AN 216

Digital Network and Spectrum Analysis Primer

A Guide To The Use Of The HP3570A And HP3571A Analyzers

APPLICATION NOTE 216

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Chapter 1
INTRODUCTION

The 3570A Network and the 3571A Spectrum Analyzers open many new areas to analysis with the joining of analog and digital technologies. Frequency synthesis and digital amplitude and phase measurements provide increased accuracy and resolution. However, as is usually the case when one strives for higher accuracy, effects which were previously insignificant now become major sources of error. The purpose of this application note is to investigate the sources of these errors, the magnitude of their effects and ways to minimize the problems they can cause.

Because it is the object of this note to explore new ground, it is assumed that the reader is familiar with the concept of the frequency domain and the basic principles of measurement behind *analog* network and spectrum analysis. If not, there are several excellent HP application notes on the subjects and these are listed in the bibliography at the end of this note. Also, throughout this note we are only discussing swept-tuned rather than real-time analysis. Real-time analysis is subject to quite different constraints and these are not discussed in the following chapters.

The organization of this note is divided into two parts, main text and appendices. The main text is further divided into chapters and sections to enable the reader to easily find the discussion of his particular measurement problem. The appendices contain the mathematical derivation of much of the material in the main text. This organization allows smooth development of the main discussion and yet presents detailed development of the arguments in the appendices for those who wish further background or desire to develop measurement criteria of their own.

What Are Network and Spectrum Analyzers?

Before we delve into the potential measurement problems, let us first attempt to define what network and spectrum analyzers are and what the differences and similarities are between them. As shown in Figure 1.1, both instruments have block diagrams similar to a super-hetrodyne receiver. Both could also be described as frequency-selective voltmeters. To understand the differences, let us look at the definitions of each instrument.

Traditionally, a spectrum analyzer has been thought of as a frequency-selective voltmeter or wave analyzer with a CRT display. However, because of the increasing need for improved accuracy and resolution, the 3571A uses digital readouts instead of a CRT. Only digital readouts can give both large dynamic range and high resolution, but if a graphic display is desired (with the corresponding reduction in resolution–dynamic range), a 'scope or X-Y plotter can be connected to the analog outputs. Obviously, since the CRT is optional, we need a new definition of what a spectrum analyzer is.

Definition: A spectrum analyzer is a frequency-selective voltmeter which measures all the frequency components of a signal in a frequency band. It is designed to allow it to sweep through this band as fast as is theoretically possible and to measure all signals accurately, i.e., not introduce extra apparent signals itself.

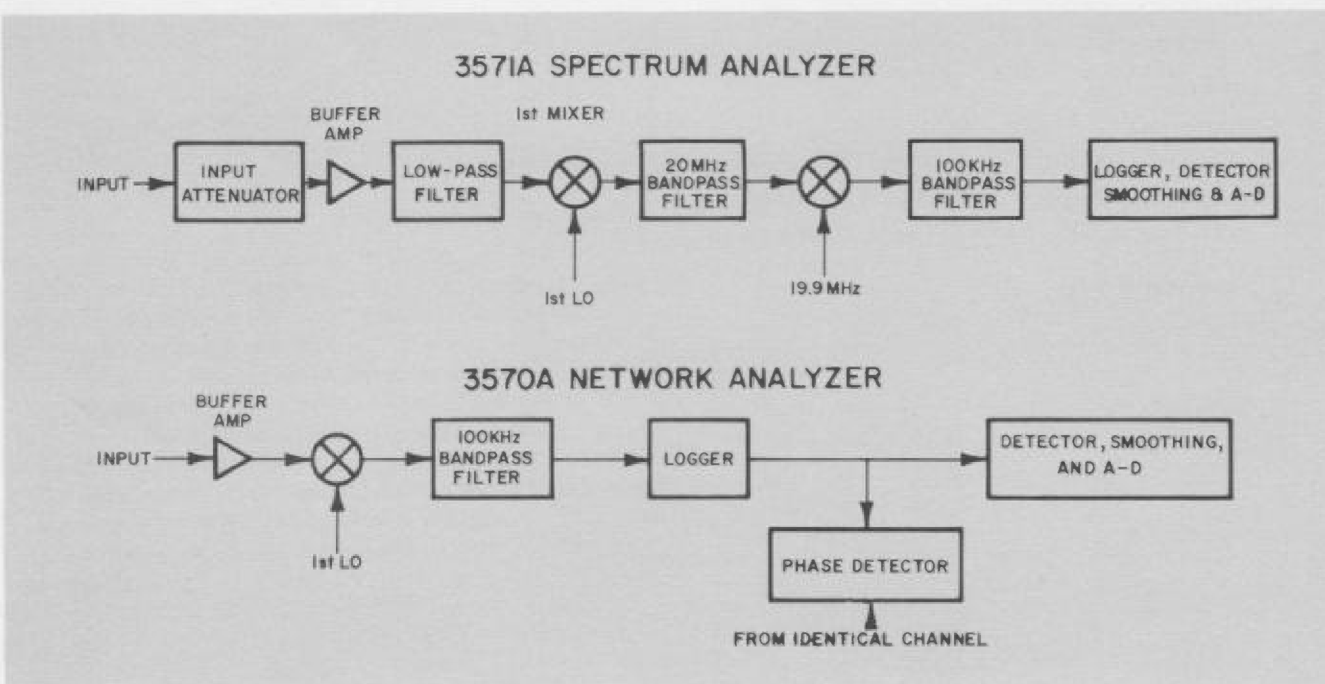


Figure 1.1. Block diagram of both analyzers.

This last point is dramatically demonstrated in Figure 1.2, where both the network and spectrum analyzers are used as a spectrum analyzer. The network analyzer introduces many spurious signals which as shown by the spectrum analyzer simply do not exist on the input signal. Thus the 3570A Network Analyzer is not, because of design trade-offs, a good spectrum analyzer.

If it isn't a good spectrum analyzer, then what is a network analyzer?

Definition: A network analyzer is a device capable of stimulating a two-port, linear network and measuring all the parameters which characterize the network. Since these parameters are, in general, a function of frequency, it must excite the network with all frequencies in the region of interest and give all parameters as a function of frequency.

For a two-port, linear network, 4 complex parameters are necessary and sufficient to completely describe the characteristics of the network. There are several equivalent ways of presenting these parameters.

For instance, some people use,

$$A, B, C, D$$

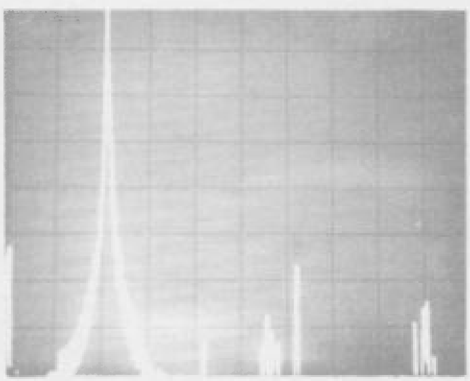
and others transform to

$$Z_{11}, Z_{12}, Z_{21}, Z_{22}$$

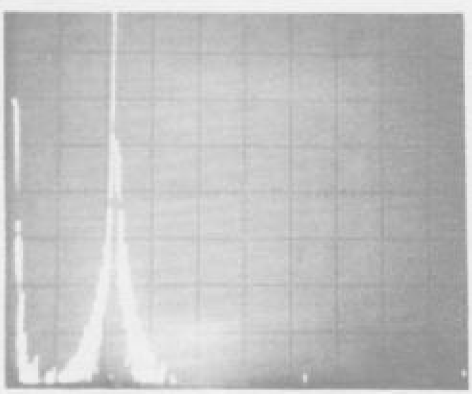
or to

$$S_{11}, S_{12}, S_{21}, S_{22}.$$

This list is by no means exhaustive, but each group of four parameters is sufficient to describe the network.



3570A Network Analyzer used as a spectrum analyzer.



Identical test span, bandwidth and sweep rate using 3571A Spectrum Analyzer.

Figure 1.2.

A spectrum analyzer with a tracking generator is capable of stimulating the network, but only measuring the magnitude of the 4 parameters, *not the phase*. Thus, it can only partially characterize the network. When the 3571A Spectrum Analyzer is used with either a 3320 or 3330 frequency synthesizer to furnish the local oscillator signals, the front panel output is a tracking generator signal. So, while the 3570A Network Analyzer cannot be used as a spectrum analyzer, the 3571A Spectrum Analyzer can be used as a limited network analyzer.

It should also be pointed out that while the 3570A Network Analyzer can measure the magnitude and phase of transfer functions, it requires external hardware in the form of hybrids, directional couplers or attenuators to measure the other two parameters of a network.

Chapter 2
CONTINUOUS SWEEPS

Because network and spectrum analyzers haven't previously had frequency synthesizers as the local oscillator, these analyzers have traditionally swept the frequency domain in a continuous manner. A frequency synthesizer, because it can change to any frequency in a time independent of the step size, need not be swept in this manner. (See the next chapter.) However, a synthesized analyzer, with its stepped sweep, can approximate a continuous sweep by making *the frequency step smaller than the bandwidth*.

