Plotting Frequency Response with the Hybrid Computer

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PLOTTING FREQUENCY RESPONSE
WITH THE HYBRID COMPUTER

BY

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B.S.E.E., Louisiana State University, 1964

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ABSTRACT

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WITH THE HYBRID COMPUTER

BY

DAVID KENNETH SWARTWOOD

This paper describes a hybrid computer microprogram which plots frequency domain responses for linear systems. The microprogram computes the real and imaginary parts of the system output and displays either Bode or Nyquist plots.

Various approaches are discussed and a detailed explanation of the one selected is presented. The major areas of discussion are the sinewave generator, the computation of real and imaginary parts of the system output, the logarithm computation and the digital control logic.

The conclusion gives a comparison of a Nyquist plot made with the microprogram with one calculated by a digital computer. Possible improvements for the microprogram are also discussed.
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INTRODUCTION

Historically, the principal use of analog computation has been for time domain analysis. This paper describes a hybrid computer microprogram which plots the frequency domain responses for linear systems. The goal of the study was to find a microprogram which would display both Nyquist and Bode plots.

Frequency domain responses are often used in system studies and it may be desirable to be able to perform frequency domain analysis directly on the hybrid computer rather than transforming the transient response. Descriptions of systems in the frequency domain often arise because of the ease in finding a frequency response in the laboratory and because of the accuracy with which this data describes a network or system. Frequency domain descriptions are also frequently used when the signals are periodic such as in power or communication systems. They are valuable when noise is present and the frequency content of the noise is known.

The program presented in this paper computes the real and the imaginary parts of the system output and displays, either Bode or Nyquist plots. The Bode plot is not complete because only the amplitude vs frequency plot is presented and a Bode plot must have both an amplitude and a phase plot. Since the real and imaginary parts of the output were calculated, the phase shift could have been
computed by taking the inverse tangent of their ratio. This was not done because trigometric function generators were not available on the computer that was used and it was too complex to do with standard analog components.
CHAPTER I

CONSIDERATION OF APPROACHES

The frequency response method differs from the time domain approach in that only the steady state response of the system to a fixed sinusoidal input is of interest, while in time domain analysis the transient response is the characteristic used to define the system. To generate the frequency response a sinusoidal input must be varied in frequency over the range of interest and the steady state output measured for each frequency input. This means that for each frequency change the transient must decay before the output measurement can be made. The delay while waiting for the transient to decay is a major obstacle to performing high speed frequency response sweeps on the analog computer.

Several techniques were considered to minimize this transient delay. For example, it is possible to apply initial conditions to the system integrators and eliminate the transients. These initial conditions would be calculated between changes in frequency. An additional microprogram would be required to perform the calculation and it would be unique for each system model. The need for this extra microprogram was the primary reason this approach was not used. This same objection eliminated the approach of generating the transient separately and subtracting it from the circuit output so that only the
steady state response would be present. In the method selected, the transient is monitored and when it exceeds a selectable value the frequency sweep is held until the transient decays to the acceptable value.

The selected approach requires no special modeling. The system model required is the same as that used for transient analysis. Frequency response curves for a new system can be made by simply changing the system model. It should be noted that the microprogram that is described in this paper could be considered a prototype for a test instrument that would plot frequency domain responses for electronic circuits. This demonstrates how the hybrid computer can be used for bread-boarding during design and development of electronics.
CHAPTER II

DESCRIPTION OF SIMULATION

The selected approach is summarized in figures 1 and 2. The flow diagram in figure 1 shows the sequence which is followed to sweep the input frequency over four decades. The sequence is started by selecting the initial system time constants and the initial frequency. The first frequency sweep is then made over one decade of frequency. If an excessive transient is sensed during this sweep, the frequency is held until the transient decays to an acceptable value. At the end of the decade of frequency sweep, the system time constants are rescaled for the next decade and the frequency reset to the initial value. This process continues until four decades are swept. At this time the system time constants are initialized and the process repeated.

