Guidelines for Designing BJT Amplifiers with Low 1/f AM and PM Noise

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Abstract—**In this paper we discuss guidelines for designing linear bipolar junction transistor amplifiers with low 1/f amplitude modulation (AM) and phase modulation (PM) noise. These guidelines are derived from a new theory that relates AM and PM noise to transconductance fluctuations, junction capacitance fluctuations, and circuit architecture. We analyze the noise equations of each process for a common emitter (CE) amplifier and use the results to suggest amplifier designs that minimize the 1/f noise while providing other required attributes such as high gain. Although we use a CE amplifier as an example, the procedure applies to other configurations as well. Experimental noise results for several amplifier configurations are presented.**

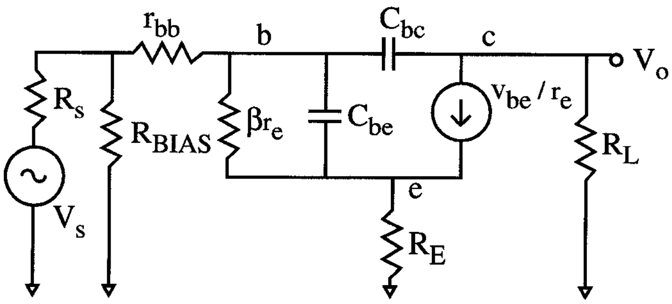

Fig. 1. Hybrid π model of CE amplifier.

I. Introduction

In this paper we discuss the design criteria for low 1/f amplitude modulation (AM) and phase modulation (PM) noise in linear bipolar junction transistor (BJT) amplifiers using the theory discussed in [1]. This theory relates AM and PM noise to current fluctuations, voltage fluctuations, impedance fluctuations, and circuit architecture. We analyze the noise equations of each process for a common emitter (CE) amplifier and use the results to suggest amplifier designs that minimize the 1/f noise while providing other required attributes such as high gain. Although this is done specifically for a CE amplifier, the same procedure can be applied to any amplifier configuration. Guidelines for the design of low 1/f noise CE amplifiers are discussed in detail. We show low 1/f noise results for high gain CE amplifiers at carrier frequencies of 5 MHz and 100 MHz. Experimental results for other configurations (common base [CB], common collector [CC]) are also presented.

II. Noise Equations

As derived in [1], for an amplifier with voltage gain = $G_V e^{j\delta}$ (where G_V is the magnitude and δ is the phase shift), the AM noise is given by:

$$\frac{1}{2}S_a(f) = \frac{1}{2}\left[\frac{\Delta G_V(i,V,\nu,f)}{G_V}\right]^2 \frac{1}{BW} + \frac{kTFG}{2P_0}, \tag{1}$$

and the PM noise by:

$$\frac{1}{2}S_\phi(f) = \frac{1}{2}\frac{\Delta\delta^2(i,V,\nu,f)}{BW} + \frac{kTFG}{2P_0}. \tag{2}$$

In the first term of (1), $\Delta G_V(i,V,\nu,f)$ refers to the fluctuations in the voltage gain due to current noise, voltage noise, and impedance fluctuations, and BW is the bandwidth of the measurement. $\Delta G_V(i,V,\nu,f)$, which depends on dc current, dc voltage, circuit parameters, carrier frequency, and Fourier frequency, is the result of baseband 1/f noise up-converted to the carrier frequency. Similarly, $\Delta\delta(i,V,\nu,f)$ in (2) refers to the fluctuations in the phase shift of the amplifier due to current noise, voltage noise, and impedance fluctuations, and is also a function of dc current, dc voltages, circuit parameters, carrier frequency, and Fourier frequency. The second term in (1) and (2) is the thermal noise of the amplifier. In this term, k is Boltzmann's constant, T is the temperature in kelvin, F is the noise figure, G is the power gain of the amplifier, and P_0 is the output power of the amplifier. Applying (1) and (2) to the CE amplifier shown in Fig. 1 yields the noise equations:

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$$\begin{aligned}
&\frac{1}{2}S_a(f) = \\
&\frac{1}{4}\left(\frac{r_e}{r_e+R_E+r_g/\beta}\right)^2\gamma^2+\frac{1}{4}\left(\frac{r_g/\beta}{r_e+R_E+r_g/\beta}\right)^2 \\
&\times\left(\frac{\Delta r_g}{r_g}\right)^2+\frac{1}{4}\left(\frac{R_E}{r_e+R_E+r_g/\beta}\right)^2\left(\frac{\Delta R_E}{R_E}\right)^2 \qquad (3)\\
&+\frac{1}{4}\left(\frac{\Delta R_L}{R_L}\right)^2+\frac{1}{2}\delta^2\left(S_\phi(f)-\frac{kTFG}{2P_0}\right)+\frac{kTFG}{2P_0},
\end{aligned}$$

$$\begin{aligned}
&\frac{1}{2}S_\phi(f) = \\
&\frac{1}{4}(\omega G_0[r_e+R_E+r_g])^2\Delta C_{bc}^2+\frac{1}{4}(\omega C_{bc}G_0)^2 \\
&\times(\Delta R_E^2\Delta r_g^2)+\frac{1}{4}\left(\frac{r_e}{r_e+R_E+r_g/\beta}\right)^2 \\
&\times(\omega C_{bc}G_0[r_g])^2\gamma^2+\frac{kTFG}{2P_0},
\end{aligned} \qquad (4)$$

where $r_g = R_s \| R_{\text{BIAS}} + r_{bb}, R_L = R_L' R_{\text{BIAS}}/(R_s + R_{\text{BIAS}}), \gamma = \Delta I_E/I_E$, δ is the phase shift across the amplifier, and $G_0 = -\frac{R_L}{r_e+R_E+r_g/\beta}$. For details on the derivation of (3) and (4) see [1].

