Positive-Feedback Oscillators

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Abstract—These lecture notes describe the basic principles of positive-feedback sinusoidal oscillators.

Index Terms—Barkhausen criterion, Laplace transform, natural frequencies, natural response, Nyquist stability criterion, passive network, positive-feedback linear sinewave oscillators, transient response, zero-input response.

I. INTRODUCTION

Our aim is to develop an electronic circuit that is able to produce sustained sinusoidal oscillations at a single frequency. No external ac source will drive the circuit; it should start from an initial state, and its zero-input response should approach a sinewave. We assume that in steady state, the circuit we develop can be described by a linear, lumped, time-invariant (LLTI) small-signal equivalent model. We shall call such a circuit linear sine-wave oscillator.

A. Terminology

Before we start the development of an oscillator, let us pay some attention to the terminology accepted in basic circuit theory [1]. Recall that a passive electric network is that combined of passive elements: resistors, capacitors, and inductors. A passive resistor never dissipates and a passive capacitor and inductor never store negative energy [1]. A circuit or its elements are active if they are not passive.

It is quite clear that being always stable, a passive network alone is not able to produce sustained oscillations: it has no poles in the right half of the Laplace plane, and, therefore, its natural response approaches zero.

To illustrate this, we recall that the transfer function of a network in terms of the Laplace transform [1],

\[ L\{f(t)\} = \int_{0}^{\infty} f(t)e^{-st}dt, \]  

(1)

where \( s \) is the complex frequency, \( s = \sigma + j\omega \), is defined as follows [1]:

\[ H(s) = \frac{L[\text{zero-state response}]}{L[\text{input}]} = K \frac{\prod_{j=1}^{m} (s-z_j)}{\prod_{i=1}^{n} (s-p_i)}, \]  

(2)

where \( K \) is a constant, the \( z_j \)'s are called zeros, since \( H(z_j) = 0 \), and the \( p_i \)'s are called poles, since \( H(p_i) = \infty \), provided that \( z_j \neq p_i \).

In terms of applications, oscillators, either sinusoidal or not, single or multi-frequency, linear or nonlinear, are generally used to synchronize a system (computers), to carry and detect information (communication), and to test a system (instrumentation).

C. Principles of Operation

Depending on the principle of operation, oscillators can be classified as negative resistance, parametric, relaxation, and positive-feedback. We shall treat here the latter type only.

II. POSITIVE-FEEDBACK OSCILLATORS

Positive-feedback oscillators employ an LLTI passive frequency-selective electric network (feedback network) \( \beta(j\omega) \) and an active, electronic amplifier \( A_{OL} \) interconnected in a positive-feedback loop. We shall assume here that \( A_{OL} \) in the first approach is frequency-independent.

A. Passive Feedback Network

Recall that a passive electric network is that combined of passive elements: resistors, capacitors, and inductors. A passive resistor never dissipates and a passive capacitor and inductor never store negative energy [1]. A circuit or its elements are active if they are not passive.

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constant, and series with a network branch. As far as the branch, or with an independent voltage source connected in parallel to a network structure of the circuit [1], [2]. This can be done either with an applying an excitation that does not change the natural frequencies of a network can be determined by function from an input variable to an output variable) and the in addition, on the initial state. the network topology, and the natural modes where the poles s vary.

Fig. 1. A parallel LCR network (a), locus of its natural frequencies (b), and its natural time response (c). The characteristic equation: \( s^2 + 2\alpha + \omega_0^2 = s^2 + \frac{(\alpha_0/Q)^2 + \omega_0^2}{2} \). The resonant frequency \( \omega_0 = 1/\sqrt{LC} \), L, and C are kept constant, and R varies. \( Q=\alpha/\omega_0L \). Since R is never equal infinity, the time response is always decaying. \( H(s) \) never equals infinity.

The partial-fraction extension of (2) can be written as

\[
H(s) = \sum_{i=1}^{n} \frac{K_i}{s - p_i},
\]

where the \( K_i \)'s are called the residues of the particular poles [1].

Applying the inverse Laplace transform \( \mathcal{L}^{-1}[H(s)] \) to (3)—this is usually done with the help of a look up table—one can obtain the natural response of the network,

\[
\text{natural response} = \sum_{i=1}^{n} K_i e^{p_i t},
\]

where the poles \( p_i \)'s are also called natural frequencies (or natural modes [2]) of the network. The \( p_i \)'s in (4) depend on the network topology, and the \( K_i \)'s (possibly complex) depend, in addition, on the initial state.