A simplified block diagram of the system is shown in figure 2. As indicated in the diagram, the hybrid computer has both analog computing modules and logic elements available. The logic devices extend the capability of the analog computer by making it possible to perform high speed logic operations which respond to and in turn control the analog computation.

The analog/hybrid computer used for this simulation was an Applied Dynamics Model AD-5 and the labeling of the devices and their
Fig. 1. Flow diagram for frequency sweeps
Fig. 2. Simplified block diagram
operation are compatible with this machine. However, the program uses only standard hybrid components and can easily be converted for another computer. Throughout this paper, the analog signals are scaled making the maximum analog signal unity.

The sinewave input for the system model is derived from a voltage controlled oscillator, VCO, which is driven from a ramp generator. These analog devices are controlled by the gates and flip flops in the control logic section. The phase and amplitude of the system output is processed by the two multipliers. The real and imaginary parts of the output are found by multiplying the output by both \( \sin \omega t \) and \( \cos \omega t \) and then storing the result with the track and store units. The input and output of one of the track and store units is measured to detect transients and the resulting signal is monitored by the control logic. The stored signals are used directly for the Nyquist plot and the square root of the sum of their squares is computed for the Bode plot. Log generators are used to find the value of frequency and magnitude in decibels for the Bode plot. These circuits are described in detail in the following sections. A schematic of the complete microprogram is shown in figure 10.
CHAPTER III

SINEWAVE GENERATOR

The sinewave stimuli for the system model is generated by the VCO shown in figure 3. It is driven by the ramp generator which ranges between scaled values of one tenth to one resulting in one decade of frequency adjustment.

A quadrature oscillator was used for the VCO. It was selected because it generates both sine and cosine outputs, gives nearly instantaneous changes in frequency and because of the continuous amplitude control it affords. The principle of the quadrature oscillator is to implement an analog loop which solves the differential equation:

\[ \ddot{x} + \omega_0^2 x = 0. \]

The undying transient solution is:

\[ x = A \sin \omega_0 t, \quad x(0) \Delta 0 \]

The loop that performs this function is formed by A05, A06 and A08. Frequency control is obtained by the use of multipliers N04 and N14. Because multipliers are used, changes in frequency can be accomplished as quickly as the slew rates of the multipliers will allow. It should be noted that no delay exists due to the integrator time constants.
Fig. 3. Sinewave generator
When the frequency is instantly changed, the integrator outputs are the same immediately before and after the change. Therefore, the VCO starts oscillating at the new frequency with the same sinusoidal position that existed before the change.

The amplitude control is unlike many techniques which are used to keep the sinewave from diverging or converging in that it is continuous. Techniques are often used which integrate the peak or average value of the sinewave to control the amplitude. Because of the integration there is a delay of several sinewave cycles in the control. The control technique used with this VCO is based on the trigonometric identity:

\[ A^2 \sin^2 \omega t + A^2 \cos^2 \omega t = A^2. \]

Since both the sine and the cosine are generated, they can be squared and summed. If the square of the amplitude is subtracted from this sum the result will be zero whenever the identity is satisfied. This identity is implemented with N22, N23, A19, A26, A27, N06 and N16 and the resulting error signal is fed back to correct the error. Because the identity is valid for all time and because it is implemented with an algebraic loop, the correction is continuous and does not allow errors to build up.

The sinewave generator is analyzed using figure 4. The circuit elements shown as circles are multipliers which are used to provide control over the amplifier gains. The gain through the multipliers is indicated by the terms \( a_1, a_2, b_1 \) and \( b_2 \).
Fig. 4. Sinewave oscillator