III. Design Guidelines

From the noise equations we can develop guidelines for designing low 1/f AM and PM noise BJT amplifiers. In general, the effect of current fluctuations on AM noise can be reduced by increasing the standing current. For PM noise, the guidelines depend on the dominant poles of the phase equation. Therefore, it is difficult to come up with a unique set of rules for low 1/f PM noise design. The general rule that always applies is to design for the largest possible bandwidth (i.e., maximize frequency of the dominant poles). For this reason we have chosen a specific configuration, a CE amplifier, to develop guidelines for low 1/f AM and PM noise design. The same procedure for deriving guidelines can be applied to any amplifier configuration.

Equation (3) shows that an unbypassed emitter resistance decreases the contribution from current noise and input impedance noise (first two terms) to 1/f AM noise. Similarly, all the terms in the PM noise equation have a factor G_0^2, suggesting that an increase in the unbypassed emitter resistance (gain reduction) will result in a reduction of PM noise in the amplifier. This agrees with experimental data showing that an increase in the unbypassed emitter resistance does reduce the PM noise [2], [3]. The problem with this approach is that as the AM and PM noise decreases, the gain also decreases. Using the noise equations one can derive guidelines for the design of low 1/f noise amplifiers without having to sacrifice the gain. These guidelines are listed below.

A. Large dc Current

The contribution of the current noise in both the AM and PM noise can be reduced by increasing the dc current through the transistor. For example, let us assume that the current noise term dominates the AM noise; that is

$$\frac{S_a(f)}{2}=\frac{1}{4}\left(\frac{r_e}{r_e+R_E+r_g/\beta}\right)^2\gamma^2. \qquad (5)$$

As I_{dc} increases, $r_e(\propto I_{dc}^{-1})$ decreases, and so does the term inside the bracket in (5) for a specific unbypassed emitter resistance. The smaller r_e, the more effective is the unbypassed emitter resistor in reducing the AM and PM noise due to current fluctuations. It is also worth mentioning that $\gamma(\Delta I_E/I_E)$ is not necessarily a constant, but depends on the noise process from which it originates, and can be a function of operating point and transistor type.

B. Servo to Reduce Current Noise

For amplifiers in which the current noise is a problem (even for large I_E) a servo loop can be used to reduce the dc current noise [4]. For the most part, the transistors used in our circuits had very low current noise; thus transconductance fluctuations did not limit the AM (or PM flicker noise). Other transistors might have larger current noise, and the addition of a servo to reduce the current noise could be needed to achieve low levels of AM and PM flicker noise.

C. Transistors with Low Input and Output Capacitances

As mentioned in [1], flicker PM noise is caused by fluctuations in the phase shift of the amplifier. In general, these fluctuations are proportional to the input and output capacitances of the transistor, so it is to be expected that transistors with small capacitances (high f_T) result in better PM noise. This might not be true for all cases, since the PM noise depends on other factors such as current noise and impedance fluctuations. If the current noise in high f_T transistors is large this might turn out to be a problem for both AM and PM noise. It is therefore important to know the current noise level of the transistor to be used, in addition to the input and output capacitances and the f_T.

D. Large Collector-Base Voltage

The dominant term in the PM noise equation (for a CE amplifier) is the collector-base capacitance fluctuations term. The collector-base capacitance is a junction capacitance which depends on the dc voltage at the collector-base terminals (V_{CB}). This capacitance has the form

$$C_{bc}=\frac{K}{(V_{bi}+V_{CB})^n}, \quad 0.1\le n\le 4, \qquad (6)$$

where V_{bi} is the built in potential of the p-n junction, K is a constant that includes the permittivity of the semiconductor and the doping concentration, and n is a parameter that depends on the doping profile of the junction [5]. Equation (6) shows that fluctuations in V_{CB} will result in fluctuations of the collector-base capacitance. The sensitivity of PM noise to V_{CB} fluctuations is obtained from

$$\left(\frac{\partial\phi}{\partial V_{CB}}\right)^2 = \left(\frac{n\omega G_0 r_g K}{(V_{bi}+V_{CB})^{n+1}}\right)^2. \tag{7}$$

Equation (7) demonstrates that higher V_{CB} results in smaller sensitivity of PM noise to collector-base voltage fluctuations (C_{bc} fluctuations).

