

DESIGN CONSIDERATIONS IN STATE-OF-THE-ART SIGNAL PROCESSING AND PHASE NOISE MEASUREMENT SYSTEMS

by

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Introduction

The recent rapid improvement of oscillator phase noise has resulted in significantly more stringent requirements for signal handling equipment. However, information concerning the phase noise performance of the two most important types of circuits - amplifiers and mixers - is often difficult to find. Some general principles are presented which allow one to estimate the phase noise performance of an amplifier. Also, techniques are described which permit one to obtain the best possible results from the traditional double balanced mixer. A measurement set-up which has 15 to 25 dB improvement in its noise floor is shown in detail to illustrate proper mixer drive and termination. Although traditional circuits can with extreme care achieve $S_{\phi} = -175$ dB or slightly better, this is not sufficient for all present requirements. One technique to obtain an additional improvement of 10 to 40 dB in measurement system noise is to reduce the mixer and amplifier contributions to the noise floor by the use of correlation techniques. A circuit to accomplish this is discussed along with some preliminary results.

One of the most frequently needed systems in the study and use of oscillators is the phase-lock loop. However, since the performance of this system is often incidental to the ultimate goal of the experimenter, e.g. the measurement of phase noise, the design of such a system is sometimes given too little consideration, resulting in unanticipated difficulties and wasted time. The design of an extremely simple phase-lock loop which is suitable for almost all high stability oscillator applications is discussed with particular attention to the advantages over more traditional circuits.

I. PHILOSOPHY OF LOW NOISE AMPLIFIER DESIGN

If an amplifier were driven from a noiseless oscillator, then the output phase spectrum would typically have a flicker noise region at low frequencies and a white noise region at higher frequencies. The break between the two is usually between one and one hundred Hz and the white noise extends out to the bandwidth of the amplifier. The source of the white noise modulation can be identified and the magnitude estimated, but similar generalizations can not be made for the flicker noise. Nevertheless, empirical guidelines can be established which should ensure against unnecessarily poor flicker noise performance.

Provided that the integrated noise of the amplifier over its entire bandwidth is small compared to the signal power, half the thermal noise power contributes to the phase modulation of the signal. Thus the spectral density of phase fluctuations due to noise of the amplifier is

$$S_{\phi}(f) = S_{\phi} / 2P_s \quad (1)$$

where S_{ϕ} is the spectral density of the noise power and P_s is the power available to a matched load. For Johnson noise, the most common situation,

$$S_{\phi} = kT = 4 \times 10^{-21} \text{ J}$$

and the achievable phase noise performance is -184 dB below a 1 rad²/Hz for a 1 V_{rms} signal from a 50Ω source.

In contrast to the white phase noise which is added to the carrier by the amplifier, the flicker phase noise is produced by direct phase modulation in the active element. It has been found empirically that a transistor stage which does not use emitter degeneration typically has phase noise given by $S_{\phi}(f) \approx 10^{-11}$ rad²/f. However the use of local RF negative feedback can reduce this noise power by as much as a factor of 10⁴. [1] Passive elements can also contribute to the flicker phase noise. [Electrolytic, ceramic, and silver mica capacitors and carbon composition resistors can give excessive flicker noise and should only be used in non-critical locations.]

Three design requirements for low phase noise amplifier design follow from the above discussion. Firstly, each stage of an amplifier must incorporate emitter degeneration to minimize the flicker phase modulation. Secondly, critical passive components should be examined for excessive phase noise. Finally, the signal level must be always maintained at a high enough level to achieve the desired white phase noise level.

In order to illustrate the influence of this philosophy on the design of an amplifier, a new isolation amplifier is described. This amplifier was developed to provide a high degree of isolation between very low phase noise RF signals which are used to compare atomic and other frequency standards.