This is desirable for two reasons. First, this means that a synthesized analyzer can produce traditional continuous analyzer results. Secondly, since a continuous sweep doesn't miss any frequencies, no unexpected anomalies in the input signal will be missed.

What potential problems could be encountered in a continuous sweep? The first is the trade-off between how fast one can sweep versus the accuracy one requires. This problem occurs in traditional analyzers, also. Two potential problems unique to synthesized analysis are phase glitching and local oscillator sidebands. All these problems are covered in the following sections.

Amplitude Compression

One of the problems encountered when sweeping in a continuous manner is amplitude compression and skewing due to too high a sweep rate. This is illustrated in Figure 2.1. This multiple-exposure photograph shows the reduction in amplitude and displacement to the right that occurs with increased sweep speed. This phenomena is a consequence of the response time of the spectrum analyzer's internal filter. The magnitude of these effects is theoretically derived in Appendix A and the results are presented in graphical form in Figure 2.2. These graphs present
· different

Sweep rate is defined as:

$$R = \frac{\Delta f}{\Delta t}$$

Where Δf = frequency step of synthesizer
and Δt = time/step of synthesizer

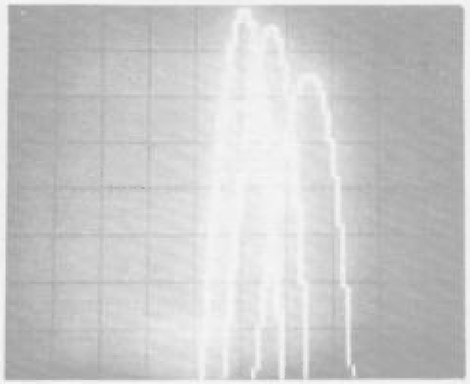


Figure 2.1. Effects of sweep rate upon the output of the 3571A Spectrum Analyzer.
Sweep rates: .1 BW^2, .83 BW^2, 2 BW^2

The sweep rate is normalized on the graphs in terms of bandwidth squared. Let us illustrate the use of these graphs by way of an example, using the 3571A Spectrum Analyzer. Suppose because we would like to resolve very close signals or because we are trying to measure very low level signals, we select a 10 Hz bandwidth. Further assume that we are not interested in very high accuracy and so we will sacrifice it for a higher sweep rate. Let us say we are willing to accept 0.4 dB amplitude compression. Also, at least at first, we will have the smoothing off. To determine the sweep rate, we first select the 10 Hz bandwidth graph. We follow the 0.4 dB error line across to the smoothing off curve and read below that the sweep rate is 0.825 BW^2 or 82.5.

Since,

$$\frac{\Delta f}{\Delta t} = 82.5,$$

then

$$\Delta t = \frac{1}{82.5} \cong 12 \text{ msec}$$

As explained in Appendix C, because of the communication between the synthesizer and the analyzer, the 10 msec/step setting actually takes 12 msec/step. So we select the 10 msec/step setting on the 3330 Synthesizer. (Otherwise we would have to select the 30 msec/step position to ensure that the amplitude compression was less than 0.4 dB.) These are the conditions used in the second trace of Figure 2.1. As one can see, the trace has about 0.4 dB compression as was predicted.

For the 3570A Network Analyzer, only the "smoothing off" curve on Graph A is needed for all bandwidths if the analyzer and not the device-under-test (DUT) is determining the bandwidth (see Appendix A). If the DUT determines the bandwidth, then the curve "no VBW" on Graph A should be used, using the bandwidth of the DUT in the calculation of sweep speed. Again, Appendix A covers some considerations involved in selecting the sweep speed, including the effect of sweep rate on phase.

If the 3571A Spectrum Analyzer is used for amplitude-only network analysis, the same considerations apply to it as to the 3570A Network Analyzer, except if the analvzer is determining the bandwidth. Then the curves are used as if they were in the spectrum analyzer case.

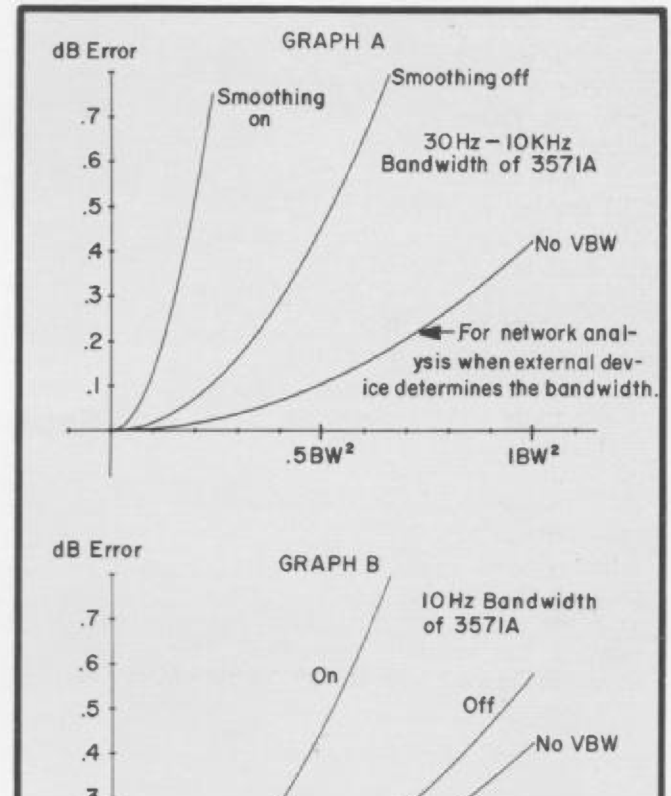


Figure 2.2. Sweep rate versus amplitude compression.

Digital Sweeping Concerns

Because of the block diagram of both the 3320 and 3330 Frequency Synthesizers, at certain cardinal frequencies some of the internal phase-lock loops range to their opposite frequency extreme. Under this condition the individual phase detectors can momentarily fall out of lock and the output of the synthesizer can phase jump as it sweeps through these frequencies. This momentary out-of-lock condition is allowed for in the frequency settling time specifications of the instrument, but it does mean that the new output can be a somewhat arbitrary number of cycles out of phase from what would be expected. Phase jumps like this cause networks and filters to "ring" the same as amplitude jumps. This can cause the minor annoyance of small "blips" on a CRT display. This occurs at 100 kHz and its harmonics when using the 3330. When a calculator is controlling the analyzers, it should wait a little bit longer before taking a reading after passing through these points (say five time constants of the network or filter involved if one is already sweeping fast).

Of more crucial concern, if only for the reason of the more serious consequences to those ignorant of this problem, are the sidebands that can develop on the synthesized local oscillator. As shown in Appendix B, the step-step sweep of a synthesizer is equivalent to a continuously sweeping oscillator with sidebands. When viewed from the analyzer output (and one assumes that the local oscillator is clean), it appears as if the signal being analyzed has sidebands (or a network has spurious responses). This condition is shown in Figure 2.3 for the 3571A Spectrum Analyzer.

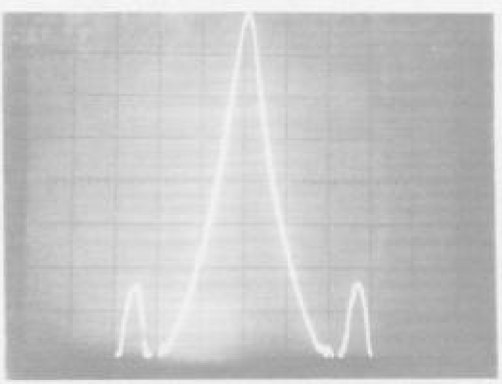


Figure 2.3. Apparent signal sidebands caused by sidebands on the local oscillator of the 3571A Spectrum Analyzer.

30 Hz bandwidth
3 msec time/step
1.4 Hz step
1000 point sweep

Equations in Appendix B predict -64 dB sidebands 333 Hz away from carrier.

Appendix B derives one fact which always demonstrates whether suspicious responses are in fact caused by the synthesizer. It is that the frequency spacing of the sidebands is equal to the inverse of the time/step of the synthesizer. *Thus spurious sidebands or responses will change apparent frequency when the time/step is changed.* Of course, true input spurious responses won't change frequency, so we have an easy test.

3571A Recommended Operating Points

Appendix B also derives the conditions necessary to keep these spurious responses below -80 dB, the dynamic range of the 3571A. These conditions are combined with the previously discussed conditions on sweep rate in Appendix C to generate a table of recommended operating points. This table is reproduced in Figure 2.4.

As noted in Appendix C, this is a rather conservative combination of 0.1 dB amplitude compression *and* 80 dB spurious response. If both are not required *at the same time*, faster sweep rates are possible.

