielding of one section from another. Switching effectiveness is demonstrated by an on-to-off ratio and cross talk of better than 110 dB. The front panel pushbuttons in the $10^{-1}$ to $10^6$ columns switch frequencies between 3.0 and 3.9 MHz into the Mixer Divider Assemblies for each column. This capability points out one of the key advantages of the direct synthesis method -- all frequencies appearing at the inputs to the Diode Switch are always present. Limitations on frequency switching speed are set only by the time constants in the filtering circuits on the control lines to the switches. Circuit bandwidths fol-

Figure 2-2. 5105A/5110B Simplified Block Diagram




lowing the switch establish propagation delay times of the changing signal traveling through switched decades.

Note that there are eight separate outputs from the switch assembly, each of which may be individually selected between 3.0 and 3.9 MHz; and each output is connected to a decade module. The first seven decades (those controlled by the seven least significant columns) are composed of two balanced mixers and one decade divider. The eighth decade is similar to the first seven except that it lacks the divider. The

Figure 2-3. HF Section Block Diagram

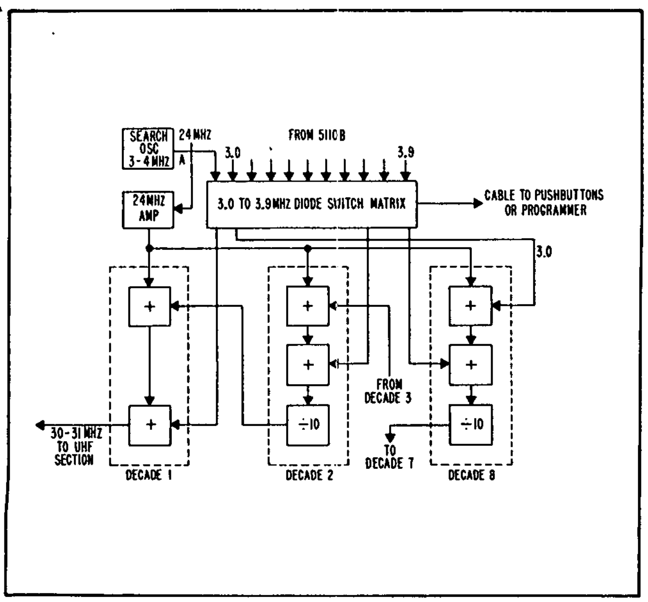
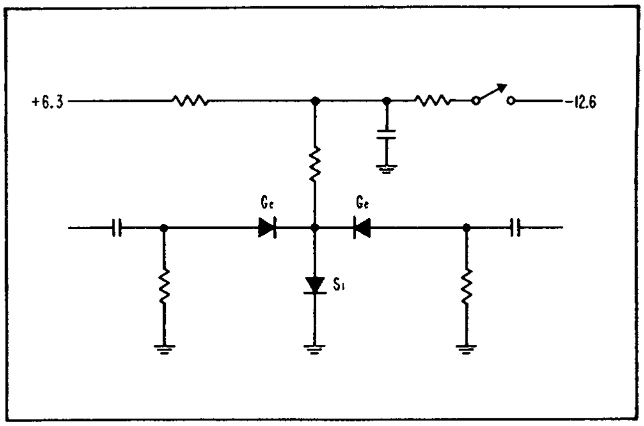



Figure 2-4. 3 to 4 MHz Switch




design is such that the optimum compromise between rejection of spurious signals and lowest signal-to-noise ratio is achieved.

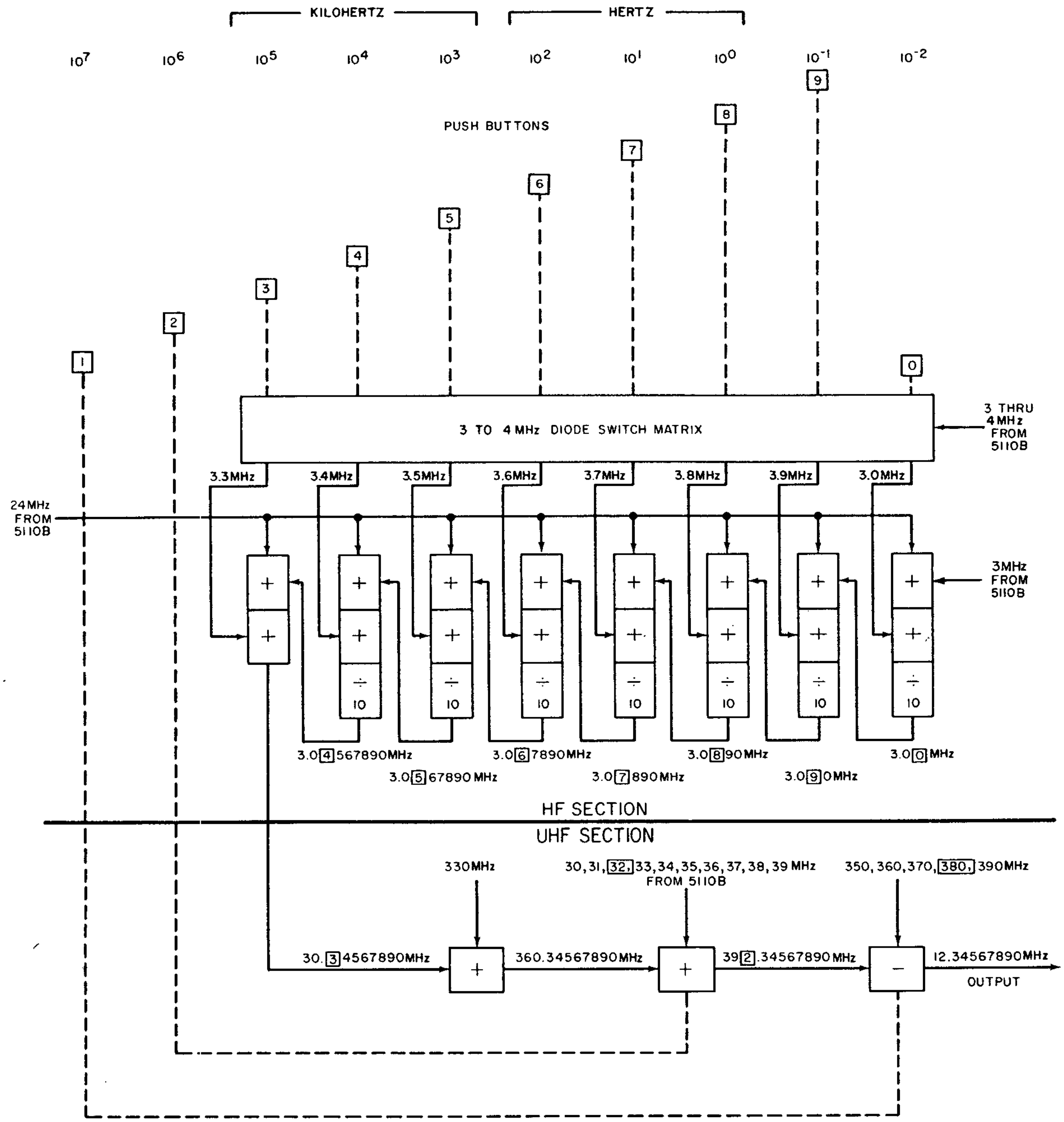
Figure 2-5 applies to the 5100B* and the following situation: The pushbuttons are shown for a selected frequency of 12, 345, 678. 90 Hz. The digit 0 is selected in the least significant column ($10^{-2}$ Hz). 24 MHz and 3. 0 MHz from the 5110B are added in the first mixer of the selected decade. The resultant 27 MHz signal is then added to a selected 3. 0 MHz signal (resulting from the pushbutton "0" in the $10^{-2}$ column being depressed) in the second mixer. Had we selected the digit "2" in the least significant column, 3. 2 MHz would have been added to 27 MHz in the second mixer. Returning to the original example, the addition of 27 MHz and 3. 0 MHz results in 30.0 MHz. This synthesized signal is now divided by ten to give 3.00 MHz. It should be noted that a great deal of filtering is carried on between successive operations in this and every other decade module to minimize the possibility of appreciable spurious signals. The resultant signal is now fed to the second decade where it is added to 24 MHz to obtain 27. 00 MHz.

If the pushbutton column controlling the second decade ($10^{-1}$) is now set to the digit "9", as shown, 3. 9 MHz from the switching matrix is added to 27.00 MHz in the second mixer to obtain 30. 9 MHz. This signal is, in turn, divided by 10 to 3. 090 MHz. The third decade first adds the synthesized 3.090 MHz from the output of the second decade to 24 MHz to obtain 27.090 MHz; 3. 8 MHz is added to obtain 30. 890 MHz. This value is divided by 10 to yield 3. 0890 MHz.

The basic operation of the hf section should now be clear - we successively add digits starting with the least significant and then divide by ten. After the least significant columns ($10^{-2}$, $10^{-1}$, $10^{0}$, $10^{1}$, $10^{3}$, $10^{4}$), the result is a synthesized signal of 3.04567890 MHz. In the eighth decade ($10^{5}$) this signal is again added to 24 MHz and the selected 3. 3 MHz signal. However, since there is no divider in this decade, it provides an output of 30. 34567890 MHz.

*The other direct synthesizers have the same principle of operation but with different frequency ranges.

Figure 2-5. 5100B Block Diagram



To generalize, each decade receives 24 MHz from the driver. The least significant decade always adds 3.0 MHz before its selected 3.0 to 3.9 MHz signal is added. All decades after the least significant one utilize the synthesized frequency of the previous decade in place of the fixed 3.0 MHz signal. The output of the hf section is always between 30 MHz and 31 MHz.

The eight least significant columns have, in addition to pushbuttons 0-9, an "S" pushbutton. If this position is selected in any of these columns, a Search Oscillator, with frequency variable between 3 and 4 MHz, is substituted for the 3.0 to 3.9 MHz signals in the selected column. See separate section on the Search Oscillator.

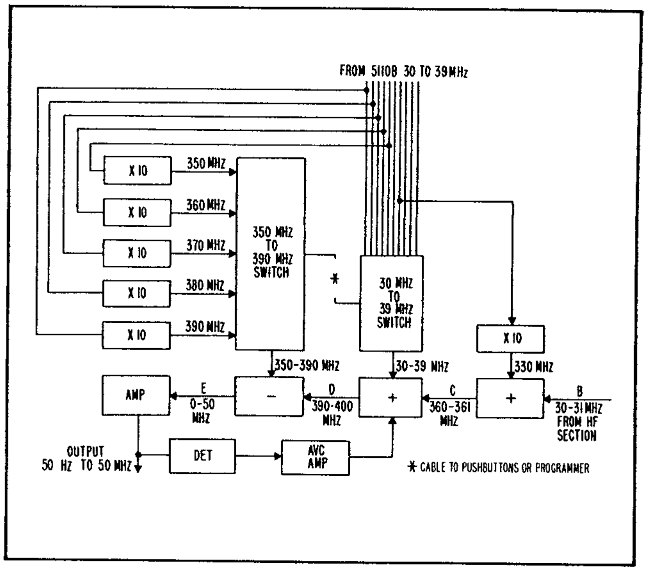
The hf section design is common to all four Hewlett-Packard direct synthesizers. A 30-31 MHz signal is precisely formed by frequency selection and mixing to represent eight digits punched into the pushbuttons. Following the hf section in the 5100B and 5105A is a uhf section to add more significant digits. But for the 5102A and 5103A low frequency synthesizers, the 30 to 31 MHz signal is directly processed to give the final output frequency. (This is explained in detail in a later section.)

## 5100B UHF Section

The uhf section translates the 30 to 31 MHz signal synthesized in the hf section to the desired frequency range of 0 to 50 MHz as the synthesizer output. Figure 2-6 shows how the signals used on our example of 12.34567890 MHz, shown in Figure 2-5, are generated. The 33 MHz signal from the synthesizer driver is multiplied by 10 to yield 330 MHz. This signal is mixed with the signal generated in the hf section to obtain a signal between 360 and 361 MHz. In our example we would add 30.34567890 MHz to 330 MHz yielding 360.34567890 MHz. This signal is then filtered and mixed with one of ten selected frequencies (30, 31, 32, 33, ..., 39 MHz) to obtain a frequency in the range of 390 to 400 MHz. The frequency in the 30 to 39 MHz range is chosen by a pushbutton (digit) in the $10^6$ column, using circuitry very similar to that used in the diode switching matrix of the hf section. In our example, 32 MHz was selected corresponding to the "2" digit, giving a 392.345,678,90 MHz signal out of the second uhf mixer.

A frequency has been generated, selectable in 1 MHz steps, with .01 Hz resolution. All that remains to synthesize our desired frequency is the most significant digit. This is achieved by mixing X10 multiplies of 35, 36, 37, 38, 39 MHz from the 5100B with our synthesized signal. (The exact value is determined by the depressed digit in the $10^7$ column.) The resulting difference frequency is variable through 50 MHz and is the final synthesized frequency. In our example the 380 MHz signal would be chosen. This, when subtracted from 392,345,678.90 Hz, gives the final output 12,345,678.90 Hz.