Recall that a network function (a particular transfer function from an input variable to an output variable) and the natural frequencies of a network can be determined by applying an excitation that does not change the natural structure of the circuit [1], [2]. This can be done either with an independent current source connected in parallel to a network branch, or with an independent voltage source connected in series with a network branch. As far as the

natural response is concerned, the above independent sources turn into an open or closed circuit, correspondingly; thus, not altering the natural structure of the network.

Consider for example a passive LCR network of Fig. 1(a). Its network function,

\[
\beta(s) = \frac{L[v_j(t)]}{L[i_d(t)]} = Z(s) = \frac{1}{s C (s - p)(s - p^*)},
\]

\[
p = -\alpha + j\sigma_d, \quad p^* = -\alpha - j\sigma_d
\]

\[
\alpha = \frac{1}{2RC}, \quad \sigma_d = \frac{1}{\sqrt{LC}}, \quad \sigma_d' = \sqrt{\sigma_d^2 - \sigma^2}
\]

can be found by connecting an independent current source \( i_d(t) \), as shown in Fig. 1(a).

Locus of the natural frequencies for the LCR network is shown in Fig. 1(b). This figure and (5) clearly demonstrate that the poles of a practical network (with a finite and positive R and, hence, \( \alpha \)) are always located to the left of the j\( \omega_0 \) axis. Or in other words, the denominator of (5) is never zero for any \( \omega \) with a nonnegative \( \sigma \). As a result, a practical LCR network cannot generate a sustained sinewave. Its output is always decaying, as shown in Fig. 1(c), and its natural response approaches zero.

One way to reach nondecaying oscillations is to compensate for the network positive (passive) resistance R (read energy
losses) with an element having negative (active) resistance \(-R\) (read energy pumping into the circuit), as shown in Fig. 2(a).

Negative-resistance oscillators are based just on this idea. They employ nonlinear electronic elements, such as the tunnel diode, having negative slope in a part of their voltage-current characteristic, to obtain—for relatively small signals—negative resistance.

Negative-resistance approach to oscillators is generally used at radio (RF) and especially at microwave frequencies, where it becomes difficult to construct a feedback circuit without introducing excess phase shift.

At relatively low frequencies, from kilohertz up to hundreds of megahertz, positive-feedback approach is generally used.

B. Feedback Loop

Another way to make the above LCR network to oscillate continuously is to integrate it in a feedback loop, as shown in Fig. 3(a), such that the total network function becomes

\[
H(s) = \frac{S_o}{S_{in}} = \frac{A_{OL}}{1 - A_{OL} \beta(s)} = K \frac{P(s)}{Q(s)},
\]

Note that contrary to (5), the denominator of (6) can reach zero for an \(s\) with a non-negative \(s\), namely, it happens when the loop gain \(A_{OL} \beta(s)\) = 1.

C. Barkhausen criterion

To reach sustained oscillations in steady state, the loop gain \(A_{OL} \beta(s)\) should be equal to unity at a physical, \(\omega\)—not complex, \(s\)—frequency, namely,

\[
A_{OL} \beta(j \omega) = 1 \Rightarrow \begin{cases} |A_{OL} \beta(j \omega)| = 1, & \text{amplitude criterion} \\
|\phi A_{OL} \beta(j \omega)| = 0, & \text{phase criterion} \end{cases}
\]

where \(A_{OL} \beta(j \omega)\) and \(\omega\) are the oscillator’s steady-state loop gain and frequency; index 1 emphasizes that in steady state the loop gain \(A_{OL} \beta(j \omega)\) is unity.

Since for a physical frequency, \(\sigma = 0\), the oscillator's steady-state poles (natural frequencies) are located directly on the \(j \omega\) axis of the \(s\) plane, as shown in Fig. 3(b), where according to (6) and (7), \(H(j \omega_0) = \infty\).

Equations (7) determine the steady-state oscillation conditions and are called the Barkhausen criterion.

D. Initial Pole Location

We now have to decide on the pole location in the initial state. Let us first suppose that there should be no difference between the initial and final pole locations; in both cases the poles can be located directly on the \(j \omega\) axis.