Also see Appendix B for the arguments as to why the spurious sidebands will not generally be a problem in network analysis. Hence it is sufficient to just use the amplitude compression curves to determine operating points for either the 3570A or 3571A when used as a network analyzer.

Table C-I
RECOMMENDED OPERATING POINTS

Bandwidth (Hz)	Smoothing	Δf (Hz)	Δt (msec)
3	both	0.4	3000
10	off	0.1	3
	on	1.5	1000
30	both	0.2	1
100	both	.4	1
		↓	↓
		47.1	300
300	both	4.2	1
		↓	↓
		141.3	100
1k	both	47.1	1
		↓	↓
		502.4	30
3k	both	423.9	1
		706.5	3
		1695.6	10
10k	both	4.71k	1
		7.85k	3

Figure 2.4. 3571A recommended operating points.

Chapter 3
CHANNEL SWEEPS

Continuous sweeps reveal in wonderful detail all the characteristics of a spectrum and can upon occasion show the circuit designer a few surprises! But in applications like production testing and quality assurance, this wealth of data is at best not needed and at worst a waste of time to gather and process even in automated testing. What is needed is to check a few judiciously chosen frequencies to verify in a minimum amount of time the performance of a device or system. Such a test we shall call a channel sweep.

Channel sweep is a term for a frequency sweep where the frequency step is much larger than the bandwidth. This occurs in spectral analysis when one wants to measure pilot tones on a communications channel or the fm sideband levels of a transmitter modulated by a test tone. In network analysis, this occurs when one wishes to check the performance of a device at a few key points, i.e., desired 3 dB points, insertion loss at center frequency, or attenuation at crucial frequencies. All these examples meet the criterion that a great deal is already known about how the spectra *ought* to appear, and if certain points don't meet our expectations, then something is wrong.

How fast can we do a channel sweep? Our carefully derived $BW^2/2$ criterion for continuous sweeps is no longer applicable. This is because instead of gradually approaching a signal (or response) and letting the filter begin to respond slowly due to its smooth skirts, we hit it suddenly with a signal. The signal was previously so far away as to be greatly attenuated, and so of insignificant effect upon the output of the filter. This is another way of saying that we are concerned with the response of the filter to a tone burst.

Tone Burst Response Times

The response of the 3571A filters to a tone burst is shown in Figure 3.1.

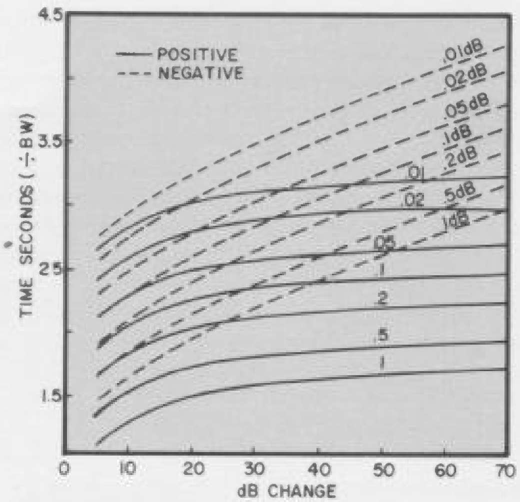
Several characteristics of these curves should be noted. The response time, of course, goes up as one increases the amplitude change or requires a more accurate reading. Therefore, one can sweep faster if lower accuracy is required or if one knows the test points will all be approximately the same amplitude. This occurs in the previously mentioned example of checking pilot tones in a communication channel or when checking the response of a wideband device.

Another characteristic to note is that there is a minimum response time, which is the time delay of the filter. This is another limitation on how fast we can channel sweep.

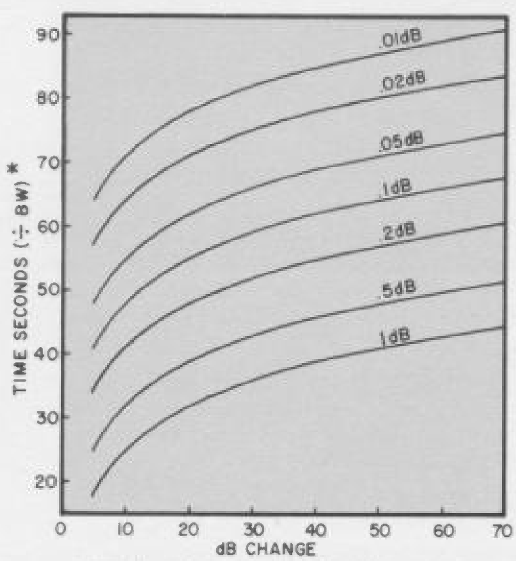
Lastly, with the smoothing off, there is a difference in response time between increasing and decreasing the amplitude (positive and negative dB changes respectively). This occurs because we are looking in a logarithmic manner at the output of an energy storage device, the filter. The energy in a tuned circuit builds up much faster (in a logarithmic sense) than it decays, hence the difference in the curves. In the general sweeping case, one must take the worst case, that for a negative amplitude change. With the smoothing on, the response time of the video filter is so much longer than these effects that they are not seen in the response and only one set of curves is needed.

The 3570A Network Analyzer doesn't have the same set of curves, but rather the table shown in Figure 3.2. The network analyzer is subject to the same kind of constraints and would have similar curves to the spectrum analyzer with the smoothing off. These curves can be inferred from the response time table.

In a network analysis, it is important to remember once again, that the external network can limit the response time. This can happen particularly with a high "Q" network. It always happens when their bandwidth is less than that of the analyzer.



Response time (smoothing off).



* Divide by 30 for 30Hz-3Hz BW.

Response time (smoothing on)

Figure 3.1. 3571A response time curves.

Typical Settling Time (following 40 dB step):

Bandwidth	90% Settled	100 % Settled
10 Hz	200 ms	800 ms
100Hz	20 ms	80 ms
3kHz	1 ms	4 ms

Figure 3.2. 3570 response time.

Chapter 4
POTENTIAL AMPLITUDE AND PHASE MEASUREMENT PROBLEMS

Now that we have studied the potential sources of error that occur when making swept measurements, let us now consider another class of potential problems—those which affect both single frequency and swept measurements. Two problems of particular significance are caused by noise and the lack of ideal grounding. These pitfalls of the real world create noise jitter and what is called the classical attenuator problem, respectively. The following sections explain these problems and present solutions to them. In the final section of this chapter the reader is reminded of the effects of return loss.

The Classical Attenuator Problem

The classical attenuator problem is another effect which has always been present and yet has not caused substantial problems until the introduction of high accuracy, large dynamic range instruments like the 3042A and 3045A systems. It is, very simply, due to the finite impedance of the shields of the cables used to connect the signals to the analyzers. Thus its effects can be reduced by using short cables with good ground continuity (e.g., double-shielded cables), but never eliminated. It may manifest itself in several ways:

- Apparent detector inaccuracies at low signal levels;
- Reduction of dynamic range at low frequencies;
- Spurious responses caused by common-mode signals.

The basics of the problem are outlined in Figure 4.1. This represents a signal source driving an attenuator that is monitored by a detector. Coaxial cables are used, and the test is being conducted at frequencies in the audio range (at high frequencies, the coaxial cables behave more like baluns and the problem is not so acute).

To simplify the discussion, the attenuator is set for infinite attenuation. It is easily seen that return currents through the cable shield to the signal source can develop a voltage, e_a, across the finite resistance of the cable shield, R_{C1}. This causes voltages across R_{C2} and Z.

If the detector input resistance, R_D, is high, the voltage drop across R_{C2} is seen by the detector so a residual signal is measured even with infinite attenuation.

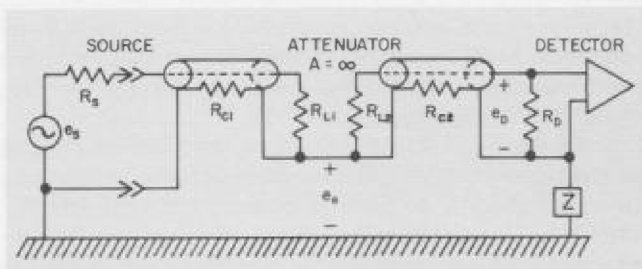
When $(R_{L2} + R_D) \gg R_{C2}$,

$$e_a = \frac{R_{C1}(R_{C2} + Z)}{(R_s + R_{L1})(R_{C1} + R_{C2} + Z) + R_{C1}(R_{C2} + Z)} e_s$$

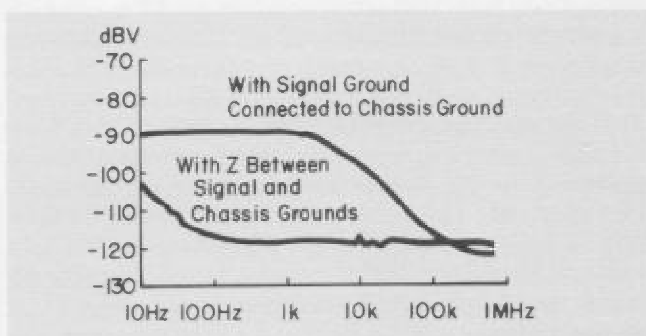
and,

$$e_D = \frac{R_{C2}}{R_{C2} + Z} e_a$$

The object is to reduce the detector signal, e_D, to zero, or at least to insignificant proportions. This would occur if either R_{C1} or R_{C2} were zero, but this would be difficult to achieve. Increasing $(R_s + R_{L1})$ and/or Z would also reduce e_D but R_s and R_{L1} are fixed by the measuring system, which leaves Z as the only variable available for manipulation.