E. Filters to Reduce Voltage Noise at Transistor Terminals

In many cases fluctuations at the transistor terminals are the result of noise in the power supply. While a large collector-base voltage (V_{CB}) will reduce the effect of these fluctuations in the PM noise, it is also useful to have some kind of filtering of the power supply noise to reduce the effective collector-base voltage noise.

F. Small dc Gain

In addition the amplifier should be designed to have a small dc gain. The reason is that the fluctuations at the base terminal show up at the collector amplified by the dc gain; therefore, a small dc gain will help minimize the effect of ΔC_{bc} in the PM noise.

G. Small Source Impedance

From [1], the phase shift of a CE amplifier (when neglecting C_{be}) is given by:

$$\delta = \frac{\omega C_{bc} R_L (r_e + R_E + r_g)}{r_e + R_E + r_g/\beta}, \tag{8}$$

where $r_g = R_S \| R_{\mathrm{BIAS}} + r_{bb}$. For a high gain CE amplifier with $r_g > r_e + R_E$, the phase shift is proportional to r_g. The use of sources with small impedance will help reduce the phase shift and thus the PM noise performance. Both the ΔC_{bc} term and the $\gamma(\Delta I_E/I_E)$ term are proportional to r_g, and a reduction in R_S will reduce both these terms.

H. Noise in Other Components Used (Resistors, Capacitors, Inductors)

The flicker noise in resistive, capacitive, and inductive components that make up the rest of the amplifier can also result in PM and AM noise if their levels are high. Several terms in the AM and PM noise equations are the result of impedance fluctuations. By carefully choosing low noise components these terms can be reduced considerably. Inductive components, which might be very convenient to use since they provide large impedances without the power dissipation of resistors, have flicker noise that depend on the type of core used. To prevent noisy inductors use air cores whenever possible.

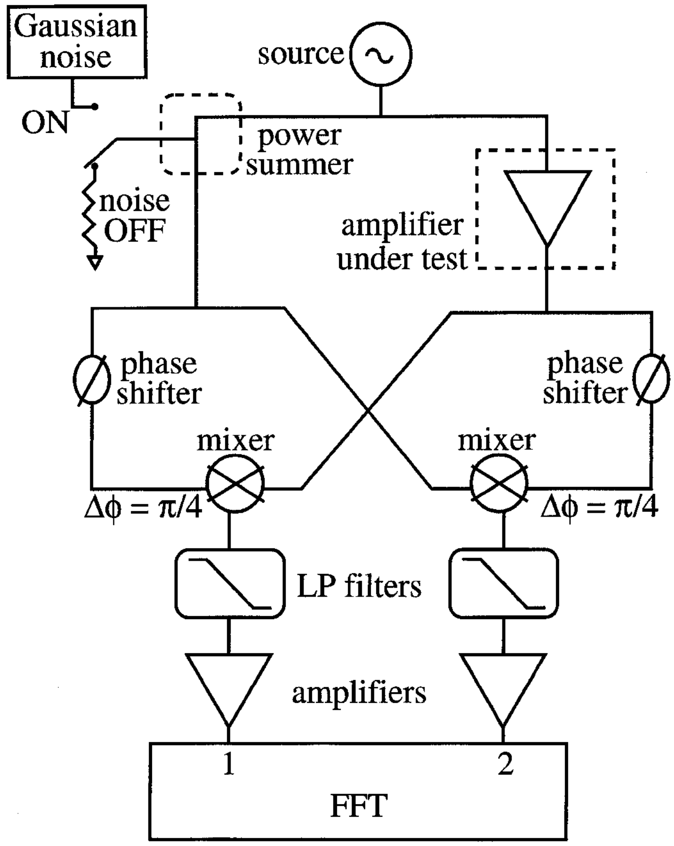

Fig. 2. PM noise measurement system for amplifiers.

As mentioned earlier, some of these guidelines might not apply to other configurations. In general, a large standing current helps reduce AM noise due to current fluctuations, but in some cases, this might jeopardize the PM noise. For example, in cases in which the base-emitter diffusion capacitance C_{be} appears in the dominant pole, a large current might degrade the PM noise since C_{be} depends on the dc collector current. An increase in the current will then cause a decrease in the pole frequency (increase in the phase shift). In addition, current fluctuations will be up-converted to PM noise through C_{be}.

We must also consider thermal noise, shot noise, and heat dissipation in the transistor when designing the amplifier.

IV. Measurement Systems

The low noise levels of our amplifiers required the use of two-channel cross-correlation measurement systems [6], [7]. The PM noise was measured using the system shown in Fig. 2. This setup has two similar phase noise detectors that are fed into a two-channel cross-correlation fast Fourier transform (FFT) signal analyzer. The phase shifter in each detector is adjusted so that the input signals to the mixer are in quadrature. The power spectral density (PSD) of the output voltage of channels 1 and 2 includes PM noise of the test amplifier, noise due to the components in the channel (mixer, amplifier, and FFT analyzer), and a fraction of the AM noise in the source that was "converted" to PM noise in the mixer (see (9) and (10)). The AM to PM conversion factor in a double balance mixer is between −15 dB and −25 dB. The cross-correlation of channels 1 and 2 includes the PM noise in the amplifier and the converted AM noise from the source as shown in (11). The uncorrelated noise (from the independent channels) is averaged away as $N^{-1/2}$, where N is the number of averages.