Magnitude of the natural response (4) in this case [see Fig. 3(b)] is a function of the initial state. It is an undesirable situation for the three following reasons. First, we do not
intend to supply the circuit with additional means to control the initial state. Second, any circuit is subject to disturbances and noise. In the case of oscillator, the uncontrolled initial state, disturbances, and noise will make its steady state unpredictable. Third, it is impossible to provide the initial pole locations exactly on the \( j\omega \) axis because of the non-zero tolerances of the circuit components.

Since it is undesirable to locate the oscillator's initial-state poles directly on the \( j\omega \) axis, they should be located in the right half of the \( s \) plane at a distance from the \( j\omega \) axis. This distance [see Fig. 3(b)] should be large enough to ensure the desirable pole location for any deviations of the circuit element values within their tolerances and temperature ranges.

The oscillator in this case will generate an increasing sinewave. When the sinewave magnitude will become sufficiently large, a \textit{negative feedback} should stabilize it by shifting the poles towards the \( j\omega \) axis, as shown in Figs. 3(b) and (c).

It is undesirable to locate the poles in the left half of the \( s \) plane because, in this case, the oscillator steady state will be \textit{unstable}.

To summarize the above: we want the oscillator to be \textit{unstable} in the initial state and \textit{conditionally stable} in the steady state. We never want it to be stable.

The simplest—but not always the best (see the concluding part of this Section)—way to determine the initial poles location in the right half of the \( s \) plane and to provide their movement to the \( j\omega \) axis, when the oscillations approach their steady-state magnitude, is to use the nonlinear properties of the amplifier \( A_{OL} \), as shown in Fig. 3(d).

In the initial state, the small-signal value of \( A_{OL} \) is simply set to be greater than it is needed to satisfy the Barkhausen criterion (7), \( A_{OL} > A_{OL1} \). As a result, the initial pole location will be either to the right or to the left of the \( j\omega \) axis [see Fig. 3(b)], depending on the specific behavior of the feedback transmission \( \beta(s) \).

For the reasons discussed above, \( \beta(s) \) for which the initial poles are in the left half of the \( s \) plane should not be employed in our design. Below (see Section III), we will show how one can easily decide whether \( A_{OL} > A_{OL1} \) shifts poles to the right or to the left of the \( j\omega \) axis.

\textbf{E. Negative Feedback}

Meanwhile, let us suppose that \( A_{OL} > A_{OL1} \) shifts the poles as desired, to the right of the \( j\omega \) axis. This forces unstable operation of the oscillator. It starts from an initial state and builds up its oscillations, as shown in Fig. 3(c). When the oscillations became large enough and reach nonlinear part of the amplifier transfer characteristic \( A_{OL} = S/R \) in Fig. 3(d), the amplifier gain \( A_{OL} \) decreases (negative feedback) and approaches \( A_{OL1} \). Accordingly, the poles move towards their steady-state location on the \( j\omega \) axis, as shown in Fig. 3(c). The steady state is reached when the amplifier gain at frequency \( \omega_0 \) equals \( A_{OL1} \), and the Barkhausen criterion (7) is satisfied.

\begin{figure}[h]
  \centering
  \includegraphics[width=\textwidth]{fig4.png}
  \caption{The effect of disturbances and noise on the oscillator output.}
  \label{fig:disturbances}
\end{figure}

\textbf{F. The effect of Disturbances and Noise}

A disturbance and noise affecting the circuit can be translated to the input of the oscillator and represented by an equivalent source \( S_{in} \), as shown in Fig. 3(a).

The source \( S_{in} \) will contribute to the oscillator output: it will either decrease or increase the oscillation caused by the initial state (the natural response).

Recall that a circuit having a single couple of \textit{purely} imaginary poles is conditionally stable. (A circuit having more than a couple of equal imaginary poles has linearly increasing natural response and, hence, is unstable.) A conditionally stable circuit has a bounded steady state only if it is not excited at the frequency of the poles (see Fig. 4). If the circuit is excited at the frequency of the poles then its forced response linearly increases with time.

Let us suppose now that the spectrum of the disturbance and/or noise is continuous and always includes a sinusoidal excitation at the frequency of the oscillator poles. This
excitation will contribute a linearly increasing part (forced response) to the oscillator natural response as shown in Fig. 4. Depending on the phase difference between the natural and forced responses, the oscillator output will either increase immediately after adding the external excitation or first decrease and after that increase (see Fig. 4).