II. WIDE-BAND LOW-NOISE ISOLATION AMPLIFIER

The amplifier shown in Figure 1 is designed to operate from one to several hundred megahertz. In order to minimize current drain a method of achieving high isolation which used a small number of stages was needed. This requirement was satisfied by a cascaded pair of common base transistor stages, Q_1 and Q_2 . A signal applied to the output port propagates towards the input through the collector-base capacitance of Q_1 . The 2N3904 was selected because of its small output capacitance, 4 pF. Since the base of Q_1 is grounded through a capacitor and the emitter looks into the high output impedance of the preceding stage, the signal is low pass filtered. A second stage of filtering is performed by the transistor Q_2 in the same way. It is also possible for a signal to propagate from the output to input through the bias chain. Transmission through this path is reduced to the same level as transmission through the transistor chain by the cascaded low

pass filters. Typical isolation which is achieved is greater than 120 dB at 5 MHz degrading to 100 dB at 50 MHz.

The common emitter input stage determines the collector current of the transistors. The 27Ω dc emitter resistance produces an average collector current of 40 mA. Noise performance is generally best when the amplifier operates well within the class A region. With a 50Ω load this amplifier can produce an output of nearly 1 V_{rms} (13 dBm) with minimum distortion. The gain of the amplifier is determined by the load resistance and the unbypassed portion, 27Ω, of the emitter resistance. With the values shown, the full output swing occurs for an input of approximately 1.5 V_{ptp}.

The white noise floor which one would estimate for this amplifier is S_φ = -184 dB. The measured noise floor is shown as curve A of Figure 2. The noise floor appears to be only -174 dB, but since this level corresponds to the measurement system noise it can only be said that the amplifier is not worse than this. The measured flicker phase noise of the amplifier is S_φ = 10⁻¹⁴ rad²/f. This performance level is reached because each transistor has a reasonable amount of local RF negative feedback. The emitters of Q₁ and Q₂ both look into the high dynamic impedance of the preceding stage while the emitter of Q₃ has the unbypassed 27Ω resistor. For a given application, this unbypassed resistor should be made as large as possible, limited only by the necessity of having full output voltage swing.

In addition to achieving low noise levels it is necessary to minimize microphonics and pickup of power-line frequencies and other signals. For this reason, no use has been made of filter inductors or coupling transformers. It is also possible that temperature changes could cause sufficient collector current variation to degrade the flicker performance. As a result, the diode has been included in the bias chain to further stabilize the collector current. It should be placed in physical contact with transistor Q₃. The amplifier has been constructed on a double sided printed circuit board measuring 3.25 cm x 9 cm. The art work for this circuit board is available from the authors.

III. DOUBLE BALANCED MIXERS AND PHASE NOISE MEASUREMENT SYSTEMS

The most common and also most sensitive method of measuring phase noise is to use a double balanced mixer. If the input ports are driven by quadrature signals, then the output voltage is proportional to the phase deviation of the input signals from the quadrature condition. The spectral density of the phase noise can be calculated from the very simple expression

$$S_{\phi}(f) = \left[\frac{V_n(f)}{V_s} \right]^2 \quad (2)$$

where V_n²(f) is the noise density in units V²/Hz at the output of the mixer and V_s is the sensitivity of the mixer in V/rad.

A variety of circuits for the measurement of phase noise have been discussed extensively in the literature. Here we will look closely at specific problem areas common to all circuits using double balanced mixers: components, input drive levels and output termination. Data are presented which show that proper treatment of these details results in a 15 to 25 dB improvement in the performance of phase noise measurement systems.

Figure 3 shows a typical double balanced mixer.

The best performance has been obtained with units which use hot carrier diodes in the ring. Some differences may also result from the type of transformers in the coupling circuits. The noise observed at the output of the mixer, consisting of mixer and amplifier contributions, is nearly constant over a range of input power level. However, the output signal, proportional to the phase fluctuations, increases with the drive power. The best signal-to-noise ratio for Fourier frequencies in the white noise region is obtained at very high drive levels. For one type of mixer, using a single diode in each arm, the best noise floor was obtained with approximately 30 mA rms current at each input. This drive level exceeded the manufacturer's maximum drive current specification. The optimum drive is not necessarily the same for all Fourier frequencies. The same mixer performed best below 40 Hz at lower drive level. Since such a double balanced mixer is a dynamic impedance the average drive current does not sufficiently describe the operating conditions. The optimum method of coupling to the mixer also depends upon the output impedance of the signal source. Although the use of 50Ω pads to attenuate the drive level is traditional, a series resistor whose value is chosen to set the desired current often gives superior performance. The improvements which are observed may be due to reduced ringing of the drive currents.