$$PSD(V_{n1}) = PM_{\mathrm{amp}} + \mathrm{NOISE}_{\mathrm{channel\ 1}} + \beta_a AM_{\mathrm{source}}, \tag{9}$$

$$PSD(V_{n2}) = PM_{\mathrm{amp}} + \mathrm{NOISE}_{\mathrm{channel\ 2}} + \beta_a AM_{\mathrm{source}}, \tag{10}$$

$$PSD(V_{n1} \times V_{n2}) = PM_{\mathrm{amp}} + \frac{\mathrm{NOISE}_{\mathrm{channels\ 1,2}}}{\sqrt{N}} + \beta_a AM_{\mathrm{source}}. \tag{11}$$

In our case the noise floor of the system was limited by the AM noise of the source. At a carrier frequency of 5 MHz, the noise floor was $\mathcal{L}(10\ \mathrm{Hz})_{\mathrm{floor}} \cong -170$ dBc/Hz. At a carrier frequency of 100 MHz, the noise floor was $\mathcal{L}(10\ \mathrm{Hz})_{\mathrm{floor}} \cong -162$ dBc/Hz.

The AM noise was measured using the system shown in Fig. 3. In this system a low noise source drives the amplifier under test. After the amplifier, the power is split and each leg is fed into an AM detector. The AM detector consists of a power splitter, a phase shifter, and a mixer. The phase shifter is adjusted so that the two inputs to the mixer are in phase. The PSD of the output voltage of channels 1 and 2 includes AM noise from the source and the amplifier, and noise from the channel. The cross-correlation of the voltages of channels 1 and 2 includes only AM noise from the source and the amplifier since the uncorrelated noise (from the individual channels) is averaged away as $N^{-1/2}$ (see (12)–(14)). Since in many cases the noise of the source was larger than the amplifier noise, a limiter was used to reduce its AM noise. The approximate noise floor achieved with this setup was $1/2S_a(10\ \mathrm{Hz})_{\mathrm{floor}} \cong -162$ dBc/Hz for a carrier frequency of 5 MHz.

$$PSD(V_{n1}) = AM_{\mathrm{amp}} + AM_{\mathrm{source}} + \mathrm{NOISE}_{\mathrm{channel\ 1}}, \tag{12}$$

$$PSD(V_{n2}) = AM_{\mathrm{amp}} + AM_{\mathrm{source}} + \mathrm{NOISE}_{\mathrm{channel\ 2}}, \tag{13}$$

$$PSD(V_{n1} \times V_{n2}) = AM_{\mathrm{amp}} + AM_{\mathrm{source}} + \frac{\mathrm{NOISE}_{\mathrm{channels\ 1,2}}}{\sqrt{N}}. \tag{14}$$

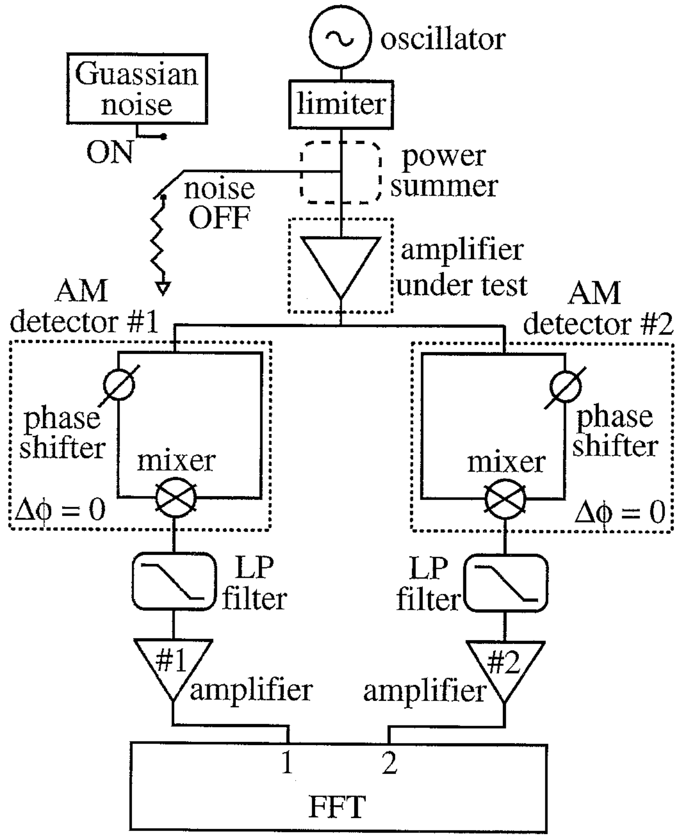

Fig. 3. AM noise measurement system for amplifiers.

V. Experimental Results on Amplifier Flicker Noise

This section shows experimental noise results on different amplifier designs and configurations. Measurements were done on amplifiers at 5 MHz and 100 MHz carrier frequencies.