Summing the inputs to amplifier A05 yields the equation:

$$
\ddot{x} + \frac{b_2}{a_2} \dot{x} = -\frac{b_1}{a_2} \dot{x} - \frac{b_1 b_2}{a_2} x - a_1 x
$$

$$
\ddot{x} + (b_1 + b_2) \dot{x} + (a_1 a_2 + b_1 b_2)x = 0.
$$

By comparison to the expression $s^2 + 2\delta\omega_n s + \omega_n^2$ where $\delta$ = damping factor and $\omega_n$ = natural frequency of oscillation, it can be determined that the term $b_1 + b_2$ controls the damping. Note that when $b_1 = b_2 = 0$ the equation reduces to $\ddot{x} + a_1 a_2 x = 0$ which is the differential equation that describes a continuous sinusoidal oscillation which neither increases or decreases in magnitude. Therefore, $b_1$ and $b_2$ can be controlled by feedback to keep the oscillation amplitude constant.
The comparison with $s^2 + 2\omega_n s + \omega_n^2$ indicates $a_1 a_2 = \omega_n^2$ where $\omega_n$ is the natural frequency of oscillation for the circuit. Thus the frequency of the circuit can be controlled by multipliers $a_1$ and $a_2$.

If control is provided about only one of the integrators, the equations reduce to:

$$\ddot{x} + b_1 \dot{x} + a_1 x = 0 \quad \text{for } a_2 = 1, \quad b_2 = 0.$$ 

This indicates that frequency and amplitude control can be obtained by control about a single integrator. The principle reason that this was not done and that a separate control was used about each integrator was to keep the signal levels and thus the phase shift on both as great as possible. This minimizes the error due to extraneous phase shifts in the integrator with the least phase shift. An additional reason is that it provides equal amplitude signals for the amplitude control circuit. This is necessary to satisfy the trigonometric identity:

$$A_1^2 \sin^2 \omega t + A_2^2 \cos^2 \omega t = A^2$$

that is, $A_1 = A_2 = A$.

The oscillator frequency is varied over one decade. The maximum frequency can be found by writing the differential equations for the circuit in figure 4 with $a_1$ set equal to $a_2$ and with $b_1$ and $b_2$ equal to zero.
\[ -\ddot{x} = a^2 x \]
\[ \ddot{x} + a^2 x = 0 \]

Therefore,

\[ \omega_n = a = \text{Natural frequency of oscillation in radians/sec} \]

\[ \omega_n = 2\pi f_n = 1000 \text{ radians/sec (maximum available on computer)} \]

\[ f_n = \frac{1000}{2\pi} \text{ when "times one" integrator inputs are used} \]

\[ = 159.15 \text{ Hz} \]

\[ f_n = \frac{10,000}{2\pi} \text{ when "times ten" integrator inputs are used} \]

\[ = 1591.5 \text{ Hz} \]

The use of the times ten input on the integrators is optional. The selection is based on the range of gains in the system model and the location of breakpoints in the frequency domain.
CHAPTER IV

FINDING REAL AND IMAGINARY PARTS
OF SYSTEM OUTPUT

The input to the system model is defined as $A_1 \sin \omega t$. The resulting system output from a linear system is $A_0 \sin (\omega t + \phi)$ where the magnitude has been changed from $A_1$ to $A_0$ and $\phi$ is the phase shift from input to output. This holds true for any linear system. Non-linear systems generate frequencies other than the applied frequency and this plotting technique will not work. The phasor output is resolved into real and imaginary parts as indicated by the polar to rectangular formula:

$$A_o e^{j(\omega t + \phi)} = A_0 \cos (\omega t + \phi) + jA_0 \sin (\omega t + \phi)$$

Since the real and imaginary parts of the phasor output are required to make the Nyquist plot, the output is operated upon by multipliers N24 and N25. This multiplication by sine and cosine results in one term which is related to the phase shift. The expressions for $V_{\sin}$ and $V_{\cos}$ in figure 5 are defined below.