Equivalent circuit



Improved performance with isolated ground.

Figure 4.1. Classical attenuator problem.

In the 3570A Network Analyzer (a predecessor to the 3571) the "Z" of Figure 4.1 is near zero, i.e., the input is not isolated at low frequencies as in the 3571A. Therefore, for this instrument, the full voltage of e_a is measured by the detector. Consider a 50Ω system using two 4-foot lengths of RG/58U cable with an infinite attenuator between. What would be the real attenuation?

For the cable used $R_{C1} = R_{C2} \cong 20\ m\Omega$ and $R_s + R_{L1} = 100\Omega$, so

$$\frac{e_D}{e_s} \approx \frac{(20 \times 10^{-3})(20 \times 10^{-3})}{100 \quad (40 \times 10^{-3})} \rightarrow -80\ dB$$

This reduction of the measurement range is why there is the 600Ω line on the sensitivity graph of the 3042 data sheets and manual. If the source is terminated in greater than 600Ω, then

$$R_s + R_{L1} > 650\Omega$$

and

$$\frac{e_d}{e_s} < -100\ dB,$$

which is below the range of the instrument.

Another way to reduce the classical attenuator problem is to make Z non-zero. Increasing Z to just 1Ω and leaving the source terminated in 50Ω yields,

$$\frac{e_D}{e_s} \approx \frac{(20 \times 10^{-3})(20 \times 10^{-3})}{100 \quad (1.0004)} \rightarrow -108\ dB$$

Thus, a small increase in Z results in a significant reduction in e_D. A similar analysis shows that common mode voltages caused by ground loops are also reduced by increasing Z.

Increasing Z, however, would not allow the barrel of the front-panel BNC connector, which is connected to signal ground, to be tied directly to earth ground. The classical attenuator error was reduced in the Model 3571A without fully floating the input connector by making Z a saturable-core inductor wound with #17 wire. This has practically zero impedance at dc but a finite impedance at frequencies where the classical attenuator problem exists. On the other hand, a large powerline signal through Z, such as might occur with a grounding error, would saturate the core, reducing the impedance of Z to much less than one ohm. This is why the input to the Model 3571A is not described as "floating," but as "isolated at low frequencies."

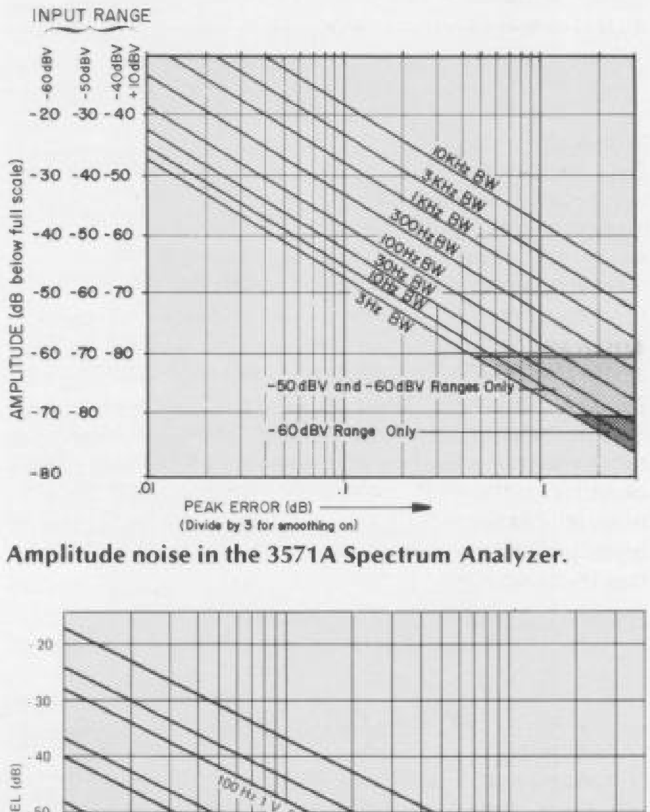
The potential reduction in measurement errors achieved by this arrangement is shown by the graph in Figure 4.1. This was made by the model 3571A measuring the output of a 120-dB attenuator fed by a one-volt signal supplied through a 4-foot cable (bandwidth: 3 Hz range: 1 mV; smoothing:on).

Noise Jitter

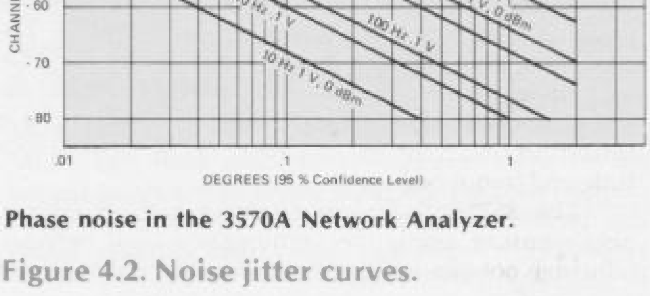
As has been stated before, extreme accuracy is possible with the use of digital instead of analog displays. But to obtain this accuracy one must be aware of noise. Noise is a ubiquitous problem, whose effects have been traditionally reduced in network and spectrum analysis by the use of video filters and "eyeball" averaging by the operator. It was previously pointed out that the 3570A and 3571A analyzers both have video filtering at all times, but if the analyzers are used in a calculator-controlled system, using the operator to estimate a noise reading by "eyeballing" is extremely inefficient. What is needed is to be able to estimate the probable error in a reading due to noise. This is what the noise curves in the instrument manuals and reproduced in Figure 4.2 can do.

To begin our discussion of these curves, it should be noted that since noise is a statistical phenomena, we can only *estimate* the noise. In this case we can estimate to a 95% probability that the error will be lower than predicted. In other words, out of 100 readings taken, about 95 will have an error due to noise of less than the predicted amount. The other five or so readings will have noise errors in excess of the predicted amount, but the chance of an excessively noisy reading falls off very rapidly with increasing error, because the noise is Gaussian in distribution. Thus the chance of a reading with an extremely large error (compared to the predicted error) is very remote. These arguments are developed in detail in almost any book on noise or statistics and so will not be further expounded here.

We note from the curves that the predicted noise error is lower if the signal to be measured is near the top of the input range, if the bandwidth is reduced or if, in the case of the 3571A, the smoothing is "on." All these things improve the signal to noise ratio, and hence reduce the noise error in a reading. Note also that these arguments apply to both amplitude and phase readings, as noise is present in both cases.



Amplitude noise in the 3571A Spectrum Analyzer.



Phase noise in the 3570A Network Analyzer.

Figure 4.2. Noise jitter curves.

Calculator Signal Averaging

If reducing the bandwidth and input range and turning the smoothing on still fail to bring the predicted noise error within acceptable bounds, due either to other constraints on these variables or the need for extremely high accuracy, there is yet another option for the user, signal averaging. Signal averaging is the calculator equivalent of the operator "eyeballing" an average reading. The calculator takes several readings at the same frequency and averages them together. The noise voltage is reduced by the square root of the number of readings taken, *if the samples are independent.* To insure this, the samples must not be taken too rapidly; that is, no faster than 1/(2VBW), where VBW is the video bandwidth of the analyzer in Hertz. (The video bandwidth of the analyzers for selected bandwidths are shown in table B-1.)

One good way to insure that samples are not taken too rapidly is to use the 3330 Frequency Synthesizer which acts as the local oscillator to trigger the readings. If the 3330 is stepped in frequency or amplitude it triggers a reading of the 3570A or 3571A analyzer after waiting the time/step programmed into the 3330. So, if the 3330 is told to take a frequency or amplitude step of zero (whichever is more convenient) with the desired sample time in the time/step, the analyzer will return a reading after the correct time.

Return Loss

Since much has been written on the subject of return loss, especially in the microwave field, it will only be stated in passing that imperfect terminations can reduce the accuracy of low frequency measurements as well. Even over the 13 MHz range of these analyzers, return loss can cause unflatness in the response significant enough to reduce the accuracy of a reading. This is why the accuracy enhancement software of the 3042A Network Analyzer System with the 9825A Calculator is only specified for networks with greater than 26 dB return loss. It should also be noted that reflecting waves caused by imperfect terminations can produce errors in phase readings because of the shift in the zero crossing point. Mathematical derivation of these effects can be found in texts on microwave theory and practice under the transmission line sections.