The first design considered was a 5 MHz CE amplifier. To emphasize the importance of the biasing scheme and the components used in the amplifier, we started with a poor design for flicker noise. Fig. 4 shows the initial CE design. In this amplifier the dc current is 6 mA and the collector-base voltage is 2.8 V. No filtering of power supply noise is included. The large collector resistor combined with the small unbypassed emitter resistance (5 Ω) result in a high rf gain of 22 dB. The dc gain is approximately 0 dB. The 4:1 transformer at the output is used to reduce the output impedance of the amplifier. The first column of Table I shows some parameters for the transistor used (2N2222A).

The AM and PM noise results for this CE amplifier are shown in Fig. 5. The thermal noise level is approximately −154 dBc/Hz. The reason for such a high level was the low output power level (2 dBm) needed to avoid compression. (Due to the small V_{CB} (collector-base dc voltage) and the small dc current, the output power of the amplifier started to compress at 2 dBm.) The flicker noise level is similar for AM and PM noise; at 2 Hz it is approximately −138 dBc/Hz.

TABLE I

PARAMETERS FOR TRANSISTORS USED IN THE 5 MHz CE AMPLIFIERS.

	Transistor in CE #1 (2N2222A)	Transistor in CE #2
f_T	300 MHz	8 GHz
C_{be}	25 pF	1.6 pF
$C_{bc}[V_{CB} = 9\text{ V}]$	8 pF	0.65 pF
F	≤ 4 db	1.5 dB
input current noise 10 Hz ($I_E \cong 18$ mA) (measured)	-210 dBA$_{rms}$/Hz	-178 dBA$_{rms}$/Hz

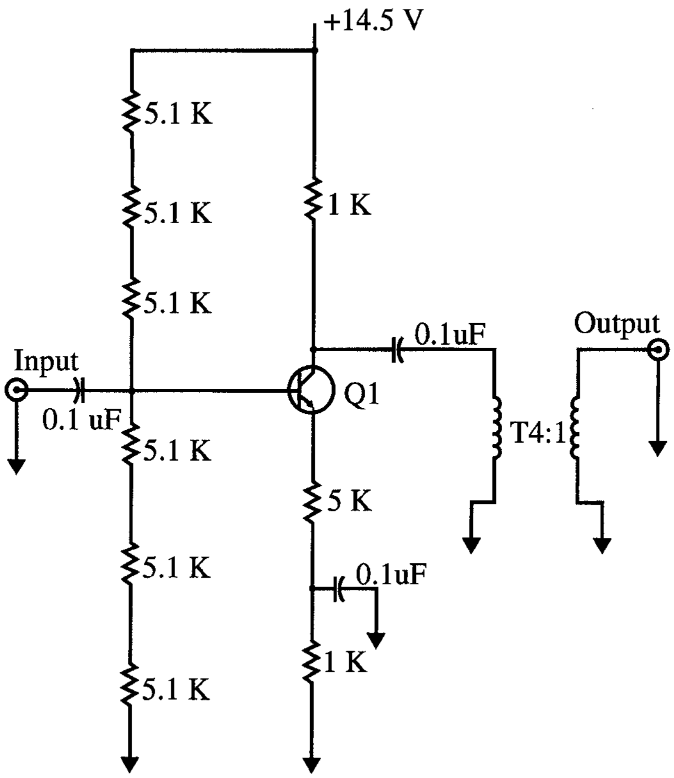

Fig. 4. Circuit diagram of CE amplifier #1.

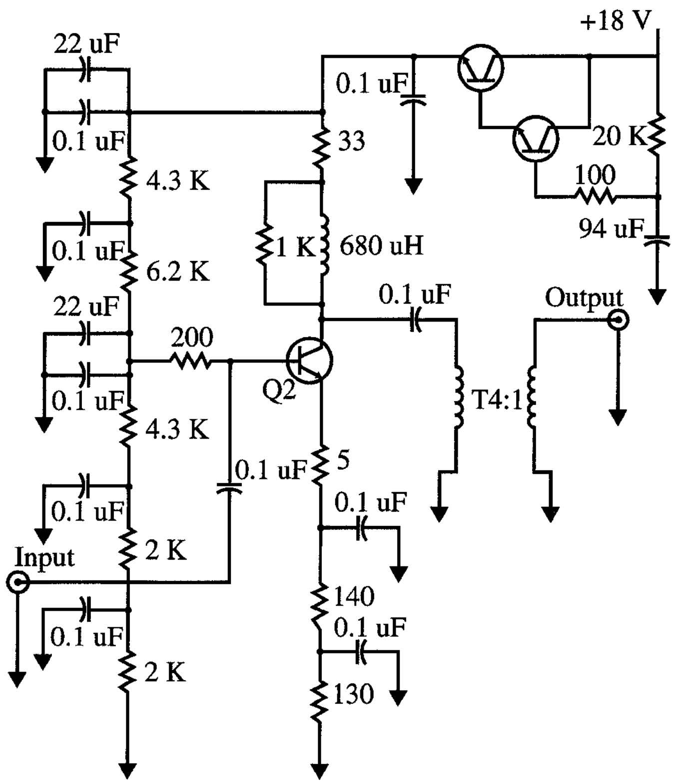

Fig. 6. Circuit diagram for CE amplifier #2.