$$V_{\sin} = A_0 \sin (\omega t + \phi) \sin \omega t$$

$$= (A_0 \cos \phi \sin \omega t + A_0 \sin \phi \cos \omega t) (\sin \omega t)$$

$$= A_0 \cos \phi \sin^2 \omega t + A_0 \sin \phi \cos \omega t \sin \omega t$$

15
\[ A_0 \cos \phi \sin^2 \omega t + A_0 \sin \phi \frac{\sin 2 \omega t}{2} = A_0 \cos \phi \sin^2 \omega t + \frac{A_0 \sin \phi}{2} \sin 2 \omega t \]

where \( A_0 \cos \phi = \text{real component of } A_0 e^{j(\omega t+\phi)} \) at \( \omega t = 2K\pi, K = 0, 1, 2 \cdots \)

\[ V_{\cos} = A_0 \sin (\omega t + \phi) \cos \omega t \]
\[ = (A_0 \sin \phi \cos \omega t + A_0 \cos \phi \sin \omega t) \cos \omega t \]
\[ = A_0 \sin \phi \cos^2 \omega t + A_0 \cos \phi \sin \omega t \cos \omega t \]
\[ = A_0 \sin \phi \cos^2 \omega t + \frac{A_0 \cos \phi}{2} \sin 2 \omega t \]

where \( A_0 \sin \phi = \text{imaginary component of } A_0 e^{j(\omega t+\phi)} \) at \( \omega t = 2K\pi, K = 0, 1, 2 \cdots \)

Fig. 5. Multiplication by sine and cosine

The coefficient of the left hand term is \( A_0 \cos \phi \) for multiplication by \( \sin \omega t \) and \( A_0 \sin \phi \) for multiplication by \( \cos \omega t \). Since these coefficients are the real and imaginary parts of the output which are desired, they must be separated from the \( \sin 2\omega t \) term. This is done by sampling the multiplier outputs when the sine and the cosine waves are zero. The sine wave generates the sample for the imaginary term and the cosine wave generates the sample for the real term. The \( \sin 2\omega t \) term is eliminated because at the sample times this
term is zero. It was found that an acceptable sample pulse width was three microseconds since the maximum real frequency generated by the VCO is 1.59 KHz.

The sample waveforms are shown in figure 6. The values for the real and imaginary parts are sampled and stored by integrators A10 and A20 which are operated as track and store units. The stored value is updated each cycle and provides a continuous output for the display.
Fig. 6. Sample pulses for multiplier outputs
CHAPTER V

FINDING LOGARITHM OF MAGNITUDE AND FREQUENCY

The Bode magnitude plot is a plot of magnitude vs frequency with both expressed in decibels. The magnitude of the gain is calculated from the real and imaginary values which are stored on the track and store amplifiers. The square root of the sum of the squares of the real and the imaginary terms is calculated by N02, N03, N12, A07 and A17. Amplifier A07 provides an offset adjustment which is essential if small values of output are to be plotted. Amplifier A13 is configured as a log amplifier so that the gain can be plotted in decibels. It uses the free diodes that are available on the general purpose amplifier so no external components are necessary. This approach was taken because no logarithm modules were available on the computer that was used. This simple log amplifier is sensitive to temperature changes and is not accurate over many decades but it was sufficient for the visual displays required in this problem. The amplifier provides a logarithmic output over two and a half decades when used in the magnitude circuit. On the output of the log amplifier, the amplifier A14 drives the Y input of the display and provides an adjustment that allows the display to be moved vertically. The logarithm of the frequency over one decade is found by the log
amplifier A03. For an input change of one tenth to one, the output varies logarithmically over a small range. Amplifier A04 multiplies this signal by ten and in addition provides an adjustment that makes it possible to move the display horizontally. Since the signal has been converted to decibels, additional decades of sweep on the display are formed by adding signals to the output of amplifier A04. The additional three decades of frequency are generated by amplifier A12. The same logic signals which change the integrator time constants in the system model are used to control the switched inputs of amplifier A12. The output of amplifier A12 drives the X input of the display through the inverter A21.