Figure 2-6. 5100B UHF Section




In summary, the uhf section of the 5100B produces an output frequency between 0 and 50 MHz from the synthesized 30 to 31 MHz signal output of the hf section. It does this by first translating the hf signal up to 390 to 400 MHz, then down to the output frequency. The advantages of this up-down translation are that undesired intermodulation products formed in the mixers can be removed by filtering and that those products or spurious signals formed that are within the desired band are of a high order and thus with proper design are at a low level (i.e, < 95 dB below the desired signal).

## 5102A and 5103A

The 5102A and 5103A Frequency Synthesizers are each contained in the same size package as the 5100B. It can be seen in Figures 2-7 and 2-8 that the 5102A and 5103A are very similar to each other and to Figures 2-1 and 2-5. The difference between the two synthesizers is in the frequencies supplied to the final mixer.

Functionally the 5102A and 5103A are very similar to the 5100B/5110B without the uhf section. Other differences are as follows:

1. The dual ranges of the 5102A and 5103A extend the instrument's capability at minimum cost and without sacrifice of convenient size. Both units have the frequency standard source and spectrum generator in the same package with the switching and synthesizing circuitry.
2. An output of $f_o + 30$ MHz has been provided on the rear panel of both models where $f_o$ is the number punched into the pushbuttons. This will be called $f_{out}$. The selected frequency $f_o$ is equal to ($f_{out}$ x 10) - 300 MHz in the 10 MHz range (5103A), $f_{out}$ -30 MHz in the 1 MHz range (5102A/5103A) or ($f_{out} \div 10$) -3.0 MHz in the 100 kHz range (5102A). The $f_o + 30$ MHz output is therefore always between 30 and 31 MHz with a S/N ratio that is better than that obtained by multiplying the output band to cover 30 to 31 MHz.

Figure 2-7. 5102A Block Diagram

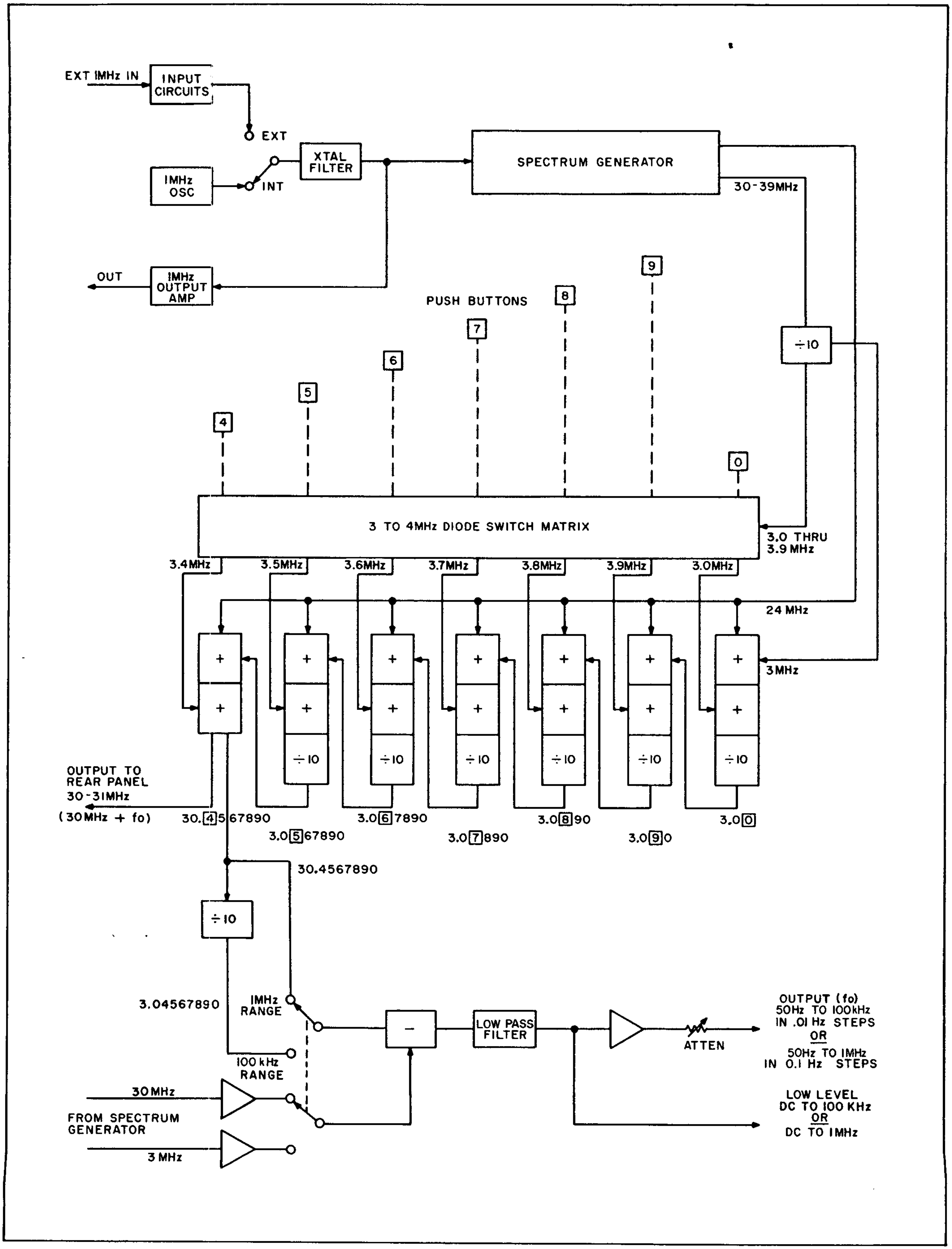
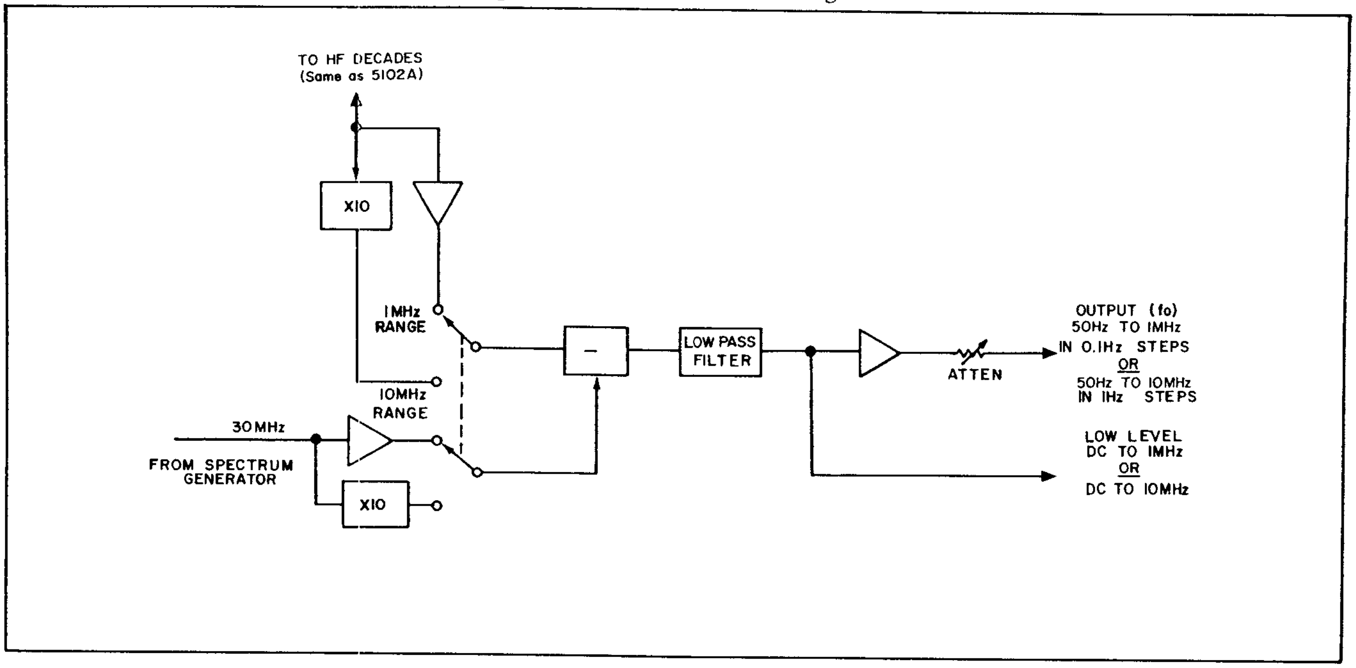


Figure 2-8. 5103A Block Diagram




3. The signal-to-phase noise and signal-to-a. m. noise ratios are superior to those of the 5100B/5110B on all ranges except for the 10 MHz range for which they are equivalent. On each range the harmonic rejection of the 5102A and 5103A is better than the 5100B. However, more elaborate filtering circuitry in the 5100B/5110B results in a superior spurious specification.

4. The RMS fractional frequency deviation at any given frequency for the 5102A and 5103A is superior to the 5100B/5110B. This is due primarily to the absence of the uhf section in the lower frequency instruments. As a result the noise levels of the fundamental frequencies are not multiplied as much, and hence do not degrade the signal as much as in the 5100B.

## 5105A

The 100 kHz to 500 MHz HP 5105A Synthesizer is like the 5100B as they both use the 5110B and have the same hf section. Figure 2-9 indicates the general block diagram.

The major difference in the 5105A as compared to the 5100B is in the uhf section. Uhf oscillators are used to generate 300 to 390 MHz instead of multiplication from a lower frequency. By having two branches for the selected uhf oscillator frequency which are difference mixed in the final mixer, variation in the uhf oscillator frequency is cancelled out to preserve the inherent stability of the 1 MHz quartz crystal of the 5110B Driver. This drift-cancelling method reduces the number of component modules required and results in the same size instrument with ten times the frequency capability at a very competitive price.

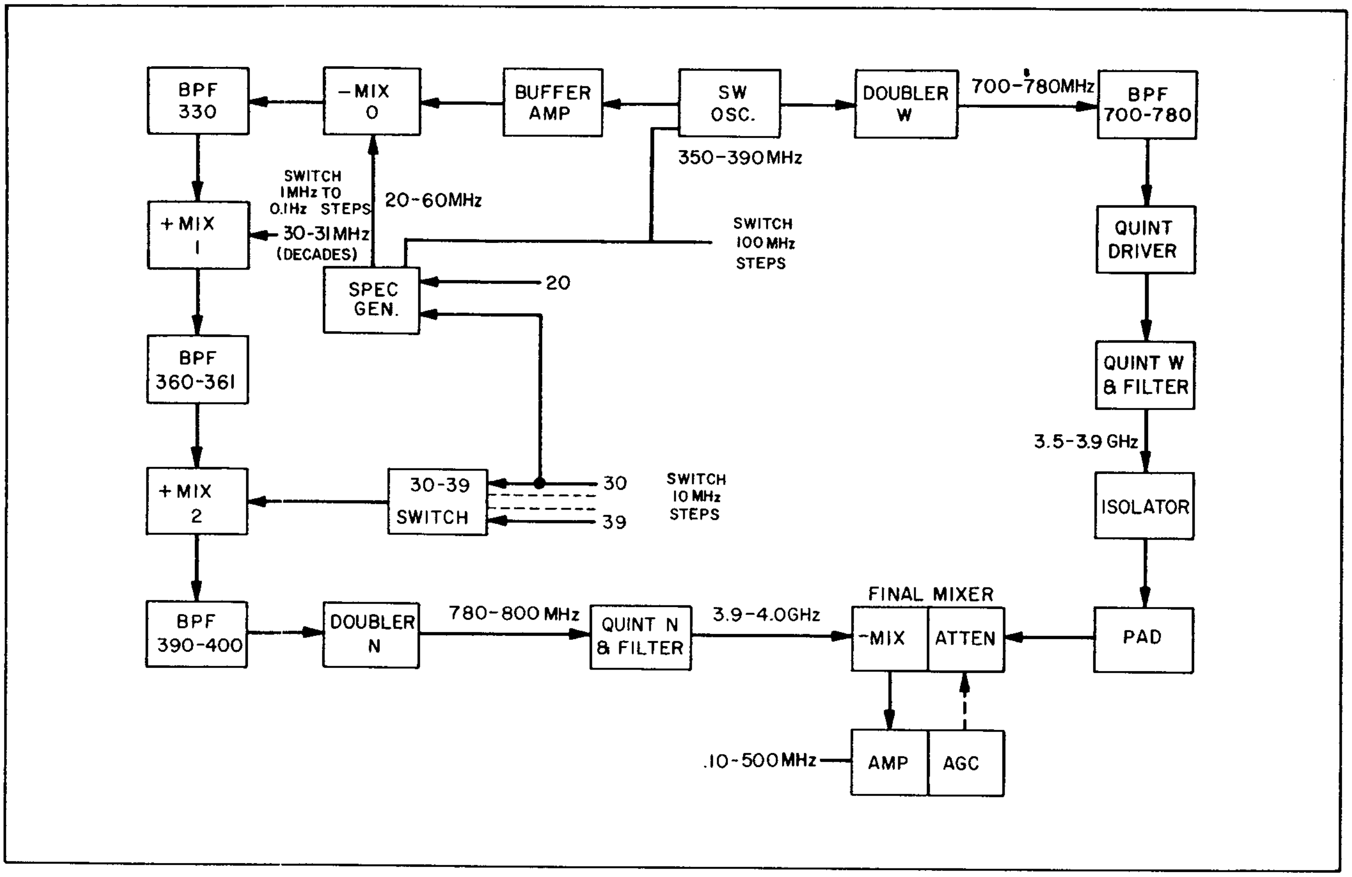
An important advantage of the "drift cancelling" technique used in the 5105A is an excellent signal-to-noise ratio of the output signal. Although its operation is much like a 5100B multiplied by 10, the 5105A yields less noise than if the frequency was obtained by multiplying the 5100B output by 10. In the conventional technique of multiplying a lower frequency to uhf in two offset channels and mixing to obtain the difference, the noise is also multiplied and modified. Thus, the noise cannot be completely eliminated when down converting.