The signal-to-noise ratio at the mixer output is also affected by the type of termination used. Since the mixer has a low output impedance, near 50Ω, the dc termination must be high impedance compared to 50Ω. Failure to observe this may result in 6 dB or more loss in signal level. However, it has been determined empirically that the mixer must be terminated differently at RF. In the circuit shown in Figure 4 the impedance to ground at the output of the mixer is 1 kΩ at dc and approximately 50Ω at 10 MHz. The net result of the high drive level and the output termination is illustrated in Figure 5, which shows the beat frequency between the two oscillators in Figure 4 at low drive level (sinewave) and high drive level (clipped waveform). The slope of the clipped waveform at the zero crossings is more than twice the slope of the sinewave. It follows from Eq. (2) that the noise floor is improved more than 6 dB by this technique. The increase in slope is not realized without appropriate termination, but the optimum circuit has not been determined. The noise floor achieved with the circuit of Figure 4 is shown in curve C of Figure 2. The spectral density of phase is -150 dB at 1 Hz and drops to a floor of -176 dB.

Several other special circumstances may occur. One may wish to measure the signal from a device which has insufficient output power to drive a double balanced mixer. Figure 6 shows a simple buffer amplifier which may be used under the circumstances. In keeping with the stated philosophy of amplifier design, this circuit can drive a mixer with a nominal 50Ω input impedance with a 1 V_{rms} signal in class A operation. The mixer input impedance appears as an unbypassed 200Ω in the emitter circuit which results in excellent flicker noise performance. As shown in curve B of Figure 2, the spectral density of phase is -149 dB at 1 Hz and falls off to the noise floor of the measurement system. The 10 dB improvement in flicker noise over the previously described isolation amplifier is probably due to the greater emitter degeneration or lower intrinsic flicker noise in the 2N5943 transistor compared to the 2N3904 or both. The 2N5943 was suggested for use in this circuit by Charles Stone of Austron Inc.

If the device being tested is capable of more output power than a standard double balanced mixer can accept, then it is possible to achieve even lower noise floors. Provided the driving voltage exceeds about 1 V_{rms}, it is possible to use a high level mixer. Such a device has more than one diode in each leg of the ring and is therefore able to achieve higher output voltage without a

corresponding increase in the noise. The circuit diagram of the mixer in Figure 3 shows two diodes in each leg. Using a mixer with three diodes per leg a noise floor of -184 dB was achieved with a drive level of 1.6 V_{rms}. If the oscillator's output impedance is low but the voltage is insufficient to drive a high level mixer a step-up transformer can be used to obtain the appropriate drive voltage. Since the signal and noise power increase by the same ratio, the spectral density of phase of the device under test is unchanged but the noise floor of the measurement system is reduced.

IV. CORRELATION TECHNIQUE

With all of the improvements described the traditional double balanced mixer phase noise measurement system is unable to resolve the noise floor of the best oscillators and amplifiers.

If times 20 or more frequency multiplier chains with noise levels 20 dB below that of the measurement system shown in Figure 2 were available, then that would solve the present problem. So far we are unaware of such multiplier chains, although some prototype multiplier chains show white noise floors 5 to 10 dB below Figure 2. It would also be convenient if the measurement system were broadband so as to accept carrier frequencies from approximately 1 to 100 MHz. Figure 7 shows the block diagram of a phase noise measurement system which is inherently very broadband and also has the capability of improving the measurement system noise by at least 20 dB. It consists primarily of two equivalent traditional phase noise measurement systems.