The decreased oscillations will increase the amplifier gain \( A_{OL} \) above its steady state value \( A_{OL1} \), and the increased gain will shift the oscillator poles in the right half of the \( s \) plane. The oscillator natural response will increase, compensate for the disturbance or noise, and the poles will move back towards the \( jo \) axis (recall Fig. 3).

The increased oscillations will decrease the amplifier gain \( A_{OL} \) below its steady state value \( A_{OL1} \), and the decreased gain will shift the oscillator poles in the left half of the \( s \) plane. The oscillator natural response will decay, and the oscillator steady state will be defined by its steady state response to the input signal \( S_{in} \) in accordance with the network function (6).

Since in the new steady state, caused by a disturbance or noise, the loop gain \( A_{OL} \beta(jo) \) will be smaller than unity, the network function (6) will be finite, and so will be the oscillator output. The specific value of the amplifier gain \( A_{OL} \) in the new steady state and thus the specific value of the network function and the oscillator output depend on the magnitude of the disturbance or noise and on the nonlinear properties of the amplifier.

Note that under influence of a disturbance or noise, the oscillator operates as a frequency-selective amplifier with the total gain described by the network function (6). At the frequency of oscillation, the total gain is at maximum and, at the other frequencies, it decreases in accordance with the frequency behavior of the feedback transmission \( \beta(jo) \). The disadvantage of the use of the nonlinear properties of the amplifier is that its output signal becomes distorted.

To avoid distortions, it is better to control with the help of a negative feedback the value of \( \beta(jo) \) in such a way that the
amplifier output is always within one third of its full scale [see Fig. 3(d)]. This can be done, for example, by using in the feedback network photo- or thermo-resistors, which values depend on average, not instant, magnitude of oscillations.

III. THE NYQUIST STABILITY CRITERION

We shall finally illustrate how one can easily decide whether \( A_{OL} > A_{OL1} \) shifts the oscillator poles from their steady-state position on the \( j \omega \) axis to the right or to the left of this axis.

Let us first assume that the denominator in (6) has only two zeros:

\[
Q(s) = 1 - A_{OL} \beta(s) = (s - z)(s - z^*),
\]

\[
|Q(s)| = |(s - z)(s - z^*)|.
\]

\[
\Phi_Q(s) = \Phi_{(s - z)} + \Phi_{(s - z^*)}.
\]

Let us also assume that the two zeros of \( Q(s) \) lie in the right half of the \( s \) plane, as shown in Fig. 5. We consider as well that for a passive \( \beta(j \omega) \), \( Q(s) \) has no poles in the right half of the \( s \) plane.

If we now encircle the entire right half of the \( s \) plane with a closed contour (so called Nyquist D-contour), as shown in Fig. 5(a), another closed contour, \( Q(j \omega) \), is created in the complex \( Q(s) \) plane around its origin, \( Q(s) = (0, j0) \), as shown in Fig. 5(b).

The function \( Q(s) \) is said to map the right half of the \( s \) plane inside the contour \( Q(j \omega) \) in the \( Q(s) \) plane. [Note also that contour \( Q(j \omega) \) in Fig. 5(b) encircles the origin of the \( Q(s) \) plane in the same direction as the right half of the \( s \) plane is encircled in Fig. 5(a).]

The above result is known as the Principle of the Argument. This principle can be easily understood if one considers an opposite case. Consider, for example, that \( Q(j \omega) \) does not encircle the origin of the \( Q(s) \) plane, as sown in Fig. 5(c). In this case, the vector \( Q(j \omega) \) will not rotate by 720° for \(-\infty < j \omega < +\infty \) as Fig. 5(a) and (8) dictate.

We are now ready to formulate the Nyquist stability criterion for a positive-feedback oscillator combined of a stable feedback network \( \beta(j \omega) \) and a stable amplifier \( A_{OL} \): such an oscillator is unstable if its loop gain \( A_{OL} \beta(j \omega) \) does encircle the point \( (1, j0) \), as shown in Fig. 6.

Fig. 6 illustrates that \( A_{OL} \beta(j \omega) > 1 \) can cause both unstable and stable operations of a circuit. Or in other words, \( A_{OL} \beta(j \omega) > 1 \) can shift the circuit’s poles either to the right or to the left of the \( j \omega \) axis, depending on the specific behavior of the feedback transmission \( \beta(j \omega) \).

REFERENCES