The switched oscillator block contains five uhf oscillators - 350 to 390 MHz in 10 MHz steps. One, only, is on at any time to reduce the possibility of spurious signals. The "on" oscillator is selected by the MHz pushbuttons in the hundreds column of the front panel. These pushbuttons also control the output frequency of the spectrum generator. Frequencies from 20 to 60 MHz in 10 MHz steps are selected in conjunction with the uhf oscillators such that the output of the "0" mixer will always be 330 MHz.

The constant 330 MHz is added to the hf section output (30 to 31 MHz) in mixer 1 to determine the output frequency in the 0. 1 Hz to 1 MHz columns. The carrier frequency, now at 360 to 361 MHz, is added to 30 to 39 MHz in mixer 2. The 30 to 39 MHz switch (similar to 5100B) selects the output frequency in the 10's of MHz column. This signal (390 to 400 MHz) is then multiplied by ten and fed to the final mixer. The final mixer (down converter) gives the difference between the synthesized frequency (3. 9 to 4 GHz) and the switched oscillator frequency that is multiplied by ten (3. 5 to 3. 9 GHz).

Note that the hundreds of MHz output frequency is determined by the 3. 5 to 3. 9 GHz signal frequency selected by the leftmost pushbutton column. The difference frequency has a maximum of 500 MHz and a minimum of 0. 1 MHz. The minimum is determined by the output amplifier coupling capacitor (ac coup-

Figure 2-9. 5105A UHF Section




by the output amplifier coupling capacitor (ac coupling) and the ALC loop time constant.

## Phase Modulation

A phase modulation input is provided on the rear panel of the 5105A. This allows modulating any output frequency at a maximum deviation of 3 radians with a deviation rate up to 1 MHz. Typical sensitivity (small signal) is 3 radians/volt with 50 ohms input impedance. The modulating signal is applied to the 330 MHz signal in mixer 1.

## Remote Programming

Any frequency or search oscillator position that may be selected by front panel pushbuttons on the synthesizer may also be selected remotely. All HP synthesizers can be remotely programmed with frequency selection typically accomplished in less than 20 microseconds (less than 1 millisecond for 5100A's manufactured before January 1, 1966). Each pushbutton on the front panel of the synthesizer is represented by a unique pin in the fifty pin connectors located on the rear of the instrument. The format is: one of eleven on, 10 off (or disregarding the search pin, one of 10 on, 9 off) for each column of the synthesizer. Four modes of operation are easily identifiable, manual remote programming (hard wiring), slow electronic control (typically 1 ms), fast electronic control (typically 20 μs) and ultra fast special versions and applications. All that is required for manual remote programming is -12.6V dc supplied to the correct pins with the local-remote switch in the remote position. Each connector includes a -12.6V dc pin controlled by the local remote switch and a grounded pin.

One or more columns can be remotely programmed while the others are under local control. In this case, the local-remote switch is placed in local position. Each column to be programmed in this mode must have all its buttons out (not depressed) and now those columns may be programmed at the rear panel. Note that the switchable -12.6 V dc at the rear panel is not available in the local position so that the non-switched -12. V dc must be used. For switching speeds of typically 1 ms the electronic control must supply -12.6V dc to the "on" digits and must return the "off" digits to an impedance of 100 kΩ or greater. To obtain "20 μs typical" operation both on digit drive (-12.6V dc) and off digit drive (+1.5V dc) must be provided and both must originate in approximate synchronism from low source impedances. Switching Speed Limitations and a more detailed discussion of switching speed is given in Section IV.

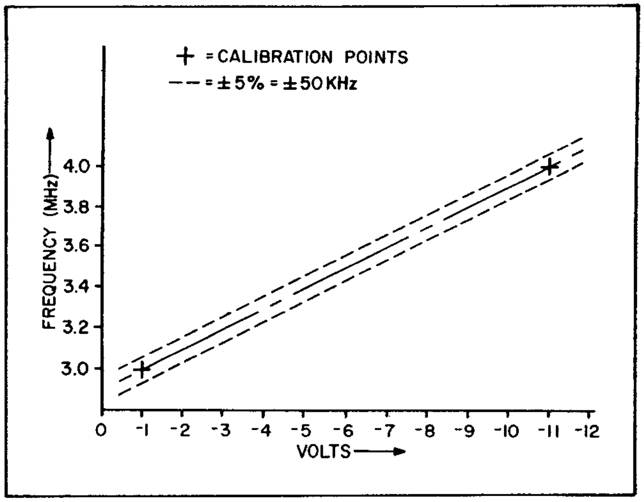
For suggestions on remote programming equipment, refer to the Bibliography at the end of this note. Also, you may wish to consider the Hewlett-Packard Model 2759A Synthesizer Programmer. This equipment was designed to operate with HP computer systems using plug-in cards for interfacing. The Model 2759A can be programmed to all columns simultaneously or sequentially with proper time delays in order to minimize switching transients. The latter feature is highly desirable in many applications and will be more fully understood by consulting Section IV of this Note.

Search Oscillator

The search oscillator may be varied between 3 and 4 MHz to provide continuous tuning and it may be frequency modulated from an external source at a maximum allowable sine wave rate of 1 kHz (for the voltage control calibration to hold). The output of this oscillator is controlled by either the front panel dial or by the application of a dc voltage. Specified dial accuracy is ± 3%, and the typical voltage vs. frequency characteristic is shown in Figure 2-10. The LC type oscillator is such that the characteristic will vary from the idealized slope of 100 kHz/volt; however, linearity with external voltage control is specified within ± 5%. Consequently, if a well defined 5 volts dc is applied, the search oscillator will be within 50 kHz of the desired 3.4 MHz (the 5% is based on the total 1 MHz swing). You may want to use a counter to determine the exact frequency produced by a particular external voltage input. Repeatability then becomes a function of how accurately it is possible to return to a given control voltage. Though careful design attention has been paid to minimizing the effects of temperature, component aging, etc., the voltage vs frequency characteristic may change slightly over an extended period of time. For certain applications it will be desirable to periodically re-determine the required control voltages for optimum repeatability.

When the search oscillator is used, the stability of the synthesizer output frequency is determined by either that of the standard instrument or that of the search oscillator -- whichever is less. The search oscillator contributes a root-mean-square deviation of approximately 1 Hz to the output frequency when used in the $10^5$ column of any of the synthesizers. When it is used in the $10^6$ column of the 5105A or 5103A the root-mean-square deviation contribution becomes 10 Hz at the output of the synthesizer. The rms deviation is arrived at by using a one-second averaging time and is symbolically denoted $\Delta f_{rms}$. If the search oscillator is used in the $10^4$ column, $\Delta f_{rms}$ = 0.1 Hz. If it is used in the $10^3$ column $\Delta f_{rms}$ = 0.01 Hz, etc. If this value is normalized by the output frequency, the result is the rms fractional frequency deviation. As an example, for the 5105A consider that the search oscillator is used in the 100 kHz step column at an output frequency in the 100 MHz region. The instability in the output due to the search oscillator is then:

Figure 2-10. Search Oscillator Linearity




$$\frac{\Delta f_{rms}}{f_{out}} = \frac{1.0\ \text{Hz}}{100 \times 10^6\ \text{Hz}} = 1 \times 10^{-8}$$

At this output frequency, and using one second averaging, the short term stability due to the synthesizer itself is on the order of $\pm 2 \times 10^{-11}$. Therefore, the search oscillator governs in this case.

Modular Construction.

In the theory of operation the synthesizer was presented by functional units - the synthesizer driver, the hf section, and the uhf section. This approach is also used in the construction of the synthesizer.

Solid-state modular construction has been used thruout the synthesizers which enables the system to meet stringent demands regarding spurious signals since the isolation it affords minimizes spurious coupling. All modules are interchangeable between synthesizers and the decades used in the least significant columns are interchangeable between columns. For example if the $10^3$ module were to fail in the 5105A, it could be replaced by the $10^{-1}$ module. The $10^{-1}$ module could then be bypassed by a jumper cable resulting in only a loss of resolution in the 0.1 Hz decade.

Modular construction allows considerable testing of circuit modules before assembly of the instrument. This pre-test step and the package and circuit design result in an instrument that meets HP's stringent operating tests.

# SECTION III

# FREQUENCY STABILITY AND SPECTRAL PURITY

We have seen that HP synthesizers derive their stability from a single quartz crystal oscillator. It is not surprising, then, that the same terminology which has been used historically to describe the merit of precision frequency standards can be used to specify synthesizer stability. However, for most synthesizer applications phase noise and spectral density plots are a more useful description of the signal stability and in fact are the preferred methods for specifying synthesizer short term signal stability.

Let's begin a discussion of the stability specifications by defining a few terms.

## Frequency Stability

There are two commonly referred to types of stability --long term and short term. Mathematically, instantaneous frequency may be defined as:

$$f(t) = \frac{d\phi(t)}{dt}$$

where $\phi$ is the instantaneous phase.

Average frequency over a period of time $t_0$ is then:

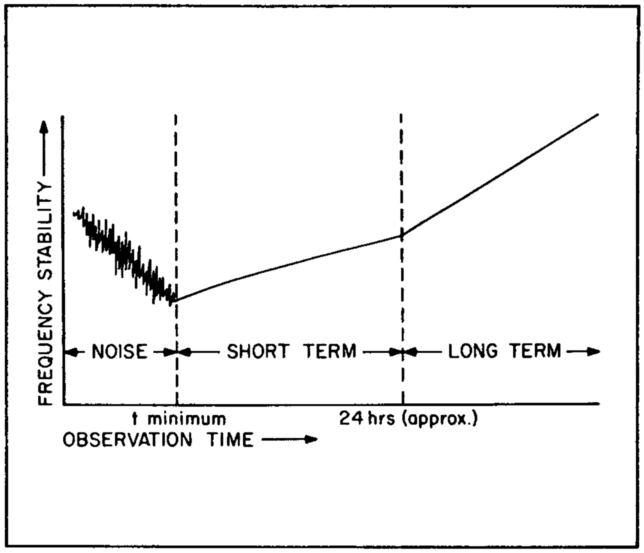
$$\overline{f(t)} = \frac{1}{t_0}\int_{t_1}^{t_1 + t_0} f(t)\, dt$$

Long term stability refers to slow changes in average frequency with time due to secular changes in the resonator or other elements of the oscillator. This is usually expressed in fractional parts per unit of time such as $\pm 3 \times 10^{-9}$ per day and is termed the crystal aging rate. In the case of a synthesizer this long-term stability is completely determined by the driving standard.

Short term stability refers to changes in average frequency over a time sufficiently short (but greater than some minimum time) so that the change in frequency due to long term effects is negligible. This situation can be shown graphically as it is in Figure 3-1. It can be seen that for times less than the minimum mentioned above, the variation in frequency can be considered as being caused by noise or incidental fm. The "pure" short term effects are due to variation of operating parameters such as supply voltage, line voltage, load, humidity, and ambient temperature. Since no clear-cut definition exists for the "minimum time" that we have previously referred to, it is our practice to include the effects of incidental fm with the effects produced by variations in the above parameters and term the resultant the short term stability.

A graphical model of the effects that various factors of stability have on the theoretically pure sinusoid of the synthesizer can be developed. To do this, we

Figure 3-1. Frequency Stability




first establish the axes shown in Figure 3-2. The vertical axis represents the imaginary component and the horizontal axis represents the real component. By Euler's equation:

$$|E_c| e^{j\phi} = |E_c| \left[\cos\phi + j \sin\phi\right]$$

the projection of the rotating vector on the real axis when plotted against time will represent a cosinusoid. We consider the vector to be rotating at an angular frequency of $\omega_c$, the frequency of the oscillation.