At the output of each double balanced mixer there is a signal which is proportional to the phase difference, $\Delta\phi$, between the two oscillators and a noise term, V_{N1} , due to contributions from the mixer and amplifier. The voltages at the input of each bandpass filter are

$$\begin{aligned} V_1(\text{BP filter input}) &= A_1 \Delta\phi(t) + C_1 V_{N1}(t) \\ V_2(\text{BP filter input}) &= A_2 \Delta\phi(t) + C_2 V_{N2}(t) \end{aligned} \quad (3)$$

where $V_{N1}(t)$ and $V_{N2}(t)$ are substantially uncorrelated. Each bandpass filter produces a narrow band noise function around its center frequency f:

$$\begin{aligned} V_1(\text{BP filter output}) &= A_1 [S_\phi(f)]^{1/2} B_1^{1/2} \cos [2\pi ft + \psi(t)] \\ &+ C_1 [S_{VN1}(f)]^{1/2} B_1^{1/2} \cos [2\pi ft + n_1(t)] \\ V_2(\text{BP filter output}) &= A_2 [S_\phi(f)]^{1/2} B_2^{1/2} \cos [2\pi ft + \psi(t)] \\ &+ C_2 [S_{VN2}(f)]^{1/2} B_2^{1/2} \cos [2\pi ft + n_1(t)] \end{aligned} \quad (4)$$

where B_1 and B_2 are the equivalent noise bandwidths of filters 1 and 2 respectively. Both channels are bandpass filtered in order to help eliminate aliasing and dynamic range problems. The phases $\psi(t)$, $n_1(t)$ and $n_2(t)$ take on all values between 0 and 2π with equal likelihood. They vary slowly compared to $1/f$ and are substantially uncorrelated. When these two voltages are multiplied together and low pass filtered only one term has finite average value. The output voltage is

$$\begin{aligned} V_{\text{out}}^2 &= 1/2 A_1 A_2 S_\phi(f) B_1^{1/2} B_2^{1/2} + D_1 \langle \cos[\psi(t) - n_1(t)] \rangle \\ &+ D_2 \langle \cos[\psi(t) - n_2(t)] \rangle + D_3 \langle \cos[n_1(t) - n_2(t)] \rangle. \end{aligned} \quad (5)$$

For times long compared to $B_1^{-1/2} B_2^{-1/2}$ the noise terms D_1 , D_2 and D_3 tend towards zero as $1/\sqrt{t}$. Limits in the reduction of these terms are usually associated with harmonics of 60 Hz pickup, dc offset drifts, and nonlinearities in the multiplier. Also if the isolation amplifiers have input current noise then they will pump current through the source resistance. The resulting noise voltage will

appear coherently on both channels and can't be distinguished from real phase noise between the two oscillators. One half of the noise power appears in amplitude and one half in phase modulation.

Curve A of Figure 8 shows $S_\phi(f)$ for the mixer and dc amplifiers in Channel 1 and 2 when used separately. The mixers in this case have three diodes in each leg instead of the two shown in Figure 3 and are driven with approximately 5 V_{ptp} at 5 MHz from 10 Ω source impedance using 33 Ω series resistors. Curve B Figure 8 indicates the correlated component of this noise between the two channels. In order to predict performance in a specific measurement using this scheme, the noise level of the isolation amplifiers used would have to be added to Curve A and proportionately to Curve B.

The above data clearly indicate that significant improvements over any presently existing phase noise measurement system can be obtained using correlation techniques. Such improvements are vitally necessary in order to measure present state-of-the-art signal processing equipment and to test future components and circuits. The simple single frequency correlator used in this experiment could be replaced by a fast digital system which would simultaneously compute the correlated phase noise for a large band of Fourier frequencies. Ultimate noise floors could probably be reduced 40 dB below the noise level of a single channel.