Chapter 5
AMPLITUDE SWEEPS

The 3570A and 3571A Network and Spectrum Analyzers are not only useful for amplitude vs. frequency sweeps, but they can add another dimension to these measurements by amplitude sweeping as well. The high amplitude accuracy and resolution of the 3330B (or 3320B) used as the network stimulus coupled with the accuracy and resolution of the analyzers, allows one to make precise, programmable amplitude sweeps. The network analyzer system, for instance, can study amplitude compression and phase distortion of a network as a function of input amplitude and frequency.

The 3571A Spectrum Analyzer can, of course, also measure amplitude compression of a network (although not phase distortion) using the local oscillator synthesizer as the network stimulus. In addition, with extra 3320B or 3330B synthesizers, one can make high accuracy harmonic or intermodulation distortion measurements and generate outputs of distortion versus input amplitude and/or frequency.

Amplitude Sweep Speed

How fast can one amplitude sweep? Certainly one constraint is how fast the analyzer responds to changes in its input. This was covered in detail in the chapter on channel sweeps and the reader is referred to that chapter.

Another limit on amplitude sweep rate is how fast the source settles to the new amplitude. The 3330B and 3320B have identical output sections and so the following discussion applies to both sources. The settling time for either instrument is 350 msec. with the leveling in the slow condition and 35 msec. with the leveling fast or off. The leveling is a thermopile-controlled leveling loop which improves the accuracy of the source's output voltage. Because the leveling loop could distort the output at low frequencies, it is necessary to turn it off below 10 Hz and to the "slow" condition when below 1 kHz. An idiosyncrasy of the leveling loop which should be mentioned in passing occurs when one first turns on the "slow" condition. Due to internal time constants of the loop, there is no output for about 7 seconds. After this initial stabilization, one can change the leveling as desired and not encounter this 7 second delay until the power is turned off and on again.

The leveling loop improves the accuracy of the source and yet the 350 msec. per step settling time necessary for frequencies between 10 Hz and 1 kHz may be unacceptably long. One possible remedy is to cleverly exploit the capabilities of the system as a whole. Figure 5.1 shows a block diagram of output section of the synthesizer. Note that the leveling loop itself only covers a 10 dB range and that the attenuator covers the remainder of the range. *Thus, if the output is changed by 10 dB, the leveling loop need not settle to a new value and the output will settle within a few milliseconds.* So the fastest way to make a multi-decade amplitude sweep is to step through the desired range with 10 dB increments and then start over, offset by the desired amplitude step size (see Figure 5.2). Continuing in this manner until the sweep is complete, we have taken the same number of measurements, but in the fastest possible way. Now we use the calculator to sort these measurements out, process and present them in the desired format (e.g., plotting). This is an excellent example of the improvements that can come with intelligent integration of the system components to overcome the individual equipment limitations and make the integrated total performance greater than the sum of its parts.

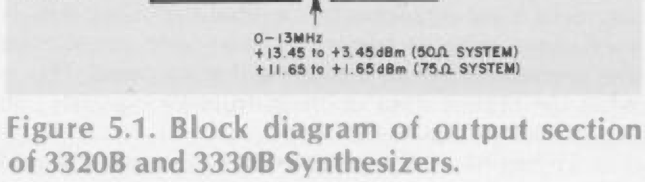


Figure 5.1. Block diagram of output section of 3320B and 3330B Synthesizers.

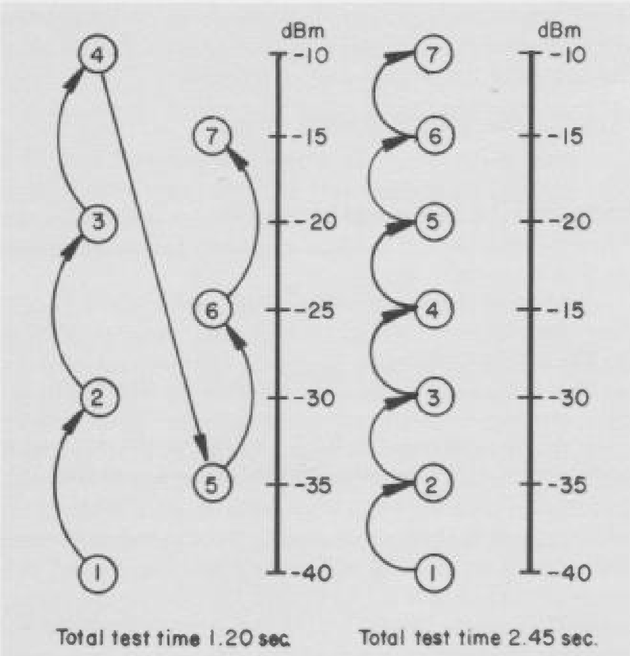


Figure 5.2. Example of fast sweep routine when the leveling is "slow." (Sweep from -40 dBm to -10 dBm in 5 dB steps at a frequency of 100 Hz.)

Chapter 6
CONCLUSION

As can be seen from the size of this somewhat voluminous note, there are many things one must consider to realize the *potential* accuracy of the 3570A and 3571A analyzers. Obviously, not all the potential problems discussed will be significant in any one measurement situation. This note should serve, however, as a handy reference in any measurement situation with these analyzers. Lastly, the reader is reminded that this note is not all inclusive, and other application notes should be consulted about particular measurement areas or problems.

Appendix A
MAXIMUM SWEEP RATE DUE TO FILTER RESPONSE TIME

This appendix contains the mathematical derivation of the classical formulas for maximum sweep rate of Gaussian filters. More importantly, it shows when these are and aren't applicable to spectrum and network analysis.

It can be shown that a Gaussian filter is the optimal filter for swept, single-frequency measurements, in the sense that it can be swept the fastest and has no overshoot in the response. Unfortunately, it is impossible to build with a finite number of realizable components. However, a multiple-stage synchronously-tuned filter is a fair approximation to a Gaussian filter and has the added advantage that it is easy to change its bandwidth. Because of these reasons, synchronous filters are used in almost all spectrum and tuned network analyzers, including the 3570A and 3571A.

Because the mathematics is much easier, the Gaussian filter is treated in the following development rather than the synchronous filter. Also, the case of a low-pass filter instead of a bandpass filter is used. The equivalence of these problems is shown in most texts on network analysis. The following analysis develops the formulas for the decrease in peak response and phase error due to sweep rate in analyzers with Gaussian filters followed by a logarithmic detector and low-pass ("video") filters shown in Figure A-1. The last simplifying assumption we shall make is that the frequency sweep is of infinite extent so that we can use Fourier transforms. This assumption is equivalent to ignoring the transient effects of the starting of a sweep. This sweep is illustrated in Figure A-2.

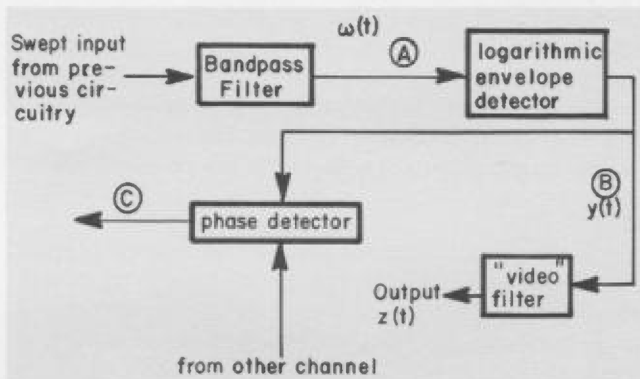


Fig. A-1. Block diagram.

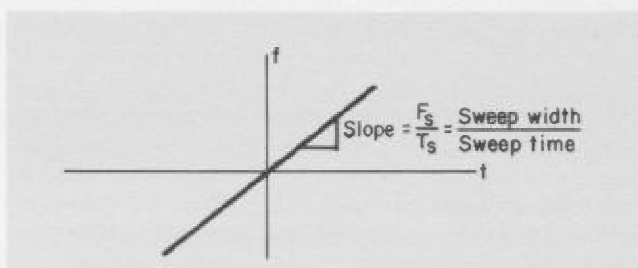


Figure A-2. Infinite, linear sweep

The Fourier transform of this sweep is,

$$S(\omega) = \tau\sqrt{2\pi}\exp\left[-\frac{1}{2}(\tau\omega)^2\right]$$

$$\text{where } \tau = \sqrt{\frac{jTs}{2\pi Fs}} \qquad \text{(A.1)}$$

The Fourier transform of a Gaussian filter is,

$$H(\omega) = \exp\left[-\frac{1}{2}(\delta\omega)^2\right] \qquad \text{(A.2)}$$

where $\delta = \sqrt{\ln 2}/\pi BW$ and BW = 3 dB Bandwidth

The time output of the Gaussian filter (point A in Figure A.1) is the inverse transform of the product of $S(\omega)$ and $H(\omega)$. This is,

$$\omega(t) = \frac{\tau}{\sqrt{\tau^2+\delta^2}} \quad \exp\left[-\frac{1}{2}\,\frac{t^2}{\tau^2+\delta^2}\right] \qquad \text{(A.3)}$$