We now introduce a representation for noise. In the first place, noise is a random process. That is, it cannot be fully defined prior to its application. However, it is generally considered to be composed of an infinite number of frequency components. Mathematically, the expression

$$\sum_{n=\infty}^{n} c_n \cos(\omega_n t - \phi_n) = E_n(t)$$

has found rather widespread use. To avoid the mathematical expansions involved, we can think of noise as being represented by an infinite number of vectors each rotating with a specific speed and having a phase with respect to some reference. The addition of these vectors at various instants in time will yield resultants which have specific phases and magnitudes (Fig. 3-3). Progressing from this instantaneous representation to one which involves the passage of time, we note that this resultant vector would effectively vary in amplitude and phase in a random manner. Because of this randomness, it is necessary to speak in terms of probabilities. In other words, during what percent of the time can we expect this resultant vector to have a

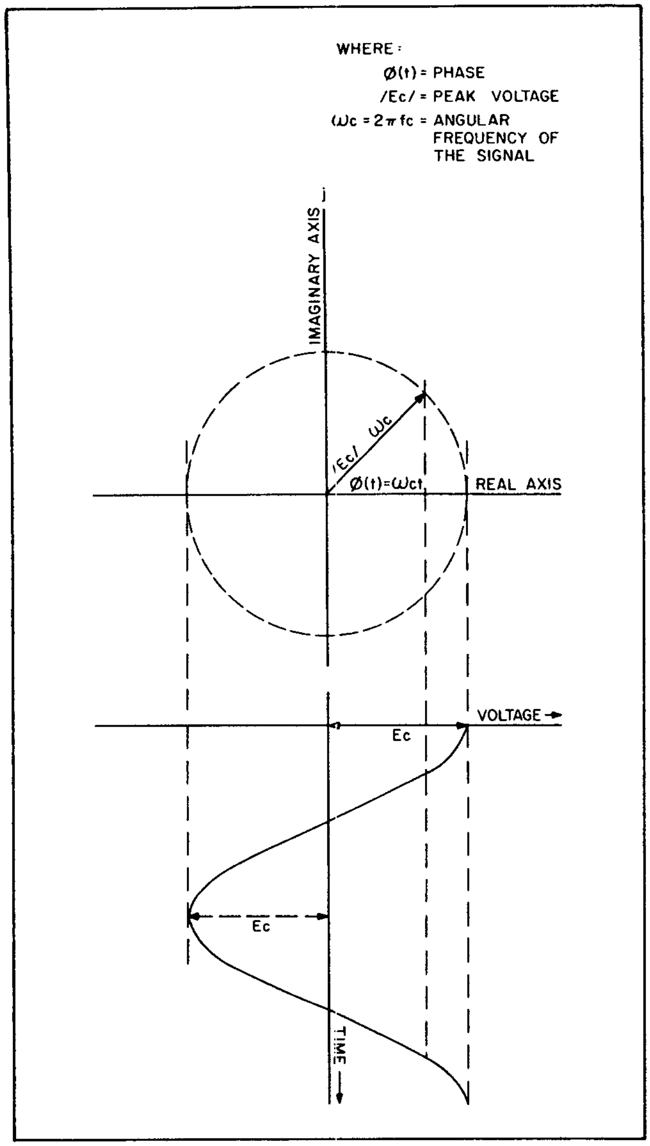



Figure 3-2. Cosinusoid From Rotating Vector

magnitude and phase which lies between some limiting values? The "central limit theorem" of the Theory of Probability tells us that the distribution of the sum of N independent random vectors approaches a "normal" or Gaussian law as N approaches infinity. This simply means that if we were to plot the frequency of occurrence (the vertical axis) of specific values of amplitude or phase (the horizontal axis), the graph would take the form shown in Figure 3-4. In this case, a mean or average value would be seen to exist at the center of the symmetrical distribution. The equation for this curve is:

$\mu$ = mean
x = abscissa
y = ordinate

$$y = \frac{1}{\sqrt{2}} \cdot e^{-\frac{1}{2}\left(\frac{x-\mu}{\sigma}\right)^2}$$

$\sigma$, commonly referred to as the standard or root-mean-square deviation, is a measure of the "spread" of the distribution. The proportion of the cases falling between two numbers or the probability of getting a value between two values is given by the corresponding area under the curve. The common statistical inferences are that 68.3% of all values will be less than one standard deviation on either side of the mean. 95.1% of all cases will vary by less than two standard deviations, and 99.7% of all cases will vary by less than three standard deviations.

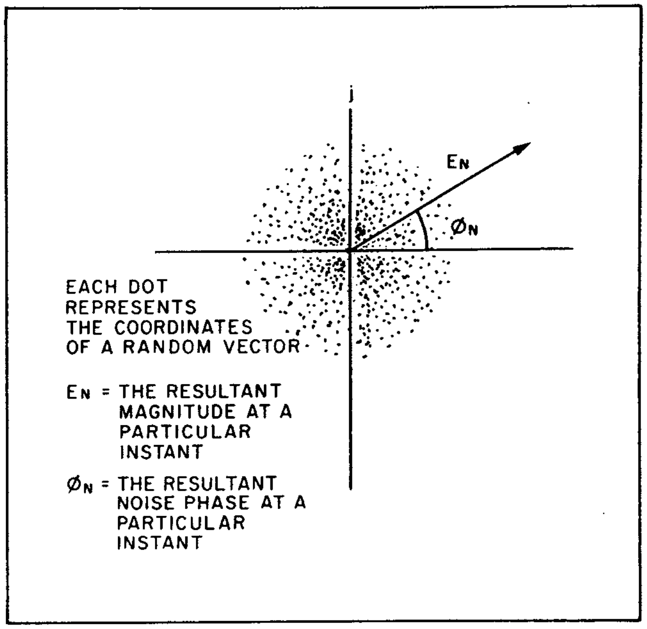



Figure 3-3. Vector Noise Representation

The basic question narrows to, "What is the standard deviation of the noise?" Examination of the mathematical expression for noise indicates that the average value of this distribution is equal to zero. However, if a large number of phase or amplitude measurements

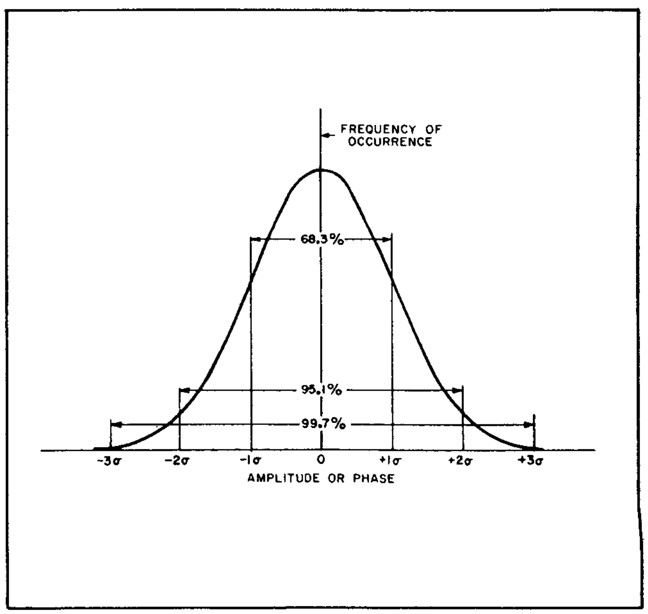



Figure 3-4. Gaussian Noise Curve

were made and the deviations of their values from the mean value were squared, then summed and divided by the number of measurements, the result would approach the square of the standard deviation. The greater the number of measurements made the closer the result would approach the true square of the standard deviation of the distribution. The important point to realize is that the very nature of the distribution dictates that some instantaneous values of phase or amplitude are more probable (or occur more frequently) than others.

The next step is to construct a graphical model for the effect that noise would have on the pure sinusoidal signal. If we consider the random noise distribution, and by extension its resultant vector, to be displaced from the origin by the length of the pure sinusoidal signal vector. The vectors shown in Figure 3-5 are a logical representation of phase and amplitude variations due to noise.

If, at each instant, the resultant noise vector is added to the carrier or signal vector, the resultant oscillator output signal is seen to fluctuate in amplitude and phase. As a matter of fact, the noise vector can be resolved into two components--one producing only an amplitude variation and the other producing only a phase variation. This vividly displays the random phase and amplitude variations.

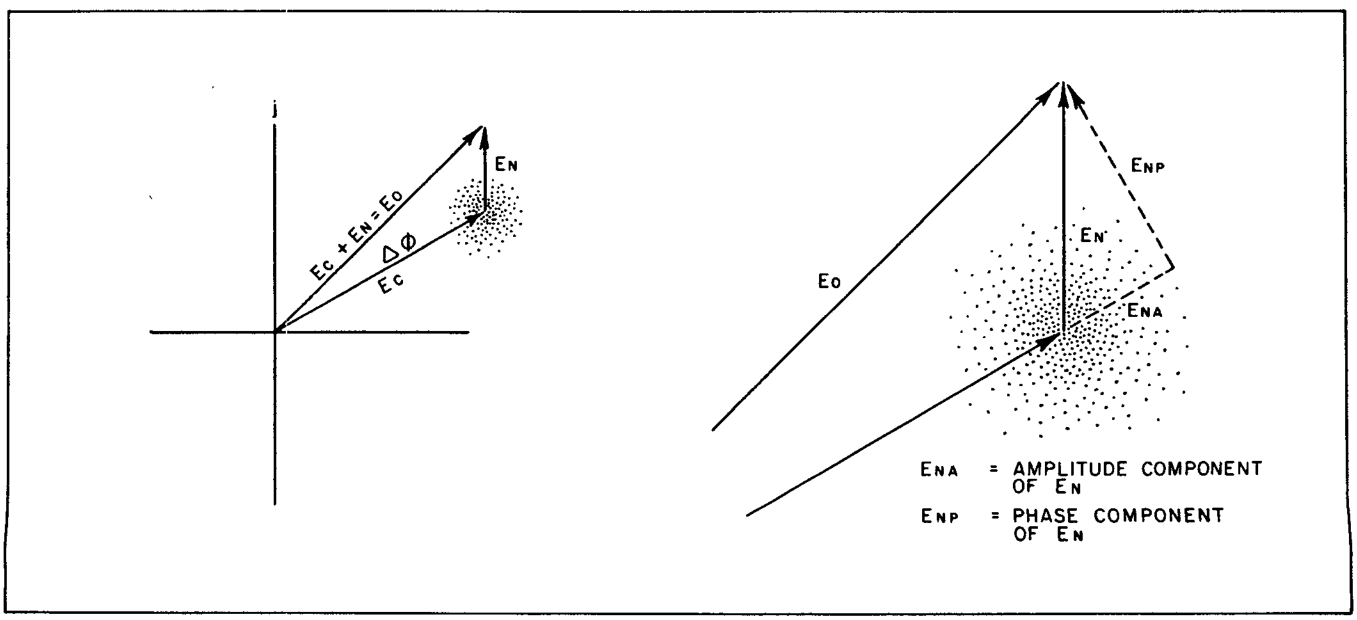
Detailed mathematical analysis shows that as the ratio of the magnitude of the carrier or signal to root-mean-square amplitude of the noise becomes large, the resultant distribution behaves like a normal law. The mean of the distribution becomes equal to the magnitude (in the case of the amplitude component) or the phase (in the case of the phase component) of the carrier and the standard deviation is equal to the standard deviation obtained for the noise distribution.

By examining the equation for the normal curve, we note that it is completely determined by the mean and the standard deviation. Consequently, once these values are determined for a particular output frequency, the common statistical inferences can be drawn. For example, if the standard deviation for the synthesizer output distribution at 50 MHz is .00015 Hz, we can say that 68.3% of all cases will differ from 50 MHz by less than that value. 95.1% of all cases will differ from 50 MHz by less than .00030 Hz. 99.7% of all cases will differ from 50 MHz by less than .00045 Hz.

## Frequency Spectrum

In addition to vector representation, the more useful and more easily measured "frequency spectrum" of

Figure 3-5. Phase and Amplitude Noise Variations

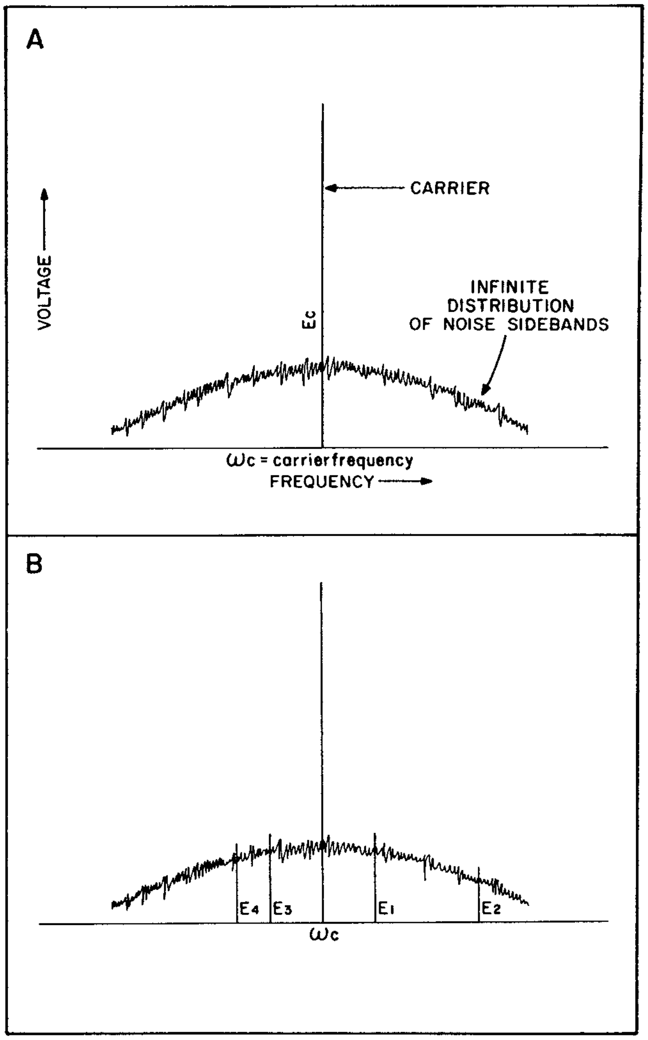


the distribution can be developed. Here it is helpful to think of an infinite number of sidebands. According to phase modulation theory, the magnitudes of these sidebands are mathematically determined by Bessel Functions of various kinds and orders. If the ratio of carrier magnitude to rms noise magnitude is large as stated previously, the only sidebands that have appreciable magnitudes are those caused by the noise components respectively interacting with the carrier --the interaction between noise components is extremely small. Under these circumstances, the frequency spectrum plot has the form shown in Figure 3-6A. The infinite number of noise components are represented by the continuous level extending across the base of the carrier.