V. PHASE-LOCK TECHNIQUES

One of the most ubiquitous elements of phase noise measurement systems is the phase-lock loop.[2,3] When the phase noise of a pair of oscillators is measured a phase-lock loop is normally used to maintain a condition of approximate phase quadrature. For accurate measurements it is necessary to keep the phase error less than about 1/6 rad despite any initial frequency offset between the two oscillators or any frequency drift during the course of the measurement. The phase-lock loop is usually the most neglected element of the measurement system because its purpose is only tangential to the measurement requirements. As a consequence it often performs marginally. In this section the requirements for a phase-lock system are discussed and an extremely simple yet elegant circuit is presented which more than meets these requirements.

A specific example illustrates the problem: Design a feedback loop to lock a 5 MHz VCO to a reference oscillator with a unity gain frequency of .16 Hz and calculate the open loop frequency difference for phase error of 1/6 rad. The VCO has a tuning rate of 5×10^{-3} Hz/V and the phase deviation from quadrature. One solution, the first order phase-lock loop, is shown in Figure 9. The 50 Ω resistor and 0.1 μ F capacitor are for proper termination of the mixer and do not contribute appreciably to the frequency response of the phase-lock loop. The open loop gain of this servo is

$$\begin{aligned} G_{\text{servo}}(\omega) &= \frac{2 (5 \times 10^{-3} \text{ Hz/V}) (0.17 \text{ V/rad}) G_{\text{amp}}(\omega)}{j\omega} \\ &= \frac{5.3 \times 10^{-3} G_{\text{amp}}(\omega)}{j\omega} \end{aligned} \quad (6)$$

where $G_{\text{amp}}(\omega)$ is the frequency response function of the amplifier. For the first order loop the amplifier has constant gain and the open loop gain of the servo system falls off at 6 dB/octave at all frequencies as shown in the dashed curve of Figure 10. This type of response results from the fact that a phase error measured at the mixer is corrected by changing the frequency of the VCO, i.e., the feedback loop contains one inherent integration.

To produce a unity gain frequency of 0.16 Hz, or an attack time of one second, requires amplifier gain of 185. The loop produces negligible correction for Fourier frequencies greater than 1 Hz and may therefore be used to make phase noise measurements in this frequency range. The open-loop frequency offset which corresponds to a 1/6 rad phase error is

$$\Delta v = (1/6 \text{ rad}) (0.17 \text{ V/rad}) (5 \times 10^{-3} \text{ Hz/V}) (185) = 2.5 \times 10^{-2} \text{ Hz} \quad (7)$$

or only 5 parts in 10^9 at 5 MHz. It is often difficult to achieve and maintain such a small frequency offset. However, the following circuit increases the dc gain of the feedback loop by 10^5 and decreases the phase error proportionally. The improved performance is achieved by adding a stage of quasi-integration which makes a second order loop as shown in Figure 11. The first stage of amplification is the same as in the first order loop. The second stage has gain equal to R_1/R_2 for Fourier frequencies larger than $1/(2\pi R_1 C_1)$. If $R_1 C_1$ exceeds the attack time of the first order loop the new phase-lock loop is stable. It is critically damped when $R_1 C_1$ is approximately four times the attack time and has good step response for $R_1 C_1$ between 1 and one and five times the attack time. The solid curve in Figure 10 is the magnitude of the open loop gain of the second order loop assuming $R_1 = R_2$. Figure 12 compares the step response of a first order loop to that of a second order loop with $R_1 C_1$ equal to the attack time.

The second order loop increases the long-term gain by the open loop gain of the second operational amplifier provided that the leakage resistance, R_L , of the capacitor is sufficiently large. The open loop frequency offset between the two oscillators which can be tolerated is therefore increased to a limit determined either by the maximum voltage swing of the second amplifier or the maximum tuning available in the VCO. The second order loop can be implemented with a single operational amplifier rather than the two which were shown for clarity. In this case the attack time is adjusted by varying the input resistor, R_2 , of the operational amplifier, A_2 and omitting amplifier A_1 in Figure 11.

ACKNOWLEDGMENTS

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REFERENCES

- [1] Donald Halford, A.E. Wainwright, and James A. Barnes, in Proc. 22nd Annual Symposium on Frequency Control (NTIS Accession No. AD 844911), 1968, p. 340.
- [2] F. M. Gardner, *Phase-lock Techniques* (John Wiley, New York, 1966).
- [3] Don Kesner, EDN 5 Jan 1973, p. 54.