The circuit operation is based on the logarithmic relationship between the current through and the voltage across a semiconductor diode. This relationship is given by the equation:

\[ i_f = I_0 \left( \frac{v_f}{nV_T} \right) \]

where \( I_0 \) = reverse saturation current

\[ V_T = \frac{T}{11,000} \text{ volts; } T \text{ is in } ^\circ \text{K} \]

\( v_f \) = forward diode voltage

\( i_f \) = forward diode current.

If the operation region is selected so that:

\[ \frac{v_f}{nV_T} \left( \frac{1}{e} \right) \ll 1 \]
then the equation reduces to the approximate expression:

\[ i_f = I_o \left[ \frac{v_f}{e^{nV_T}} \right] \]

\[ \ln \frac{i_f}{I_o} = \frac{v_f}{nV_T} \]

\[ v_f = nV_T \ln \frac{i_f}{I_o} \]

Therefore, the diode voltage and current have a fixed logarithmic relationship when \( I_o \) and \( nV_T \) are constants. These coefficients remain constant for constant temperature.

When the diode is used in the feedback of a high gain operational amplifier, the input and output voltage of the resulting circuit are related logarithmically. The equations for the circuit in figure 7 are based on the assumption that the high gain causes the amplifier input voltage, \( e \), to be approximately zero and that the high input impedance causes the amplifier input current, \( i_a \), to be approximately zero. The gain equation is derived as shown below.

\[ \frac{e_i}{R} = i_f \]

\[ e_o = nV_T \ln \frac{i_f}{I_o} \]

\[ e_o = nV_T \ln \frac{e_i}{I_o R} \]
The analog devices are controlled by the binary logic as discussed in chapter II. The modes of the ramp generator integrator are controlled by gates G20 and G27. The integrators on the AD-5 computer have hold, HD, and operate, OP, inputs which control the integrator's modes of operation. The HD input is dominant placing the integrator in the hold mode whenever it is a logic one. When both inputs are logic zero, the integrator is in the initial condition, IC, mode.

Operation of the circuit is initiated by depressing the latching operate button on the master control console. This causes the operate signal to switch from logic zero to logic one and starts the frequency sweep. When the operate signal is reset, G24 provides a logic one to the inputs of G20 and G27 placing integrator A00 in the IC mode. In the IC mode, the integrator output is controlled by potentiometer P02. It is adjusted for an initial value of one tenth. The application of the operate signal places a logic one on the inputs of G20 and G27. If the other inputs to these gates are also logic zero the integrator is switched to the operate mode and the ramp from one tenth to one is initiated.

The remaining inputs to G20 and G27 provide additional control of the ramp. One of these controls places the integrator in the hold
mode if an excessive transient is detected on the system output. The VCO frequency remains fixed as long as the integrator is in the hold mode and this allows the transient to decay to an acceptable value. This detection is accomplished by measuring the peak amplitude of the output waveform on two successive cycles and subtracting these values. If a steady state has been achieved by the output, the two successive cycles will be equal. Track and store amplifier A10 stores the peak output value for display as described in chapter IV. This stored value is also used to determine when the transient is within acceptable limits. Amplifier A22 performs this function by adding the output of A10. Since A10 is an inverting amplifier, this is equivalent to subtracting the two successive output values. The difference between these two values is monitored by the two comparators CO1 and CO2. These comparators are enabled at the peak value of the waveform so that the difference is only measured at this time. Potentiometers P01 and P02 are set so that the comparators generate a logic one if the difference exceeds a selectable limit. The comparator outputs are applied to gate G20 so that when either becomes a logic one G20 and G27 place integrator A00 in the hold mode.

The ramp generator is reset to one tenth when the ramp reaches a normalized value of one as shown in figure 8. The output is monitored by comparator CO0 which is only enabled every 100 microseconds to prevent false or double triggering. This comparator sets FF22 when the ramp exceeds a value of one. The flip flop is also enabled every 100 microseconds. This guarantees that the flip flop will be set for at least 100 microseconds which is sufficient time to
Fig. 8. Ramp generator
reset the ramp generator. When set, FF22 applies a logic one to the inputs of G20 and G27. This places integrator A00 in the IC mode and resets the ramp.