When two signals are mixed to produce a desired output frequency, components in addition to the sum and difference frequencies are obtained. These intermodulation products are due to imperfections in the mixer characteristics and may be either harmonically or non-harmonically (spurious signals) related to the output frequency. In either case, they will appear as discrete signals in the output frequency spectrum. For the complete spectrum of the output signal, these signals have to be added to the random noise spectrum (Figure 3-6B).

By considering the rotating vector representation, it is not difficult to see that the presence of spurious signals further degrades the output signal. For the sake of simplicity, we shall only consider a single spurious signal as shown in Figure 3-7. Notice that the instantaneous orientation of this signal causes further variation in the Phase ($\Delta \phi$) of the output signal. It is apparent that the total instantaneous frequency variation produced by spurious signals and noise depends on their relative magnitude and instantaneous orientation.

Figure 3-6. Frequency Spectrum




## Phase Noise Evaluation.

As mentioned in the previous section, an extremely useful technique in the measurement and specification of the short term stability of a signal source such as a frequency standard or synthesizer is the plot of phase noise and spurious signals versus frequency offset from the carrier.

For high quality signal sources the amplitude modulation is assumed to be much smaller than the phase modulation. (In HP synthesizers and standards this is a perfectly valid assumption.) If this is not the case, a similar limit plot for amplitude modulation sidebands would be of interest because of its direct effects and easy conversion to phase modulation in the processing of the signal.

The effects of phase noise on a signal may be represented by a vector diagram as shown in Figure 3-8. Since the noise contributions being observed are truly random, they will appear as numerous symmetrical sidebands around the carrier, $E_c$. Looking at a single pair of sidebands $e_{SL}$ (lower) and $e_{SU}$(upper), the peak deviation contributed by these can be shown to be 2 $e_S$. (Since they are symmetrical pairs, their absolute magnitude and instantaneous angular velocity with respect to the carrier is identical.) The resultant signal is $E_H$. For small values of $\phi$:

$$E_H = E_c$$

and

$$\sin \phi = \phi \text{ (in radians)}$$

From trigonometric relations it follows that

$$\phi = \frac{2e_S}{E_H} = \frac{2e_S}{E_c}$$

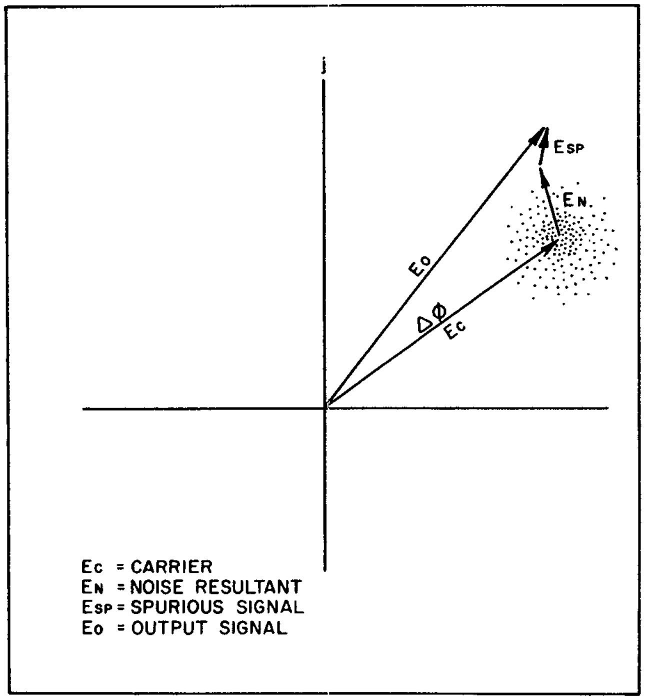
and

$$e_S = \phi\, E_c/2$$

The sideband distribution, then, consists of symmetrical pairs whose relative amplitude compared to the carrier is equal to 1/2 of the peak phase deviation of that component in radians.

In this measurement, synchronous signals are compared by means of a phase detector. The instrumentation setup shown in Figure 3-9 is an example of a typical system. The output of the phase detector, $e_N(t)$

Figure 3-7. Vectors of Single Spurious Signal




is the instantaneous voltage analog of the phase noise contribution under test. For the phase detector to be held to a zero output except for the phase noise contributions, the oscillator under test must be kept in quadrature with the reference oscillator. This is accomplished by using a direct coupled amplifier to sense a zero phase detector output and drive the test oscillator (source 1) to phase quadrature.

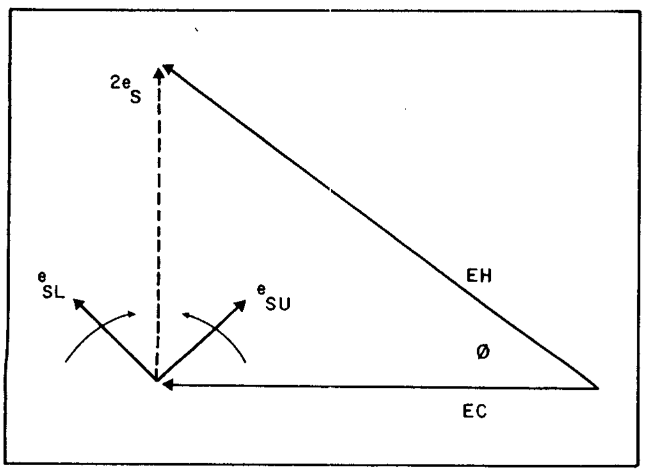
The phase noise will now be represented by a voltage out of the phase detector, $e_N(t)$, that is related to the phase noise by some constant.

$$e_N(t) = A \cos\left[\pi/2 + \phi_N(t)\right] = A \sin \phi_N(t),$$

for small values of $\phi$, $e_N(t) = A\, \phi_N(t)$.

This constant, A, can be determined by the use of a calibration source with an offset. The attenuator

Figure 3-8. Vector Representation of Phase Noise Modulation




(set to 60 to 80 dB) is used in the calibration procedure to prevent overloading the low noise amplifier and is set to 10 dB for the measurement on the test oscillator (to maintain linearity in the phase detector). When the two sources have the same frequency and are in quadrature phase position, the output phase noise can be monitored with a low frequency analyzer for phase noise plots, or a band-limited voltmeter can be used for a gross measure.

The plot obtained on the X-Y Recorder is easily calibrated since A has been determined for $e_N$ and $\phi_N$. The sideband amplitude ($e_S$) at a particular offset frequency (f) can be expressed:

$$e_S = \phi_{Nf} E_c/2 = e_{Nf} E_c/2A,$$

since $\phi_{Nf} = \dfrac{e_{Nf}}{A}$ and

where $\phi_{Nf}$ = Phase noise at a particular offset frequency,

$e_{Nf}$ = Phase noise analog at a particular offset frequency,

$E_c$ = Main signal or carrier amplitude,

A = Constant.

Since it is desirable to express the resultant plot as rms phase noise and the wave analyzer is an average reading device that measures rms noise 1 dB low, a correction of 1 dB must be added to the plot for the average-to-rms conversion. The resultant plot may also be converted to a single-sideband phase noise plot by subtracting 6 dB (for a total of minus 5 dB for conversion to a single-sideband rms phase noise plot). Figure 3-10 is a phase noise plot obtained for an HP 5103A 10 MHz Synthesizer using a test setup similar to that shown in Figure 3-9.

There are several advantages in using the phase noise versus offset frequency technique for obtaining short-term stability measurements and spectrum plots:

1. The test setup and calibration procedures are simple and straightforward.

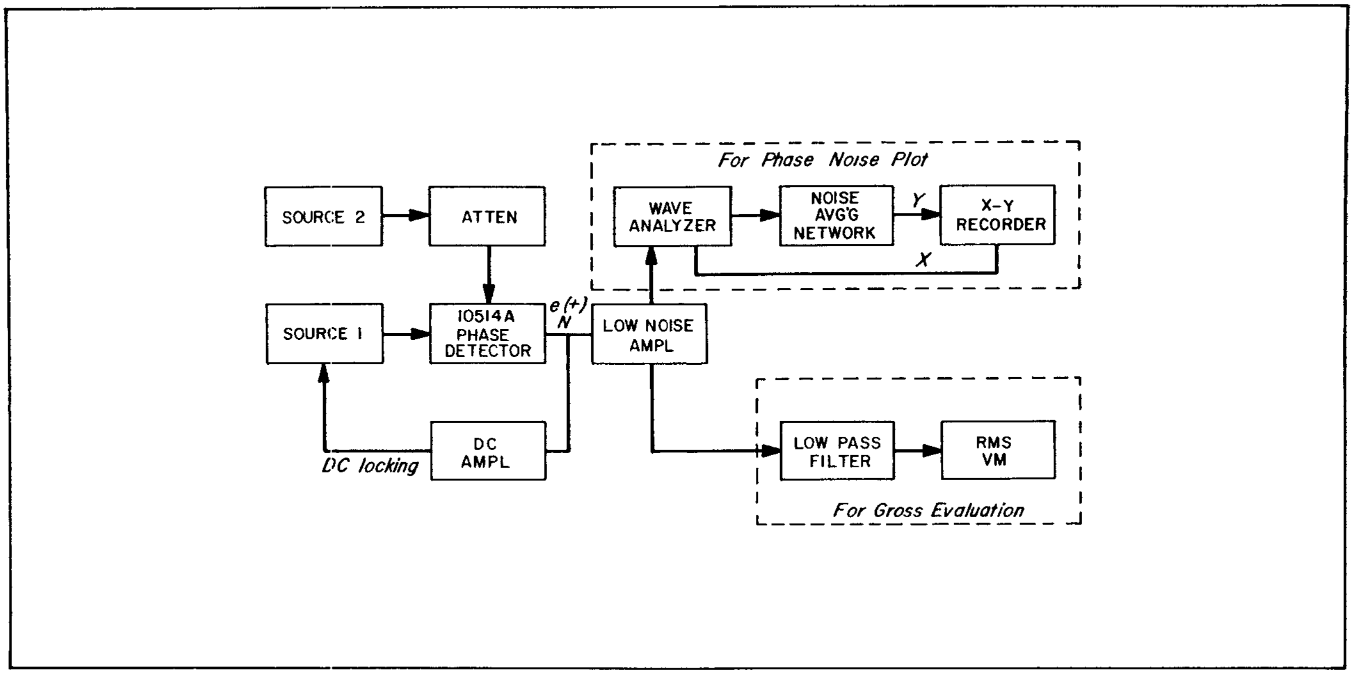
2. The phase noise and spurious signals in any band of interest may be studied by this measurement procedure.

3. The results obtained will allow the necessary compensations to be made for unwanted noise in any system where the oscillator being tested might ultimately be used.

4. The measurement technique outlined here is capable of measuring the noise characteristics of the best quartz oscillators available today.

5. Phase noise vs offset frequency plots yield sufficient information to calculate all the other commonly quoted specifications used to describe short term stability (i.e., rms fractional frequency deviation, rms incidental fm, rms phase noise, etc). The inverse calculation, however, is very difficult and in most cases not possible. For further information see Ref. 3.

Figure 3-9. Frequency Stability Measurements on High Quality Signal Sources



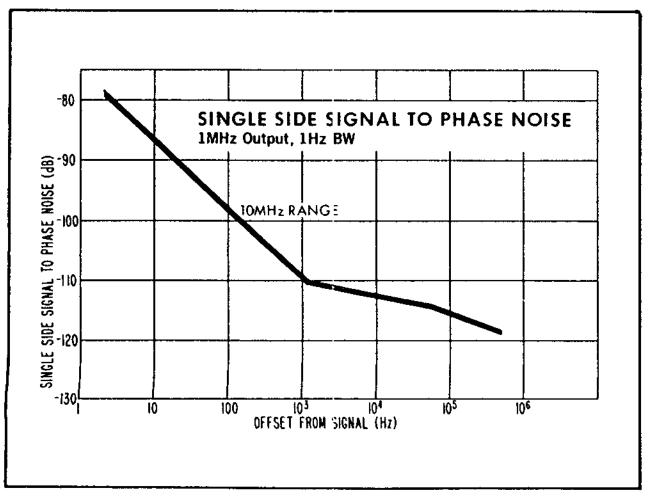

6. A disadvantage of this method is that for events occurring slower than 1 s or for spectral components closer than 1 Hz to the carrier the information is difficult to obtain. A superior method for obtaining this information is the rms fractional frequency deviation measurement which is discussed in detail in a later section.