Substituting back for τ and δ and expressing in polar notation gives,

$$\omega(t) = \frac{1}{\left[1 + \left(\frac{2\ln 2}{BW^2}\frac{Fs}{Ts}\right)^2\right]^{1/4}} \exp\left[\frac{\frac{-\pi^2 BW^2}{2\ln 2} t^2}{1 + \left(\frac{\pi BW^2}{2\ln 2}\frac{Ts}{Fs}\right)^2}\right]$$

$$\exp j\left[\frac{\frac{\pi^3 BW^4}{4(\ln 2)^2}\frac{Ts}{Fs} t^2}{1 + \left(\frac{\pi BW^2}{2\ln 2}\frac{Ts}{Fs}\right)^2} + \tan^{-1}\left(\frac{\ln 2}{\pi BW^2}\frac{Fs}{Ts}\right)\right] \tag{A.4}$$

Since the phase is determined by a zero-crossing detector type of circuit, the logger doesn't affect the phase information. Also, since the phase information is never used in the 3571, let us look only at the magnitude as it passes through the logger. Therefore, the output of the logger (point B in Figure A-1) is,

$$y(t) = -\frac{1}{4}\log\left[1 + \left(\frac{2\ln 2}{\pi BW^2}\cdot\frac{Fs}{Ts}\right)^2\right] - \frac{1}{2}\log(e)\frac{\frac{\pi^2 BW^2}{2\ln 2}t^2}{1 + \left(\frac{\pi BW^2 Ts}{2\ln 2\, Fs}\right)^2} \tag{A.5}$$

To get the analyzer output, z(t), we will convolute the time response y(t) with the impulse response of a single pole low-pass filter of radian bandwidth B (the "video" filter). The result of that convolution is,

$$z(t) = \frac{-1}{4}\log\left[1 + \left(\frac{2\ln 2}{\pi BW^2}\frac{Fs}{Ts}\right)^2\right]$$

$$-\frac{\log e}{4\ln 2}\bullet\frac{\pi^2 BW^2}{1 + \left(\frac{\pi BW^2}{2\ln 2}\frac{Ts}{Fs}\right)^2}\left[t^2 - \frac{2}{B}t + \frac{2}{B^2}\right] \tag{A.6}$$

The parabolic terms in time show that with video filtering the peak output does not occur at t=0, but somewhat later at time t=1/B. This amounts to an error in estimating the frequency of a signal when sweeping of

$$F_{error} = \frac{Fs}{Ts}\bullet\frac{1}{B} = \frac{Fs}{2\pi VBW\, Ts} \tag{A.7}$$

where VBW = B/2π (video bandwidth in Hz)

The other error term which contributes to swept frequency error is the time delay of the filter. This adds directly to the time error contributed by the video filter.

The amplitude error in dB at the peak is

$$A_{error} = 20\, z\left(\frac{1}{B}\right) \quad \text{(dB)}$$

or

$$A_{error} =$$

$$-5\log\left[1 + \left(\frac{2\ln 2}{\pi BW^2}\frac{Fs}{Ts}\right)^2\right] - \frac{5\log e\, BW^2}{4\ln 2\, VBW^2\left[1 + \left(\frac{\pi BW^2}{2\ln 2}\frac{Ts}{Fs}\right)^2\right]} \tag{A.8}$$

Since both the 3570A and 3571A are high accuracy instruments with .01 dB resolution, let us consider only small amplitude errors.

Then,

$$A_{error} \cong -\frac{20(\ln 2)^2}{\pi^2\ln 10}\left(1 + \frac{\log e\ln 10}{4\ln 2}\frac{BW^2}{VBW^2}\right)\frac{1}{BW^4}\left(\frac{Fs}{Ts}\right)^2 \tag{A.9}$$

For the special case of the video bandwidth much greater than the Gaussian bandwidth we find that the sweep rate necessary to create only .1 dB error is

$$\frac{Fs}{Ts} \cong \sqrt{\frac{.1\ln 10}{20}}\,\frac{\pi}{\ln 2}\, BW^2 \cong \frac{BW^2}{2} \tag{A.10}$$

This $BW^2/2$ criterion is the standard rule of thumb used in spectrum analysis work.

The video bandwidths used in the 3571A with the smoothing on and off are given in Table A-I. These are used in conjunction with equation A.6 to generate the error curves in Figure 2.2. These curves tell at a glance the sweep rate which must not be exceeded if a desired accuracy is to be maintained.

It should also be noted that the combination of the A-D converter in the analyzers and the calculator can act as a digital filter if the appropriate algorithms are employed and this can further improve the smoothing of the results. The user is reminded that the digital filter's effect on the output error must also be included in equation A.6.

	Approximate Video Bandwidth	
IF Bandwidth	**Smoothing Off**	**On**
10 kHz	5 kHz	1.7 kHz
3 kHz	1.5 kHz	500 Hz
1 kHz	500 Hz	170 Hz
300 Hz	150 Hz	50 Hz
100 Hz	50 Hz	17 Hz
30 Hz	15 Hz	5 Hz
10 Hz	15 Hz	5 Hz
3 Hz	15 Hz	5 Hz

Table A-I. Video Bandwidths of the 3571A Spectrum Analyzer.

Network Analyzer

Now that we have the results for spectrum analysis, we shall tackle the more complicated case of network analysis.

To begin with, the phase error at the output of the Gaussian filter (point A in Figure A-1) is (according to equation A.4),

$$\measuredangle\ \omega(t) = \frac{\frac{\pi^3 BW^4}{4(\ln 2)^2}\frac{Ts}{Fs}t^2}{1+(\frac{\pi BW^2}{2\ln 2}\frac{Ts}{Fs})^2} + \tan^{-1}(\frac{\ln 2}{\pi BW^2}\frac{Fs}{Ts}) \quad (A.11)$$

Once again, because of the high accuracy and resolution of the 3570A we are only interested in small phase errors due to sweeping and thus we can approximate equation A.11 with,

$$\measuredangle\ \omega(t) \cong \pi\frac{Fs}{Ts}t^2 + \frac{\ln 2}{\pi BW^2}\frac{Fs}{Ts} \quad (A.12)$$

Now the phase reading of the 3570A is the difference of the phase of the two channels. It is important to realize at this point, however, that both channels don't necessarily have the same bandwidth. The bandwidth of a channel can be determined either by the internal bandwidth (selected by the front panel switch) or by the bandwidth of the device-under-test, *whichever is smaller.*

Therefore, the output of the phase detector (point C in Figure A-1) is

$$\phi_{error} = \omega_B(t) - \omega_A(t)$$

$$\phi_{error} \cong (\frac{1}{BW_B^2} - \frac{1}{BW_A^2})\frac{\ln 2}{\pi}\frac{Fs}{Ts} \quad \text{(radians)} \quad (A.13)$$

Note that ideally *there is no phase error* when the bandwidth of both channels is the same. This occurs when the bandwidth of the device-under-test is much larger than the bandwidth of the analyzer itself. This zero phase error happens because the errors of the two identical channels cancel. If one is making a "B-A" amplitude measurement under these same conditions, one also gets ideally *zero amplitude error.* However, if one is just making a single channel measurement (A or B), or the device-under-test bandwidth is less than the analyzer bandwidth, then the amplitude error cancellation doesn't occur and equation A.6 still applies. The bandwidth used in this equation is once again the analyzer or device bandwidth, whichever is smaller.

It should be pointed out that equation A.13 is not nearly as good an approximation to the phase error as equation A.9 is to the amplitude error and that the phase error will always be greater than predicted. This is because the synchronous filter is a much better approximation to a Gaussian filter in the amplitude sense than in the phase sense, especially if the number of poles in the synchronous filter is small. Also, if the external device is determining the bandwidth, the phase and amplitude errors can be very much larger than predicted. This is because many common filters such as Chebyshev, Butterworth and elliptic filters have much worse transient response than a synchronously-tuned filter. Thus it may be necessary to sweep much slower than predicted. In general, it is impossible to predict the necessary sweep speed because it involves a convolution which must be evaluated numerically. However, the above equations do serve as a basic guide to determining sweep rate.

Appendix B
SIDEBANDS CREATED BY SYNTHESIZER SWEEPS

A sweep by a synthesizer looks as shown in Figure B-1. As one can see, the difference in frequency between the desired linear sweep and the synthesizer sweep is a sawtooth-shaped error function. This step-step approximation to a linear sweep generates sidebands on the synthesizer which can appear as spurious signals on a network or spectrum analyzer. This appendix will derive the conditions necessary to keep these sidebands below any desired level.