Flip-flop FF22 also provides a differentiated output for the four state counter which is composed of FF20 and FF21. This output is one microsecond wide and one pulse occurs at the end of each ramp. The binary counter increments one count for each differentiated pulse. The counter is used to control the time constants for the system model and the frequency sweep.

Sample pulses for the control of the two track and store devices are generated by flip flops FF23 and FF41. The track and store units hold the real and imaginary values of the system output for display or plot as required. The timing diagram for FF23 is shown in figure 9. The timing for flip flop FF41 is identical to flip flop FF23 and will not be described separately.

Comparator C03 detects when sin ωt passes through zero. The comparator output is allowed to change state when a logic one is applied to the enable input. The comparator is enabled once every three microseconds. A transition in the comparator output either from logic zero to logic one or from logic one to logic zero will result in a one microsecond pulse from gate G23. As shown in the figure, comparator C03 does not change state until it receives a clock pulse even though sin ωt passes through zero between clock pulses. When C03 output changes from logic one to logic zero, it generates a positive one microsecond pulse from the inverted input of gate G22. The pulse is terminated when FF25 changes state at the end of the enabling clock
Fig. 9. Zero crossing detector
pulse. Similarly, a positive one microsecond pulse is generated on the D output of FF25 when the CO3 output changes from logic zero to logic one. These two pulses are inputs to OR gate G23 so that either pulse causes FF23 to change states from logic one to logic zero. These pulses enable CO1 and CO2 at the peak value of the imaginary part of the output. Because the FF23 toggle input is a logic one until the next enabling pulse occurs, it is reset to the initial condition. This generates a three microsecond pulse at the output of FF23. This pulse and the corresponding one from FF41 are the sample pulses for the track and store units.
Fig. 10. Microprogram for making frequency domain plots
The goal of this project was to develop a microprogram for use on the hybrid computer which would display frequency domain curves in both Bode and Nyquist plots. This was achieved with the exception of the phase plot which is part of the Bode plot. This could have been accomplished with trigonometric function generators but they were not available on the hybrid computer used. The technique which was demonstrated in the microprogram could be used in a test instrument which would display Nyquist and Bode plots of electronic circuits.

The sinewave generator which was used provided an extremely stable and high quality frequency source. The frequency control is particularly outstanding because the frequency can be changed essentially instantaneously since it is only dependent on the slew rates of the amplifiers and multipliers. Its versatility makes it an approach that should be considered in other microprograms where performance is of particular importance. The generator operation is discussed in chapter III.

A Nyquist plot of a second order system is shown in figure 14. For comparison, a digital computer solution is shown with the microprogram output. The maximum error between the two curves is less than 3 percent and is primarily due to errors in the sampling circuits which monitor the system output. This could be improved by adjusting
Fig. 11. Nyquist plot

Transfer function

\[
G(s) = \frac{(s+2)}{(s+2) + (s+1)}
\]
potentiometers PO1 and PO2 so that the transients must achieve a smaller value before the frequency sweep is allowed to continue. It is also possible to monitor both the real and the imaginary outputs rather than only the imaginary output. This has the advantage of better sweep control when the imaginary output is near zero.

A unique feature of the program is the use of analog sampling to separate the real and imaginary parts of the output from the higher frequency terms. These higher frequencies result from the multiplication of the system output and a reference signal. They are usually separated with a low pass filter which attenuates the unwanted higher frequencies, however, a low pass filter is difficult to use when the frequency is not constant. The sampling approach is covered in chapter IV.

An area which could be improved is the dynamic range. The principal limitation was the range in amplitude which can be used without excessive errors. The amplitude range was approximately two and a half decades. This could be increased by the use of more accurate, temperature compensated logarithmic amplifiers and by the use of a hybrid computer with greater amplitude range.
LIST OF REFERENCES