Significance of Good Spectral Purity

For more information leading to possible applications for the synthesizer, the effects on a signal when it is multiplied into the microwave region will be briefly discussed.

Even a very ordinary oscillator will have a reasonably good spectrum close to the frequency of oscillation.

Figure 3-10. Frequency Synthesizer Phase Noise Plots




The spectrum rapidly degrades with frequency multiplication, however, so it is necessary to start out with an extremely good spectrum to have an acceptable spectrum after the fundamental frequency signal is multiplied into the microwave region.

The total power in a frequency-modulated wave is constant. If the frequency multiplying device is broadband, the ratio of total sideband power to signal power increases as the square of the multiplying factor. Consequently, the increased sideband power must come from the carrier. The spectrum of the signals begins to spread since the increased sideband amplitude causes intermodulation between sidebands to become appreciable. It is now obvious why the highest possible signal-to-noise ratio for the original is desired.

An expression for the spectrum of the multiplied signal may be given in terms of the ratio of total power in the observed sidebands to the carrier power.

$$\frac{P_N}{P_S} = n^2 \frac{P_{ON}}{P_{OS}}$$

where $P_N$ = total sideband power in the multiplied frequency.

$P_{ON}$ and $P_{OS}$ = initial (before multiplication) sideband and signal power.

n = frequency multiplication factor.

The above formula is valid only if the ratio $P_N/P_S$ is very much smaller than unity. To have a narrow spectrum with good signal-to-noise ratio after high order frequency multiplication, the initial signal-to-noise ratio must be very good and/or a narrow band filter must be used between the oscillator and the multiplier. As a result of this formula, the synthesizer noise specifications are degraded by 6 dB per octave or 20 dB per decade by "ideal" frequency multiplication.

## Spurious Measurement

As previously discussed the undesired signals -- noise, and spurious responses -- produce an effective phase modulation and consequently, a frequency instability of the output signal. The frequency spectrum of the type just discussed for the synthesizer at a specific output frequency, indicates the stability and spectral purity of the signal at that particular frequency. Each of the possible output frequencies will have a slightly different spectrum. By examining the specifications and some typical data, we can gain some insight into the general form of the spectra.

For example, the 5100B specification states that spurious signals are more than 90 dB below the selected output frequency. Consequently, we can construct a "generalized" plot of the type shown in Figure 3-11. Notice that the amplitude of each spurious component can be no greater than 1/31,623 of the carrier amplitude. Note also that spurious signals are finite frequencies which are nonharmonically related to the carrier and may appear as either symmetrical sidebands or as single components.

Figure 3-12 shows the type of system used in determining spurious levels. It is interesting to note that this system must be extremely sensitive and have a narrow post detection band in order to discriminate the spurious signals from noise. As a matter of fact, this system has a dynamic range of 120 dB and a sensitivity of 1/10 of a microvolt into 50 ohms. And because of the wide band mixer, it needs no recalibration as the frequency is changed.

## Signal-to-Phase Noise

We specify signal-to-phase noise ratio of the 5100B as being more than 54 dB down in a 30 kHz band centered on the signal excluding 1 Hz centered on the carrier. The system used for plotting this information is shown in Figure 3-13A. With the switch closed, noise due to the internal standard is correlated out. With the switch open (both synthesizers operating on their internal standards) the effect of the internal standard is taken into account and the signal-to-phase noise ratio measured is reduced by 3 dB under the assumption that both units are contributing equally to the noise spectrum. The validity of this assumption can be verified by additional comparisons with a third unit if there is lack of confidence in the results.

Figure 3-11. Generalized Spurious Response

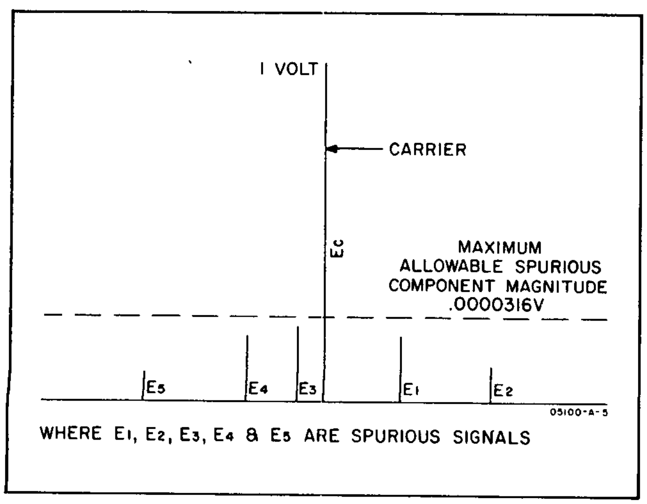



Figure 3-12. Spurious Measurement System

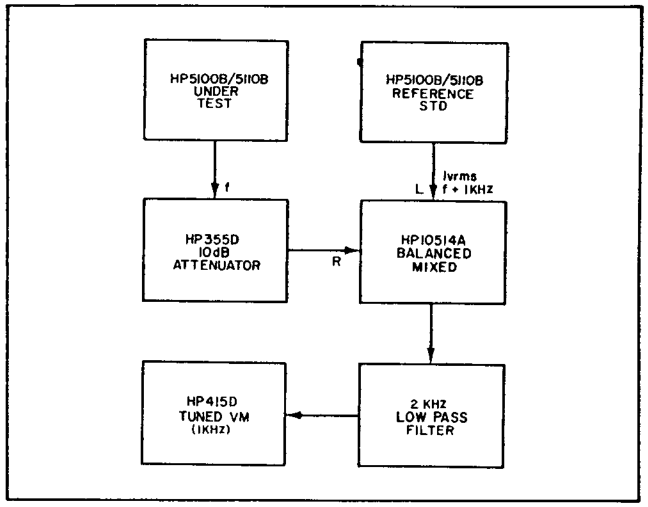
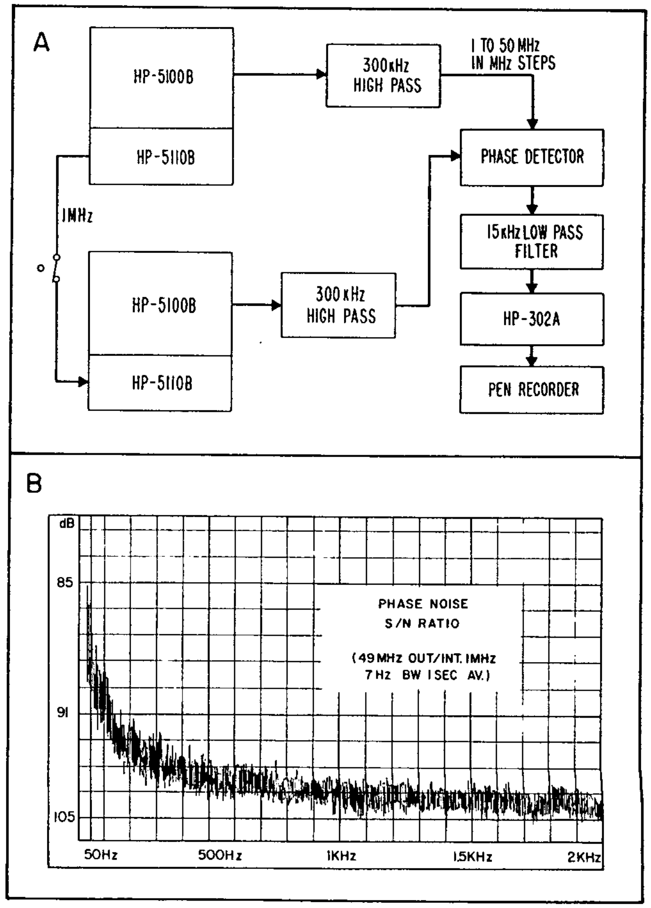



Figure 3-13. Phase Noise Plotting System

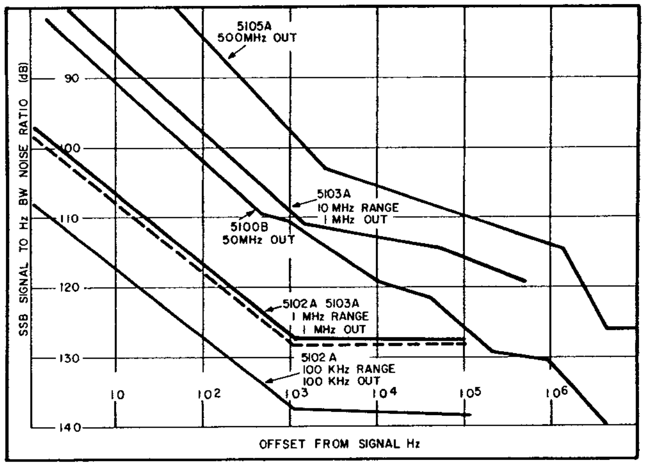


The signal-to-phase noise ratio specifications for all HP synthesizers includes the effects of the standard. When this method is employed, for a particular output frequency, the resulting plot has the appearance shown in Figure 3-13B. Conceptually, the representation can be rotated around the carrier to obtain a "spectrum" as shown in Figure 3-14. The implication is that the ratio of the total noise energy contained in a band which extends 15 kHz on either side of the carrier to the energy of the carrier is 1 to 250, 000.

The phase noise versus offset frequency plots of Figure 3-15 give a comparison of the signal to phase noise characteristics of all four Hewlett-Packard direct synthesizers. For lowest phase noise at a particular frequency the specifications of the synthesizers should be compared, considering the effect of signal multiplication on noise. For instance, using a 5100B below 100 kHz as compared to a 5102A sacrifices 20 dB of signal to phase noise ratio. However, at 50 MHz the 5100B provides a 14 dB improvement in signal to phase noise ratio as compared to a 5102A with an ideal (X50) multiplier.

Signals which are harmonically related to the output signal of the 5100B are specified to be at least 30 dB down. These signals, however, are well removed from the carrier and could effectively be removed by additional filtering if desired. Subharmonics of the carrier fall into the spurious category and consequently are at least 90 dB down.

Figure 3-15. Composite Phase Noise Plot for HP Synthesizers

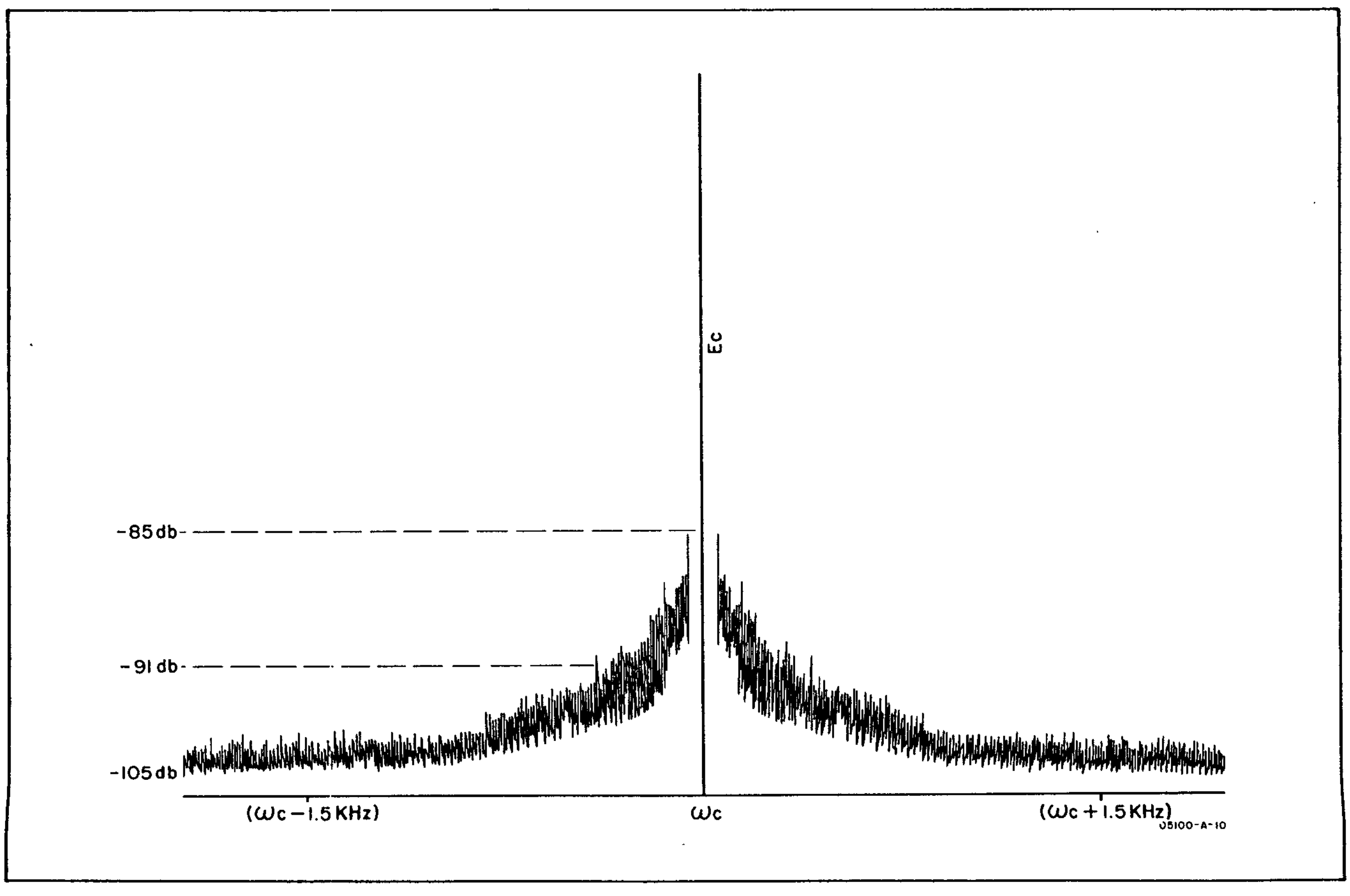



## RMS Fractional Frequency Deviation

In an application for the synthesizer for which the events of interest occur at 1 s or longer intervals it is more usual to describe the stability of the signal in terms of its rms fractional frequency deviation.