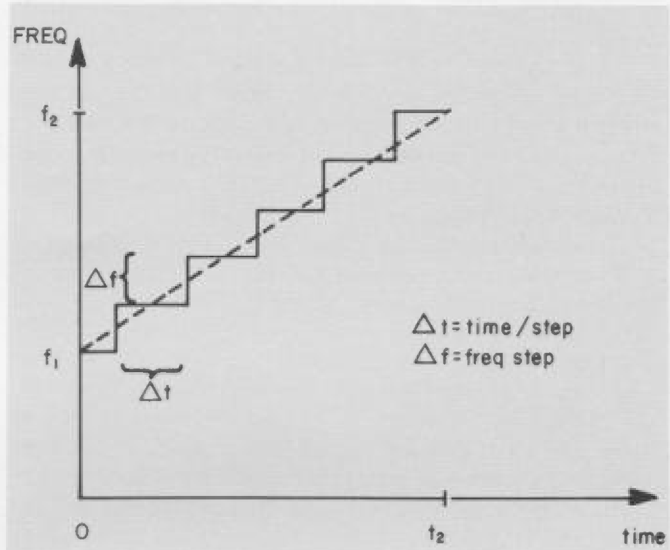


Figure B-1. Frequency sweep by a synthesizer.

The sawtooth frequency error function can be expressed as a Fourier Series (see any table of Fourier Series) as follows:

$$f_{err} = \frac{\Delta f}{\pi}\sum_{n=1}^{\infty}\frac{(-1)^{n+1}}{n}\sin\omega_n t \quad (B.1)$$

where $\omega_n = 2\pi n / \Delta t$

Δf = frequency step (Hz)

Δt = time/step

However, we are interested in the phase error function and we integrate the frequency error function to get:

$$\phi_{err} = \frac{\Delta f \Delta t}{\pi}\sum_{n=1}^{\infty}\frac{(-1)^n}{n^2}\cos\omega_n t \quad (B.2)$$

because $\phi = \int 2\pi f\, dt$

So the swept synthesizer output is as follows:

$$LO = \sin(\omega_{LO}t + \phi_{err}) \quad (B.3)$$

where $\omega_{LO} = 2\pi\frac{\Delta f}{\Delta t}t + 2\pi f_1$

and f_1 = start frequency

Using the trig sum of angles formula and realizing that we are only interested in small phase errors, we get approximately,

$$LO \cong \sin \omega_{LO} t + \phi_{err} \cos \omega_{LO} t \qquad (B.4)$$

Expressing ϕ_{err} as the Fourier series we derived above and using yet another trig identity, we get,

$$LO \cong \sin \omega_{LO} t + \frac{\Delta f \Delta t}{2\pi} \sum_{n=1}^{\infty} \frac{(-1)^n}{n^2} [\cos (\omega_{LO} - \omega_n) t + \cos (\omega_{LO} + \omega_n) t] \qquad (B.5)$$

This equation is in the form of a signal at frequency ω_{LO} with sidebands at all harmonics of $2\pi/\Delta t$ falling off at a $1/n^2$ rate. Also note that the magnitude of the sidebands can be reduced by decreasing either Δf or Δt.

Let us next determine the frequency step and time/step necessary to reduce these sidebands to the specification levels for the 3571A Spectrum Analyzer. The reader may desire to hold the sidebands to some other level, and may modify the following development to suit his own needs.

The arguments for the 3570A Network Analyzer are not presented, because unless the external device under test has extremely steep skirts, the sidebands will not show up. This is illustrated in Figure B-2. In the two examples we can see how an extremely steep network shows the sideband response, but the more gentle network buries the sideband response in the response to the main signal. We can also see how selecting a narrower analyzer bandwidth would further reduce the sideband response in both cases. In any case, the level of the sidebands themselves, whether significant or not, is still determined by equation B.5.

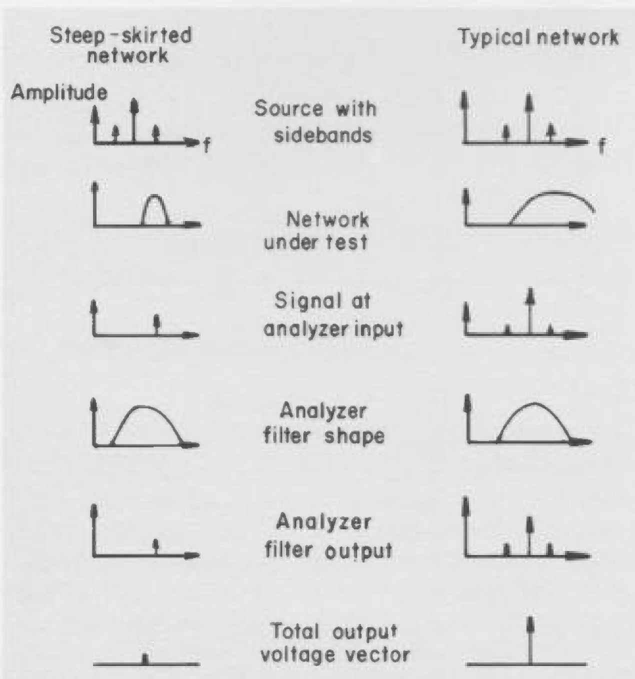
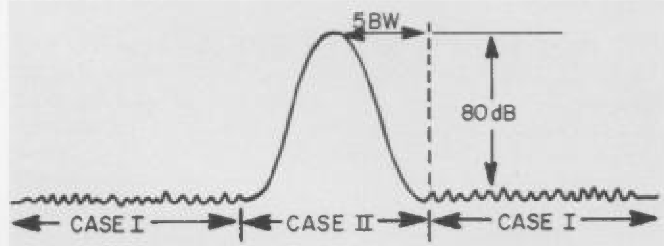


Figure B-2. Effect of oscillator sidebands on network analysis.

To determine what conditions are necessary to bury these sidebands within the other responses of the instrument, we must look at the three cases shown in Figure B-3.



Case III is the frequency step greater than the bandwidth and is covered in the chapter on channel sweeps.

Figure B-3. Typical filter response curve.

Case I is when the first sideband frequency is beyond the 80 dB response of the filter. Here, all the sidebands must be below 80 dB, the level of other spurious responses in the instrument.

Case II occurs when the first sideband is within the 80 dB bandwidth of the filter. Here we want all the sidebands to be below the filter response curve out to the 80 dB point. If this is the case, they will all be hidden by the filter response.

Case III is when the frequency step is greater than the bandwidth. Here, the approximations used break down and we consider this case in the chapter on channel sweeps.

To determine the break point between Case I and Case II, we will consider the 80 dB point of the filter to be 5 bandwidths from the center. This is based on a rough average of the shape factors of all the filter bandwidths. Therefore, Case I occurs when

$$\frac{1}{\Delta t} \geq 5 \text{ BW or } \Delta t \text{ BW} \leq 1/5 \qquad (B.6)$$

For this case, we see from equation B.5 that,

$$\frac{\Delta f \Delta t}{2\pi} < 10^{-4} \qquad (80 \text{ dB}) \qquad (B.7)$$

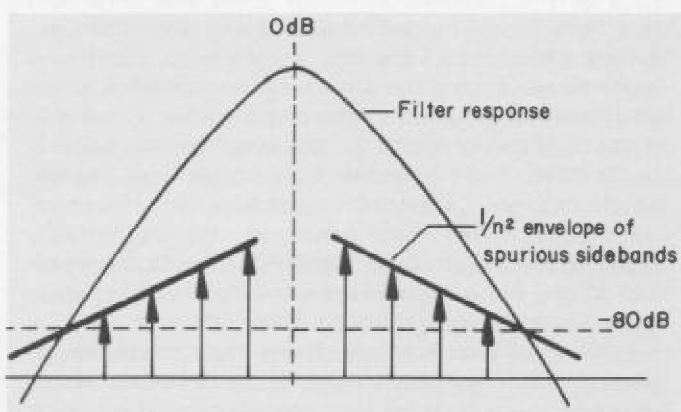


Figure B-4. Case II — first sideband frequency within the 80 dB bandwidth.

Case II occurs when

$$\Delta t BW \geq 1/5$$

and we can determine the conditions on Δf and Δt from figure B-4.

We note from Figure B-4 that the filter response falls off much quicker than the l/n^2 envelope of the sidebands. Therefore, it is sufficient to ensure that the envelope crosses the filter response at the -80 dB point. To establish this, we need

$$\frac{\Delta f \Delta t}{n^2 \bullet 2\pi} < 10^{-4} \quad \text{(from equation B.5)} \qquad \text{(B.8)}$$

when

$$\frac{1}{\Delta t} \bullet n = 5\ BW.$$

Eliminating n from the two equations and simplifying gives,

$$\frac{\Delta f}{\Delta t} < \frac{\pi}{2} \times 10^{-2} BW^2 \qquad \text{(B.9)}$$

Appendix C combines all of these derived conditions and graphically comes up with some recommended operating conditions.