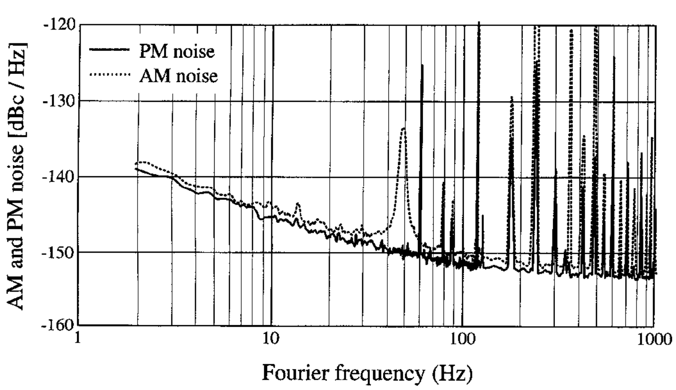

Fig. 5. AM and PM noise for CE amplifier #1.

A separate CE amplifier was designed and built according to the guidelines discussed earlier. The circuit diagram of this CE amplifier is shown in Fig. 6. The dc collector current is 23 mA and $V_{CB} = 9$ V. Several filters for the supply noise were added, including a Darlington pair configuration at the output of the power supply, and several RC filters at the base bias resistive chain. In addition, a transistor with very small input and output capacitances was used (see Table I, column 2 for transistor parameters). A large inductor at the collector was added in parallel to the 1 kΩ resistor. This achieves a low dc gain and does not affect the gain at the rf frequency of operation (5 MHz) due to the inductor's large impedance at that frequency. We expect a gain similar to the gain of the CE amplifier #1 since R_C, R_E (unbypassed emitter resistor), and the output transformer are the same. The resulting rf gain is

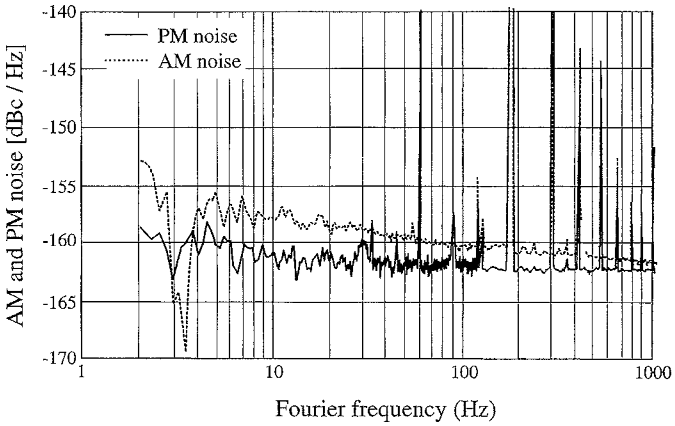

Fig. 7. AM and PM noise for CE amplifier #2.

24.7 dB, somewhat higher than the gain of the CE amplifier #1, primarily due to the larger dc current.

Fig. 7 shows the AM and PM noise results for the CE amplifier #2. The thermal level can be calculated using (15):

$$\mathcal{L}(f)_{\text{thermal}} = \frac{1}{2}S_\phi(f)_{\text{thermal}} = \frac{1}{2}S_a(f)_{\text{thermal}} = \frac{kTFG}{2P_0}. \quad (15)$$

For this amplifier $G = 24.7$ dB, $P_0 = 13$ dBm , so the thermal noise for a $F = 0$ dB is $1/2S_a(f)_{\text{thermal}} = \mathcal{L}(f)_{\text{thermal}} = -177.1$ dBm $+ 24.7$ dB $- 13$ dBm $= -165.4$ dBc/Hz.

From the PM noise trace, the thermal level is approximately -162 dBc/Hz; therefore the noise figure of the amplifier is 3 dB. At 10 Hz from the carrier, the PM noise level is $\mathcal{L}(10\text{ Hz}) = -161$ dBc/Hz, only 1 dB above thermal. At 2 Hz from the carrier, the PM level is approximately $\mathcal{L}(2\text{ Hz}) = -159$ dBc/Hz. This number includes both flicker and thermal noise; the flicker level can be calculated by subtracting the thermal noise:

$$\mathcal{L}_{\text{flicker}}(2\text{ Hz}) \cong 10\log(10^{-15.9} - 10^{-16.2}) \cong -162\text{ dBc/Hz}. \quad (16)$$

The AM level at 10 Hz from the carrier is about $(1/2)S_a(10\text{ Hz}) \cong -158$ dBc/Hz. This value is limited by the noise floor of the measurement system (-162 dBc/Hz) and the thermal noise level (-162 dBc/Hz). The actual flicker level can be calculated by:

$$\begin{aligned}(1/2)S_{a,\text{flicker}}(10\text{ Hz}) &\cong 10\log(10^{-15.8} - 10^{-16.2} - 10^{-16.2}) \\ &\cong -165\text{ dBc/Hz}. \quad (17)\end{aligned}$$

Comparing results for the two different CE amplifiers, we see that at 2 Hz from the carrier, the PM noise was reduced by approximately 20 dB, and the AM was reduced by about 15 dB. These numbers include the contributions of the thermal noise and the measurement system noise; thus, the actual reduction of 1/f noise is probably larger.