The system shown in Figure 3-15 is used to determine the "root-mean-square-fractional frequency deviation". Consider that the synthesizer's output frequency

Figure 3-14. Phase Noise Spectrum



is 20,000,100.00 Hz which is then mixed with the 20th harmonic of the internal 1-MHz standard. Mixing will result in the generation of a difference frequency which will nominally equal 100 Hz. There are two things to notice at this point. First, the measurement system is such that the instability of the internal 1 MHz standard is effectively correlated out. Second, the 100 Hz signal will not, in fact, be equal to exactly 100 Hz since any noise or spurious signals introduced by the synthesizer itself will cause the phase (and therefore the frequency) of this "beat note" to vary. A multiple period average measurement can be made of this difference frequency with a frequency counter and the result printed out on a digital recorder. These are all the data that are required; the next step is to develop mathematical formulae. The mathematical relationship used to relate frequency and time deviation is:

$$\left|\frac{\Delta f}{f}\right| \simeq \left|\frac{\Delta t}{t}\right|$$

For our purposes, f will represent the offset frequency of 100 Hz and $\Delta f$ will represent the deviation of the offset frequency from that value giving the expression:

$$f' - \bar{f} = \Delta f$$

where f' is the observed value of f. $\Delta f$ then, is the frequency deviation from the mean or average value of f. Consequently, $\Delta t$ is the observed period deviation from 10 milliseconds. Taking each value that has been printed out and subtracting 10 milliseconds from it, the result is a series of mean deviations which may be used to calculate the rms or standard deviation. This can now be expressed in terms of frequency:

$$\frac{\Delta t_{rms}}{t} = \frac{1}{t}\sqrt{\frac{\Sigma(\Delta t)^2}{N}} = \frac{\Delta f_{rms}}{f}$$

$$\Delta f_{rms} = \frac{1}{t}\sqrt{\frac{\Sigma(\Delta t)^2}{N}}$$

Normalize $\Delta f_{rms}$ by dividing it by F, the value set on the synthesizer pushbuttons:

$$\frac{\Delta f_{rms}}{F} = \frac{1}{Ft^2}\sqrt{\frac{\Sigma(\Delta t)^2}{N}}$$

The value of $\Delta f_{rms}/F$ is called the rms fractional frequency deviation (see note below). Notice that the specification as written may be interpreted as "parts in $10^{x}$". Thus, it is a stability specification and is thus a justifiable measure of the short term stability. Because of the averaging times and number of samples required to get meaningful data, it is clear that this statistic contains not only the effect of noise, and spurious signals, but also that of the other short term effects. It should also be realized that this short term stability must be considered in conjunction with the short term stability of the driving standard.

Note

The long term drift or long term stability must be removed from the data, as $\Delta f$ must be the deviation from the average f with no long term effects present. This requires in most cases for the data to be processed by a computer to obtain a least squares fit of the data to an average frequency curve.

Data for the 5100B Synthesizer, for example, gives an rms fractional frequency deviation of 1 x 10-11 at 50 MHz using a 1-second averaging time. The rms frequency deviation or standard deviation is then

$$\frac{\Delta f_{rms}}{F} = 1 \times 10^{-11}$$

$$\Delta f_{rms} = 1 \times 10^{-11}(50 \times 10^{6}) = .0005\ \text{Hz}$$

This tells us that in 68.3% of all cases in which the 5100B Synthesizer is set to 50 MHz, variation due to the Synthesizer's operation on the standard signal is less than .0005 Hz.

Examination of Figure 3-17 indicates that the rms fractional frequency deviation is degraded at lower frequencies. The point to remember here is that the rms frequency deviation is constant and looks worse when expressed in the "parts in $10^{x}$" form of the fractional deviation.

Figure 3-16. Frequency Deviation Measurement

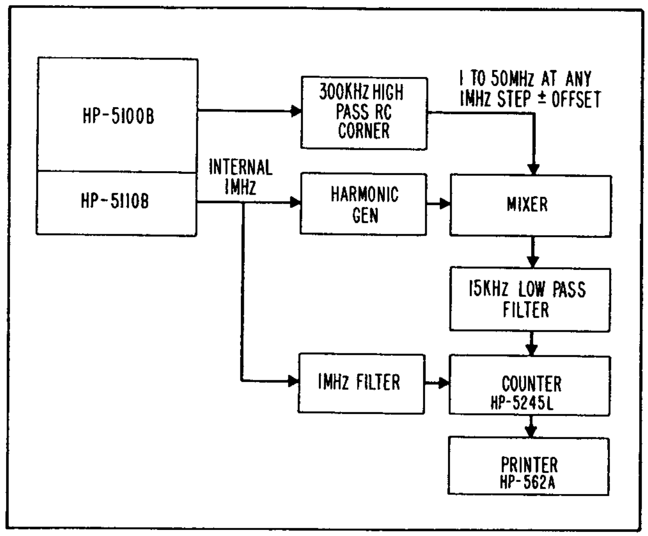
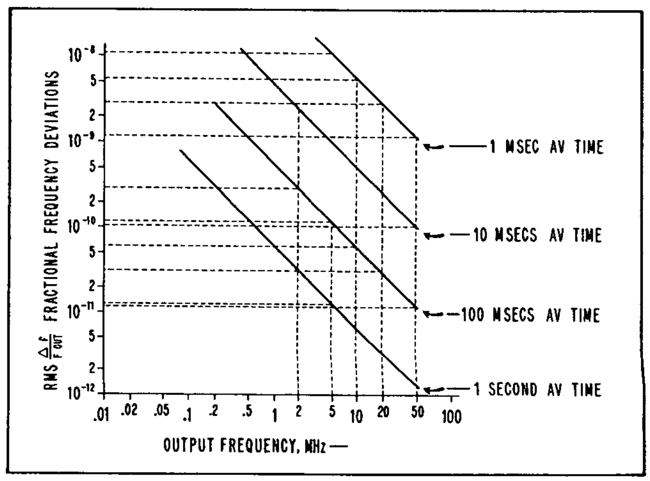



Figure 3-17. Rms Fractional Frequency Deviations of 5100B/5110B with 30 kHz Noise Bandwidth



# SECTION IV

# SWITCHING TIME LIMITATIONS

## Delay Times

With direct synthesis the limitations on switching speeds are set by the time constants in the filtering circuits on the control lines to the switches and by the circuit bandwidths. The specified typical switching time of Hewlett-Packard instruments says that most selected frequencies can be switched within 20 $\mu$s. By knowing what columns are to be switched, a reasonably accurate estimate of the actual switching time can be made.

Switching speeds of the switching matrix diodes by remote control are primarily limited by the charge and discharge time of the .01 $\mu$F filter capacitors from each control line to ground. With driving impedances below 50 ohms, both for on drive of -12.6V and off drive of +1.5 volts, this time is less than 1 $\mu$s. See the following section on "controller requirements" for additional details.

The remote control lines enter the synthesizer in parallel while the frequency development occurs in series. Thus, the principal period of frequency switching time is in the propagation delay through the decades if more than one column is changed at the same time (see Figure 4-1). If only one column is switched the time is virtually that of the switching diode networks for columns of $10^6$ per step and below, ($10^7$ on 5105A).

The bandwidth limiting circuitry in these decades causes the effect of a frequency change in one decade to be delayed at the output of the following decade by up to 4 microseconds relative to a change in that decade. This delay will momentarily cause an incorrect output frequency to occur whenever the incremental step change in output frequency is over one decade. The period of the incorrect frequency may last from 4 microseconds for switching adjacent decades, to 28 microseconds when the $10^5$ and $10^{-2}$ decades of the 5100B are switched. Its effect will be recurrent deviations from the desired frequency vs time function with the resultant formation of sidebands. These deviations will occur at decade submultiples of the switching rate, and their frequency and amplitude depend on the particular decade delays and switching rate used.

About one microsecond delay time occurs in the frequency injection from the hf section to the uhf section. This occurs between the $10^5$ and $10^6$ columns of the 5100B and the $10^6$ and $10^7$ columns of the 5105A. In these synthesizers, although the uhf section diode switching differs from the hf section, the delays introduced are similar.

An additional delay is introduced in the highest ($10^7$) column of Model 5100B by the switching on of the multipliers. A finite time elapses after switching in which phase error is present until a steady state phase condition is reached. The phase changes with time, interpreted as frequency changes for $10^7$ column switching at time $t_0$ are:

$$t_0 + 10^{-4}\ \text{s} \qquad f = f_0 \pm 2\ \text{kHz}$$

$$t_0 + 10^{-3}\ \text{s} \qquad f = f_0 \pm 1\ \text{Hz}$$

The highest ($10^8$) column of Model 5105A also has a unique time delay factor. Frequency is changed by switching the "drift cancelling" oscillators and is affected by a number of delay times: variations in path lengths of drift cancelling signals, switching characteristics of the 20-60 MHz signal from the spectrum generator, and the reaction time of the frequency multipliers. The switching action occurs within 2 $\mu$s followed by a period of phase changes similar to that of the 5100B.

Figure 4-2 indicates that the total switching time depends on which columns are switched. In the example shown, the frequency of a 5105A is switched from 80.2 MHz to 79.9 MHz. The output frequency was

Figure 4-1. Typical Delay Times for the 5100B

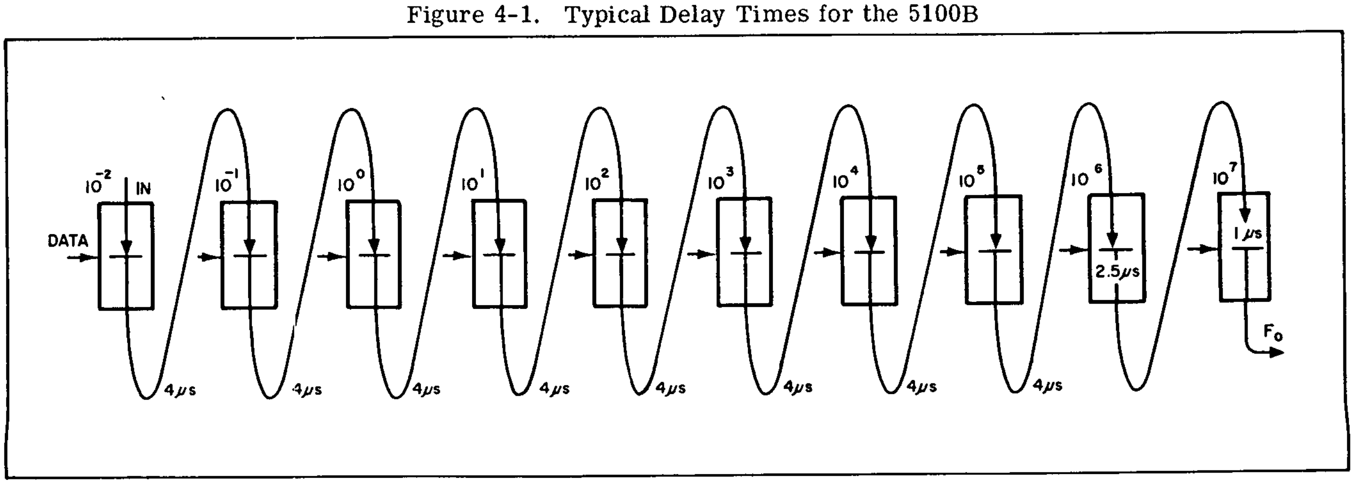
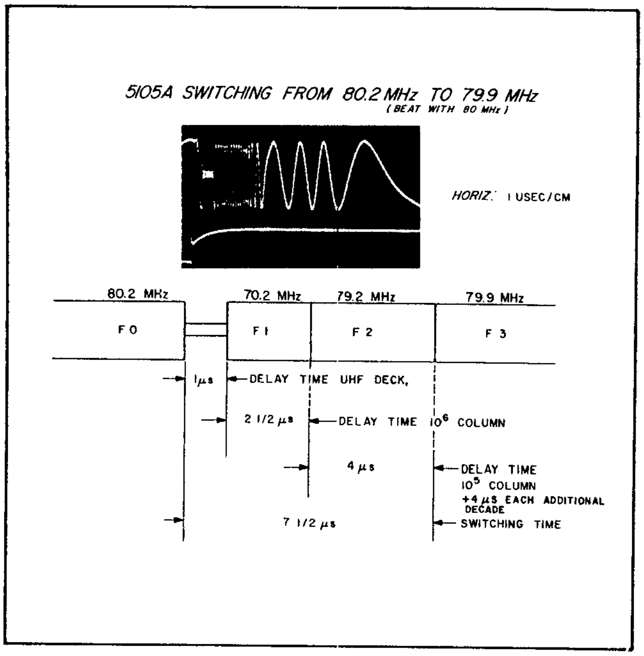


Figure 4-2. Switching Time vs Column Switched




mixed with an 80 MHz signal and then displayed on an oscilloscope. The oscillogram shows that there is approximately 1 μs delay time in the uhf deck. A 70.2 MHz signal (this appears as 9.8 MHz signal in the oscillogram) exists for approximately 2-1/2 μs due to the delay time in the $10^6$ decade. Then a 79.2 MHz frequency (this appears as 800 kHz in the oscillogram) exists for 4 μs due to the $10^5$ column decade delay. Thus, 7-1/2 μs is required to switch from 80.2 MHz to 79.9 MHz with the 5105A used in this demonstration. Another 4 μs delay would be contributed for each additional switched decade. So if the frequency in the $10^{-1}$ column had been switched in this example, an additional 24 μs would have been required to reach the desired output frequency.