ISOLATION AMPLIFIER

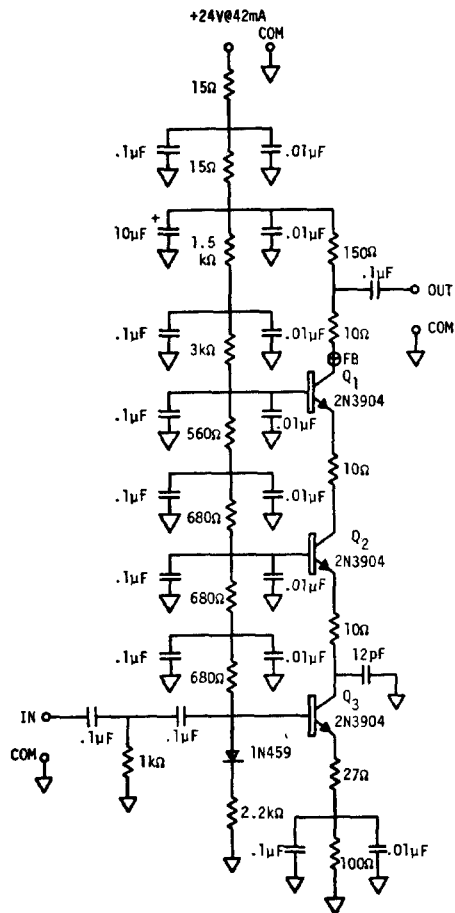


Fig. 1 Schematic of 1 MHz to 200 MHz isolation amplifier. The isolation is -120 dB at 5 MHz. Art work for this amplifier is available from the authors.

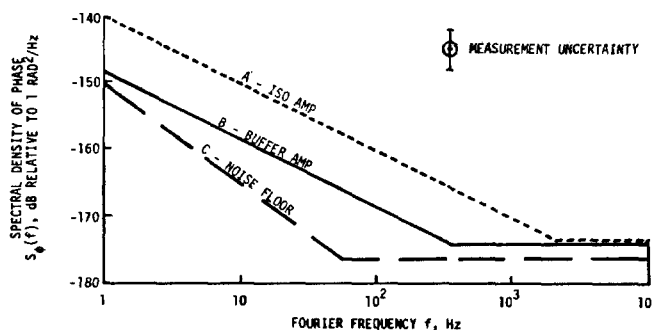


Fig. 2 Spectral density of phase, $S_\phi(f)$, for A) the isolation amplifier of Figure 1 at 5 MHz, B) the buffer amplifier of Figure 6 at 5 MHz, C) noise floor of the measurement systems of Figure 4 with a single diode mixer.

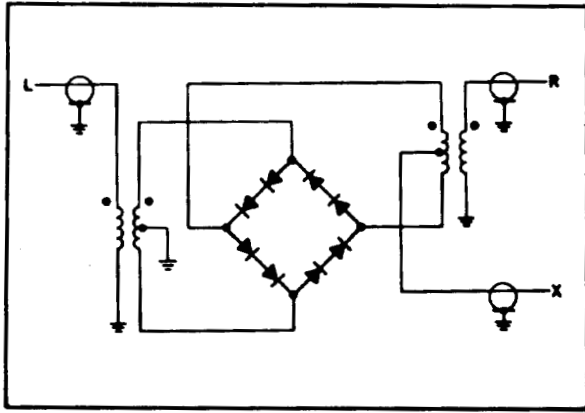


Fig. 3 Schematic of a double balanced mixer. This particular mixer is shown with two diodes in each leg.