A similar CE amplifier design was tried at a carrier frequency of 100 MHz (Fig. 8). This amplifier has a dc current of 19 mA, a large collector-base voltage (9 V), a small dc gain (less than 0.05), and an ac gain of 18.5 dB. As in the 5 MHz CE amplifier #2, a high f_T transistor is used (Table I, column 2); in addition, a Darlington pair voltage regulator and RC filters are used to reduce noise voltage from the supply.

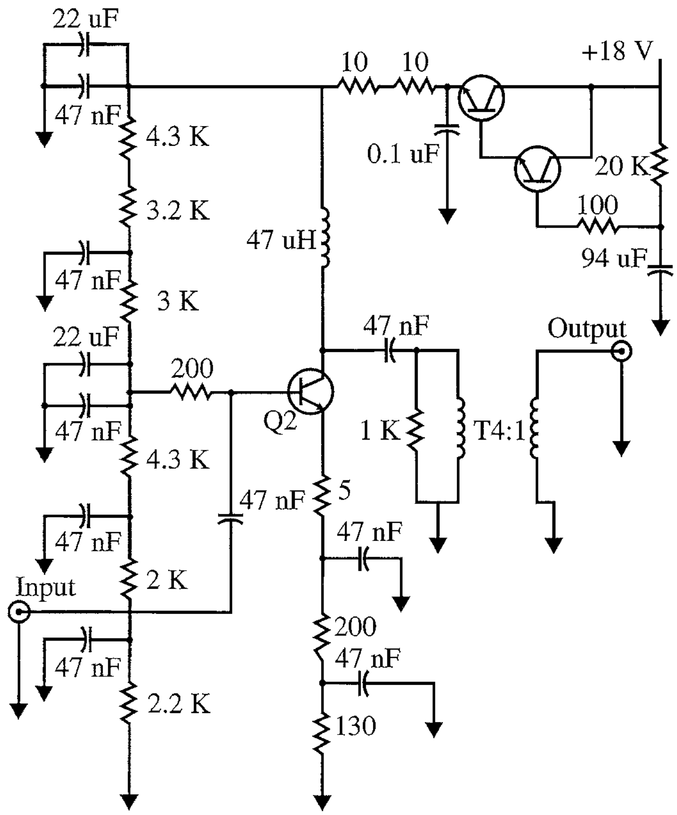

Fig. 8. Circuit diagram for a low noise 100 MHz CE amplifier.

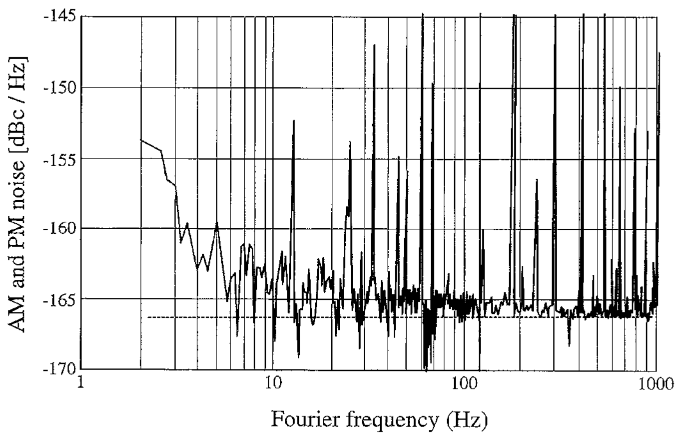

Fig. 9. PM noise for a CE amplifier at 100 MHz.

Fig. 9 shows the measured PM noise for the 100 MHz CE amplifier just discussed. At Fourier frequencies above 100 Hz the PM noise is dominated by the thermal noise (-166 dBc/Hz). This agrees with the theoretical value calculated from (15) for a gain of 18.5 dB, $P_0 = 11.5$ dBm,