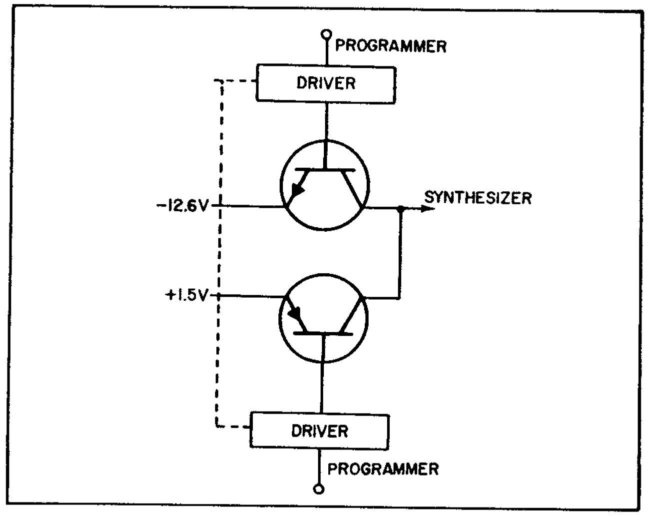
Controller Requirements.

Rapid switching of the synthesized output frequency may be accomplished by an external controller to generate the required control signals. They are applied to the synthesizer through the three, fifty-pin connectors including one pin for each pushbutton frequency selection. Each connector also supplies a ground and -12.6 volt line which may be used in remote programming. When switched to REMOTE operation all pushbuttons are disconnected, and all programming leads assume an off voltage (+1.5 V dc on $10^{-2}$ to $10^5$ decades).

The exact requirements on the control signal depend on the switching time that is to be achieved. However, in all cases activation of a desired digit is accomplished by application of -12.6 volts by the controller to the pin corresponding to that digit. In this case a simple transistor switch may be used to drive the control signal to the required -12.6 volts with recovery to the off state accomplished by fixed bias circuits in the synthesizer. The resulting internal RC time constant of recovery is sufficient for 1 millisecond switching and recovery to the off-state is assured if the controller presents an off impedance of 100 kΩ or greater.

For remote programming with minimum switching time (20 μs typical) a further requirement is placed on the control signal and controller. The controller must program both the "on" and "off" states -- the voltage level changing from -12.6 V dc (on state) to +1.5 V dc (off state) from a 50 ohm or lower source impedance. Rise and fall times should be less than 40 nanoseconds into an open circuit (to meet this requirement the length of the control lines between the controller and synthesizer should be minimized) and "on" digit control must be synchronous with "off" digit control. A circuit suitable for driving a synthesizer control is shown in Figure 4-3.

Figure 4-3. Remote Programming Driving Circuit




For load currents required by various synthesizer control lines see Table below.

Control Line Load Currents

| Column | Synthesizer | | | |
|---|---|---|---|---|
| | 5102A | 5103A | 5100B | 5105A |
| $10^8$ | | | | 25 mA |
| $10^7$ | | | 16 mA | 14 mA |
| $10^6$ | | 8 mA | 14 mA | 14 mA |
| $10^5$ - $10^2$ | 8 mA | 8 mA | 8 mA | 8 mA |

As internal circuitry is susceptible to lf ripple from the power source, stringent requirements on the power supply ensure that ripple-induced spurious signals will be 90 dB down (ripple < 10 μV). For remote programming the maximum allowable low frequency current that can be injected into the internal supply is 200 μA or, considering 10 columns are being programmed, 20 μA per column. This does not hold if an external -12.6 volt supply is used; the -12.6 volts may have up to 1 mV ripple. (See references at the end of this book.)

# SECTION V

# APPLICATION AREAS

Direct frequency synthesizers are used where rapid change of frequency through specific values is desired and a degree of stability and spectral purity comparable to that of precision fixed frequency standards is needed.

The following paragraphs will cover a number of specific examples where the use of frequency synthesizers will greatly enhance the measurement or operation.

## Nuclear Magnetic Resonance

Atomic or molecular resonance measurements require spectrally pure, precisely variable signals. Nuclear magnetic resonance methods are increasingly used to determine, among other things, the qualitative and quantitative make-up of materials. In this method the strength of an applied dc magnetic field and the frequency of a simultaneously-applied RF field needed to produce nuclear resonance in the material must be known. The dc magnetic field can be controlled at a defined reference value with great stability by previously developed means. The synthesizer now provides the rf excitation frequency at the high precision needed to greatly enhance the precision of NMR measurements. The ease of frequency control in the synthesizer allows for automatically testing for the presence and quantity of several elements in a sample and does so with such speed as to make NMR in-process control a real possibility.

## Space Doppler

This application requires a precisely controlled center frequency which must be continually changed to match the Doppler shift.

Determining the velocity of far-out space vehicles through Doppler frequency measurements involves operation at X-band with receivers having bandwidths of but a few cycles to minimize noise levels. As the vehicle velocity changes, the receivers local oscillator must be changed to keep the received signal in the center of the IF bandwidth. Here again the synthesizer is ideal because its frequency can be changed in known selectable increments.

## Secure Communications

The requirement here is for rapid precise frequency shifting in a known random manner to prevent interception.

The HP Synthesizer's spurious free performance makes it well suited to use as the master oscillator in a transmitter and as the local oscillator in a receiver. If the transmitter and the rf section of the receiver are untuned, an extremely fast switching system can be used to change the local oscillator (synthesizer) frequency to achieve communications systems of high integrity.

The synthesizer can greatly facilitate surveillance work if it is used as the local oscillator in a receiver designed to accurately determine the frequency of remote transmitters. The ease and speed with which the synthesizer frequency can be changed will allow monitoring of a multiplicity of channels with a single receiver by sequencing the local oscillator (synthesizer) through the desired channels.

Sequencing the synthesizer output through a group of desired frequencies can also permit a single instrument to operate as an automatic calibrator for a multiple transmitter installation. The arrangement can provide for phase-locking the transmitter frequencies to the synthesizer by a circuit with a time constant long enough to maintain the transmitter frequency for the duration of the sequencing cycle.

## Radio Sounding

The best transmission frequency must be determined for dependable hf long distance communications. In hf communications work, dependable long-distance communications requires the use of a frequency near the maximum useable frequency which is determined by ionospheric conditions. Since these conditions can change rapidly, test transmissions over the hf spectrum are used at frequency intervals to insure operation at the optimum frequency. The fast switching and electronic remote control of the synthesizer make it a natural part of such a "radio sounding" system.

## Stability Studies

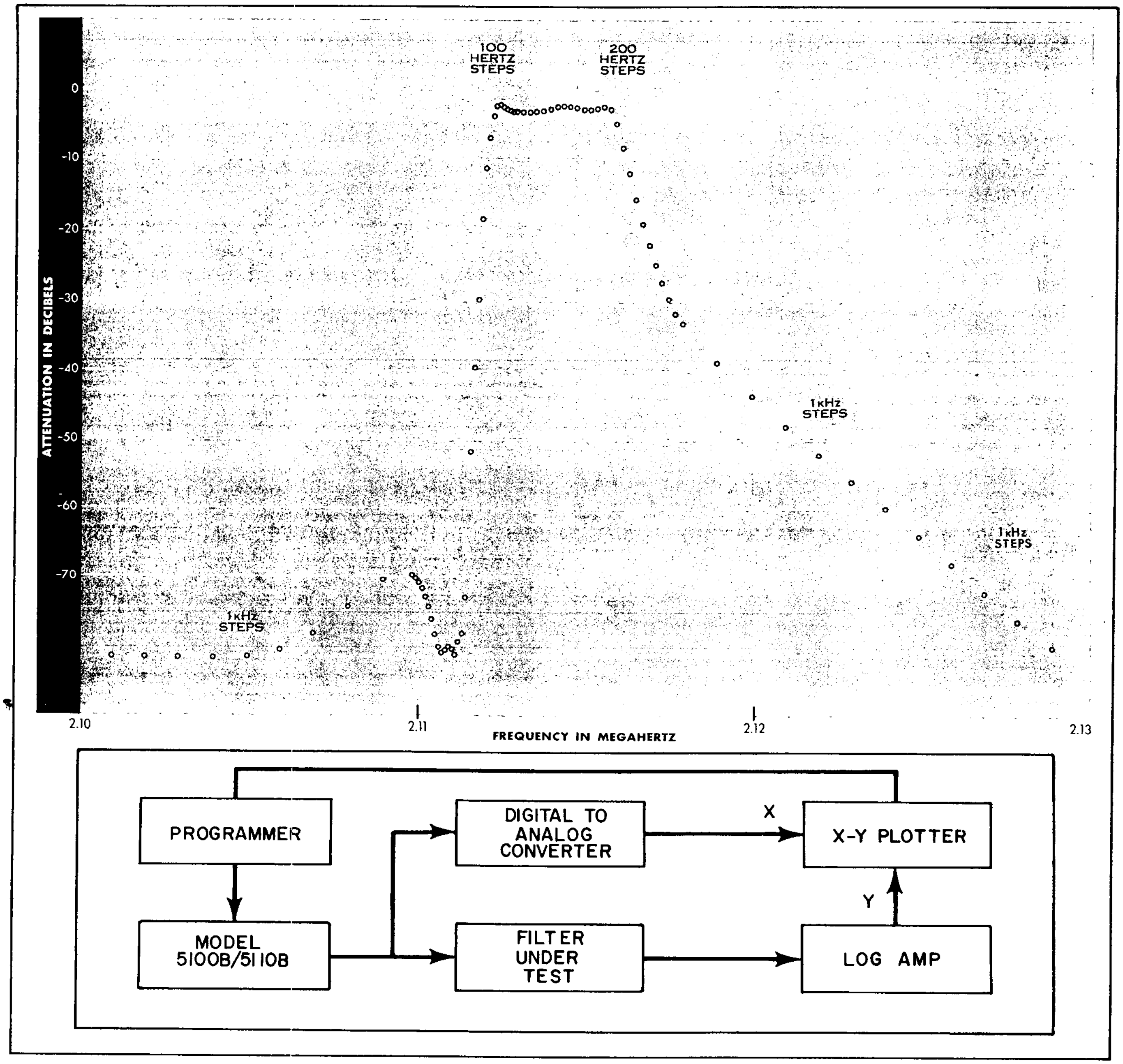
The excellent frequency stability of a synthesizer makes it useful as a standard in measuring the frequency stability of other signal sources. In frequency stability measurements the synthesizer signal can be subtracted from the signal under test, thereby translating instabilities of a source to a lower frequency where they can be measured by a frequency counter, low frequency analyzer or other means.

It is also interesting to note that the synthesizer can be used in measuring the phase stability of such devices as amplifiers, frequency multipliers, frequency dividers, trigger circuits and resonant devices. In such work a phase comparator is used to synchronously mix the input (supplied by the synthesizer for stability) and the output signal of the unit under test. By adjusting the phase of the signals to a quadrature relationship any phase perturbations introduced by the unit under test will be readily observable at the comparator output. This technique is very sensitive and powerful and can be used as a measure of the reliability of the device under test even if phase stability is not of special importance.

## Production Testing

Frequency sensitive devices can be rapidly, automatically tested with a precisely known, spectrally pure, frequency synthesizer. As an example of one of many applications made possible by a synthesizer's remote programming capability, the plot in Figure 5-1 shows the response of a single side band crystal filter. Less than one minute was required for the complete plot.

Figure 5-1. Crystal Filter Response Plot



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