BEAT WAVEFORMS

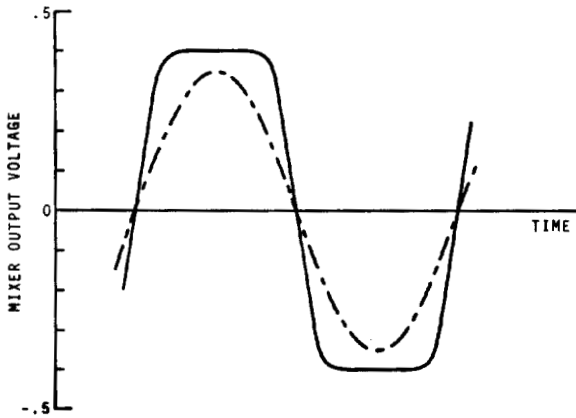


Fig. 5 Filtered waveform at the output (x port) of double balanced mixer due to frequency difference between signals at R and L ports. The solid curve is obtained at high drive levels while the dashed curve is obtained at low drive levels.

PHASE NOISE MEASUREMENT SYSTEM

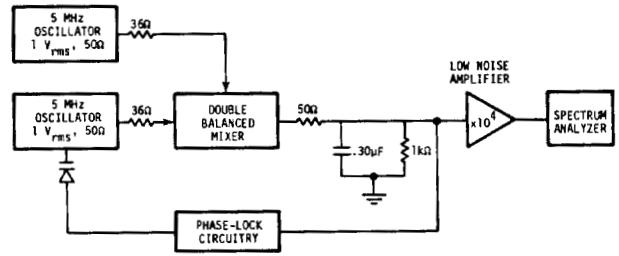


Fig. 4 Typical system for measuring $S_{\phi}(f)$ of a pair of equal frequency oscillators. Noise floor for this system is shown in Figure 2, curve C.

BUFFER AMPLIFIER

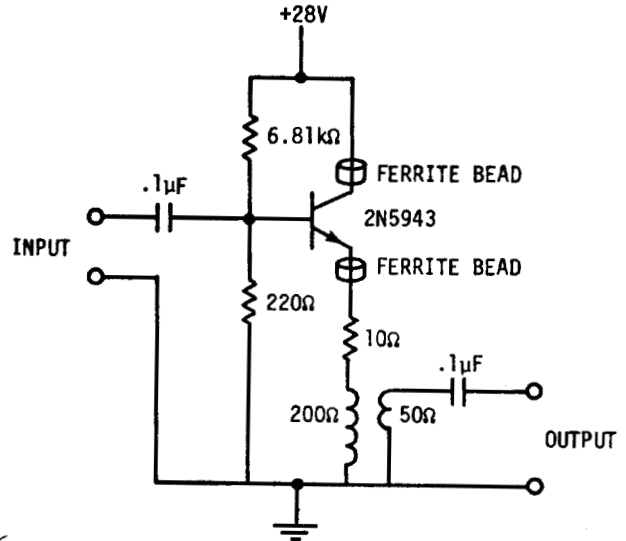


Fig. 6 Very low noise buffer amplifier which can be used to drive mixers. The noise performance is shown in Figure 2 curve B.

PHASE NOISE MEASUREMENT SYSTEM USING CORRELATION TECHNIQUE

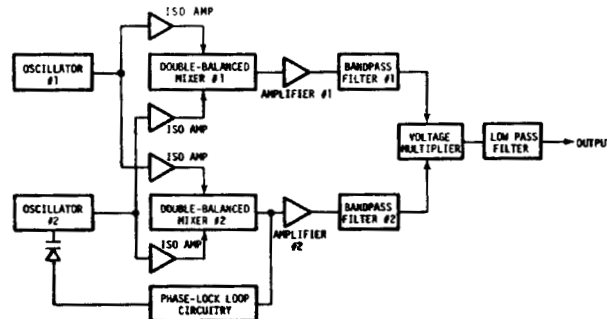


Fig. 7 Block diagram of a new phase noise measurement system which is broadband and features low noise.