Appendix C
DERIVATION OF RECOMMENDED OPERATING POINTS

The following derivation is presented in detail because the reader may wish to use his own operating criterion and derive his own operating points. The operating points developed here are derived graphically from the equations and criterion derived in appendices A and B.

Because the graph from which the operating points are taken has many lines on it which may tend to confuse the user, it will be developed from a series of graphs. These graphs each portray a different limit on the selection of an operating point. Since the object is to select the frequency step and time/step, the axes are the normalized variables of Δf/BW and ΔtBW.

Figure C-1 shows the limits on Δf and Δt which are simply based on the smallest steps the 3330 synthesizer can make (.1 Hz and 3 msec/step). The time/step is equal to the programmed time/step plus 2 msec for the exchange of data between the source and the analyzer. If a calculator is connected to the system, the time/step can be considerably longer than 2 msec extra/step, depending on the speed of the calculator and the efficiency of the software used. But, in any case, the line segments in Figure C-1 depict the minimum Δf and Δt that can be used and other valid operating points are to the right of the line segments.

The results of Appendix A were used to generate the curves in Figure C-2. These lines are limits which must not be exceeded if the peak amplitude is not to be in error by more than 0.1 dB. Note that, as was shown in Appendix A, one must sweep slower with the smoothing control on.

Figure C-3 contains the final set of limits on the selection of Δf and Δt, the sideband level limitations derived in Appendix B. These come from equations B.7 and B.9.

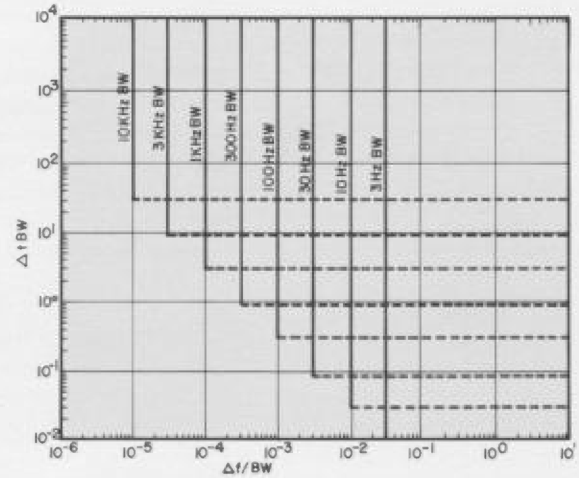


Figure C-1. Instrument limits.

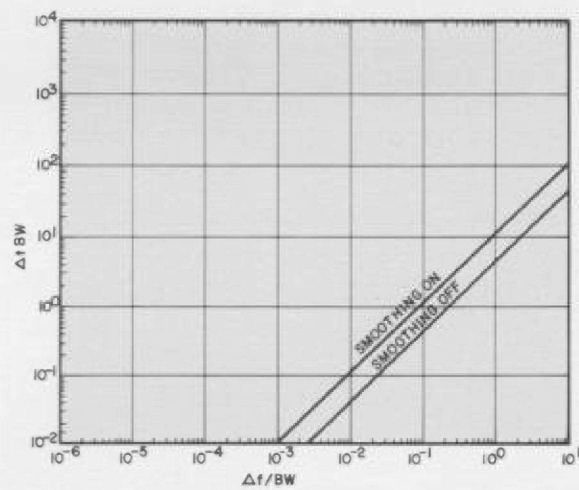


Figure C-2. .1 dB peak compression

These constraints are all combined to produce Figure C-4, the combined limitations for selecting Δf and Δt. From it we can select an operating point for each bandwidth which allows one to sweep at the fastest rate possible and yet not violate any of the above limitations. To sweep as fast as possible, we need to use as large a Δf as we can, while simultaneously using as small a Δt as possible. This is equivalent to saying our desired gradient points to -45° (see Figure C-4). Using this criterion, several operating points are shown in Figure C-4. Table C-I has a complete list of recommended operating points.

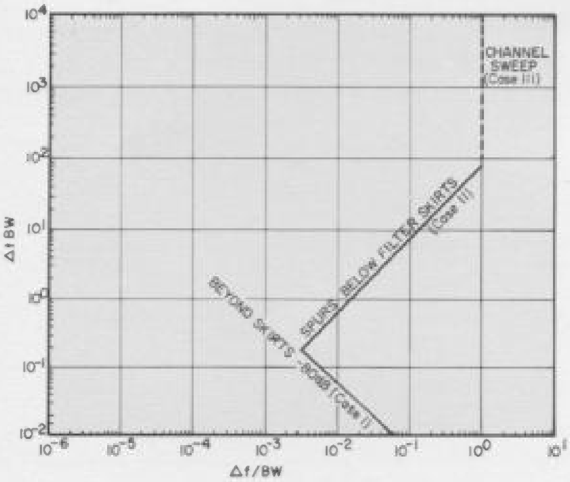


Figure C-3. Sideband level limitations.

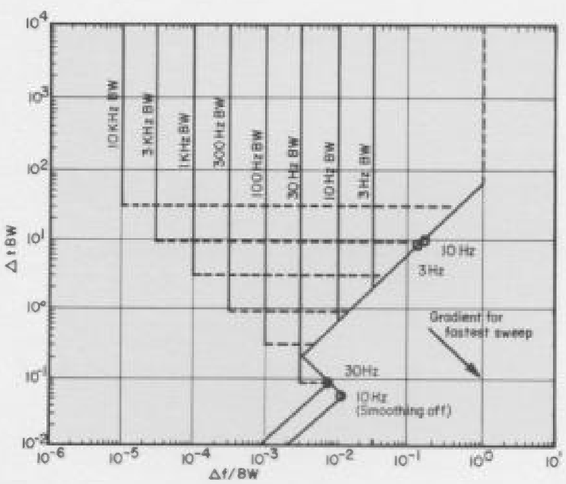


Figure C-4. Selection of recommended operating points.

To illustrate how the operating points are selected, let us look at the 10 Hz bandwidth case. With the smoothing on, only points above the Case II line are within all the limits. Therefore, only the 100 msec, 300 msec, 1 and 3 second time/step positions are valid. Since we only have 0.1 Hertz resolution on the frequency step size, at 100 msec/step we can only step .1 Hz without violating the limits and this is not as fast as we can sweep. So, let us try 300 msec/step. Here we could use 0.4 Hz/steps, which is sweeping much faster (further in the direction of the gradient), but it turns out that with 1 sec/step and 1.5 Hz/step we sweep the fastest of all without violating any of our limits. Thus, this was selected as the operating point. With the smoothing off, a valid operating point just happens to fit in the corner of the smoothing off curve and the Case I curve. This is the fastest sweep rate available for this bandwidth, again because it is the farthest in the direction of the gradient.

In the widest bandwidth it should be noted that the recommended operating points do not have the frequency step exceeding the bandwidth as the continuous sweep approximation would break down.

Lastly, it should be noted that this rather conservative combination of limits makes for fairly slow sweeps. If the user doesn't need 0.1 accuracy and 80 dB spurious performance *simultaneously*, then he can redraw the curves in this appendix and select operating points which yield a faster sweep rate.

Table C-I
RECOMMENDED OPERATING POINTS

Bandwidth (Hz)	**Smoothing**	**Δf** (Hz)	**Δt** (msec)
3	both	0.4	3000
10	off	0.1	3
	on	1.5	1000
30	both	0.2	1
100	both	.4	1
		↓	↓
		47.1	300
300	both	4.2	1
		↓	↓
		141.3	100
1k	both	47.1	1
		↓	↓
		502.4	30
3k	both	423.9	1
		706.5	3
		1695.6	10
10k	both	4.71k	1
		7.85k	3

ANNOTATED BIBLIOGRAPHY

The following Application Notes detail other common problems in spectrum analysis and are available from HP.

150 Spectrum Analysis . . . Spectrum Analyzer Basics

Describes the theory and operation of spectrum analyzers and their application. Includes information on the fundamentals, harmonic mixing, pre-selection, and tracking generators. Also included is a glossary of spectrum analyzer terms.

150-1 Spectrum Analysis . . . AM and FM

A discussion of measurement of AM and FM spectra using a spectrum analyzer. The theory of basic modulation is given, and several examples acquaint the user with basic techniques.

150-2 Spectrum Analysis . . . Pulsed RF

Extends the discussion of AM to the special case of pulsed RF signal analysis. Provides techniques for the measurement of pulse parameters from the spectrum analyzer display. Includes a handy sliderule to simplify pulse calculations.

150-4 Spectrum Analysis . . . Noise Measurements

Measurement of random and impulse noise signals. Includes carrier-to-noise ratio, amplifier noise figure, white noise loading, oscillator spectral purity, and noise corrections.

150-7 Spectrum Analysis . . . Signal Enhancement

Discussed techniques of improving sensitivity of spectrum analyzer measurements including use of pre-amplifiers. Also describes using wave analyzers for detailed analysis of the video output from the spectrum analyzer.

Also see the latest Application Note Index for new notes in areas of interest.

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