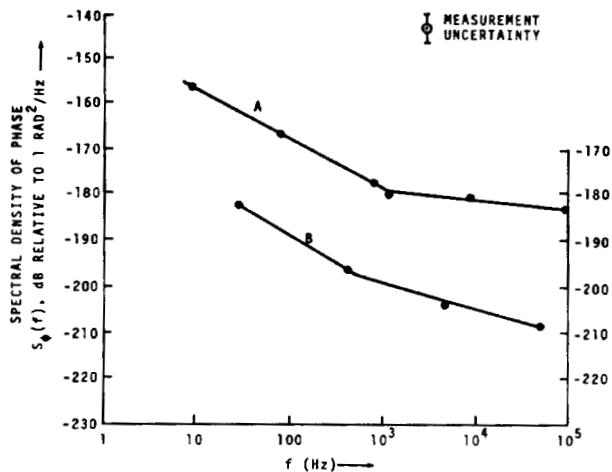


Fig. 8 Curve A shows $S_{\phi}(f)$ for each channel of the measurement system of Figure 7 excluding the isolation amplifiers. Curve B shows the correlated component of $S_{\phi}(f)$ between the two channels.

FIRST ORDER PHASE-LOCK LOOP

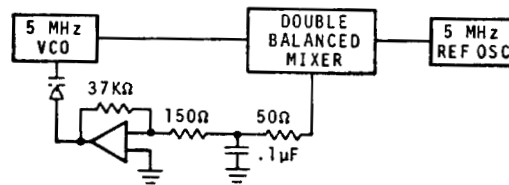


Fig. 9 Schematic of typical first order phase-lock loop. Loop attack time is adjusted by changing amplifier gain.

SECOND ORDER PHASE-LOCK LOOP

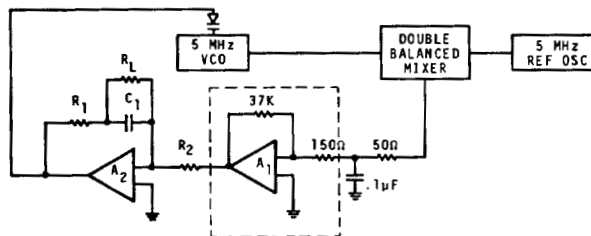


Fig. 11 Second order phase-lock loop. The attack time is adjusted by changing gain as in Figure 9. The time constant R_1C_1 is adjusted to be 1 to 5 times longer than the attack time.

OPEN-LOOP GAIN

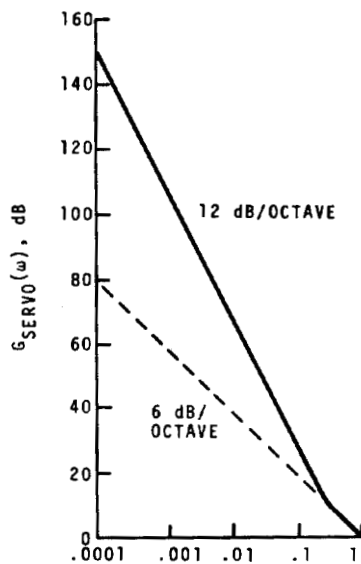


Fig. 10

The dashed curve shows the open-loop gain of the first order phase-lock loop of Figure 9, while the solid curve shows the open-loop gain of the second order phase-lock loop of Figure 11.

PHASE-LOCK LOOP STEP RESPONSE

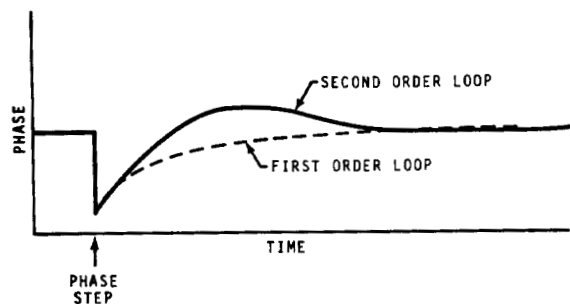


Fig. 12 The dashed curve shows the response of the first order phase-lock loop of Figure 9 to a phase step, while the solid curve shows the response of the second order phase-lock loop of Figure 11, with $R_1C_1 =$ the attack time.