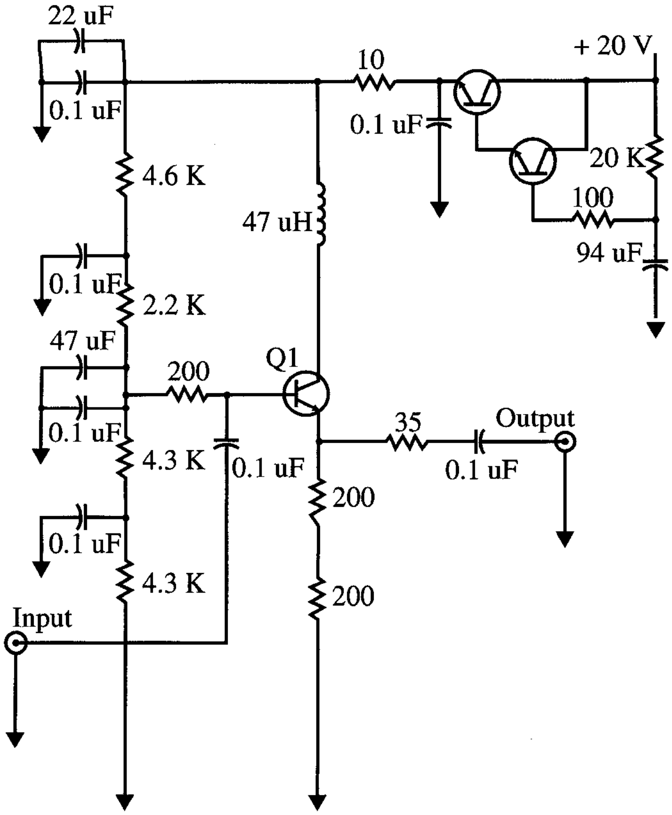

Fig. 10. A 5 MHz CC amplifier.

and $F = 4$ dB:

$$\begin{aligned}\mathcal{L}(f)_{\text{thermal}} &= -177.1\ \text{dBm} + 18.5\ \text{dB} - 11.5\ \text{dBm} + 4\ \text{dB} \\ &= -166.1\ \text{dBc/Hz}.\end{aligned} \tag{18}$$

This calculated thermal noise is shown in the graph as a dotted line. At Fourier frequencies below 20 Hz the PM noise is limited by the measurement system noise. At 10 Hz, the measured PM noise is -162 dBc/Hz, and at 2 Hz the noise is -154 dBc/Hz. These numbers are very close to the measurement system noise floor.

Another BJT configuration studied was the common collector (CC) amplifier. Fig. 10 illustrates the circuit diagram of the 5 MHz CC amplifier tested. In this CC amplifier $V_{CB} = 8.5$ V, and the dc collector current is 24 mA. Parameters for the transistor used are shown in column 1 of Table I. Due to its small rf gain (-1.5 dB) and small phase shift, we expect (from theory) both the AM and PM noise of this amplifier to have very low levels. Fig. 11 shows a plot of the measured PM and AM noise. As expected, the levels are very low; at 10 Hz from the carrier $\mathcal{L}(10\ \text{Hz}) = -169$ dBc/Hz and $1/2S_a(10\ \text{Hz}) = -162$ dBc/Hz. These values are approximately equal to the measurement system noise floor.

Due to its excellent noise characteristics, this stage can be used as a buffer stage either at the input or output of multiple stage amplifiers. One benefit of adding this stage at the input of a CE amplifier is that it would decrease the input impedance to the CE stage, reducing the total phase shift of the CE. As discussed earlier, this should improve the PM noise of the CE stage.

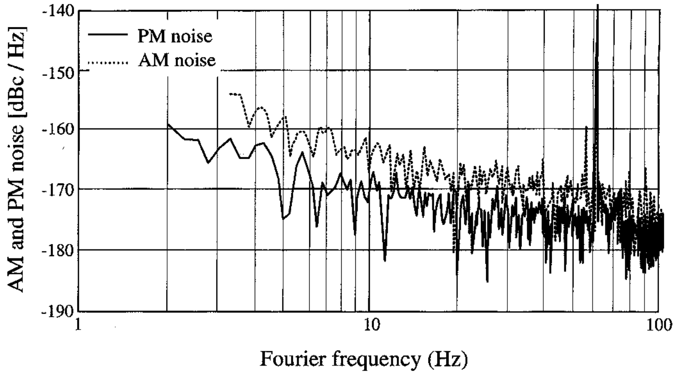

Fig. 11. AM and PM noise for a CC amplifier at 5 MHz.

Another amplifier to be discussed is the 5 MHz distribution amplifier described in [8]. This amplifier, shown in Fig. 12, consists of three common base stages that provide a high isolation between the input and output ports. The dc current through the collectors is approximately 28 mA and the rf gain of a single channel is approximately 7 dB. The details of the design are discussed in [8] and will not be repeated here. Additional filtering of the power supply noise was included to further reduce the already low PM noise of this amplifier.

PM and AM noise results for this amplifier are shown in Fig. 13. Trace 1 (in Fig. 13) shows the measured PM noise for this amplifier. Notice that this trace is close to the noise floor of the measurement system (trace 2). Trace 3 was obtained by subtracting the noise floor of the measurement system from the measured PM noise and includes only the PM noise of the amplifier. The PM noise level at a Fourier frequency of 10 Hz is approximately $\mathcal{L}(10\ Hz) = -168$ dBc/Hz and the thermal noise $\mathcal{L}(f)_{\text{thermal}} = -171$ dBc/Hz. The AM noise seems to be limited by the noise floor of the AM measurement system. At 10 Hz from the carrier $1/2S_a(10\ \text{Hz}) = -160$ dBc/Hz, compared to the noise floor level of -162 dBc/Hz.

VI. Conclusion

We have developed guidelines for designing CE amplifiers with high gain and low 1/f AM and PM noise. These guidelines minimize the up-conversion of baseband voltage noise, current noise, and impedance noise to noise about the carrier. According to the theory, the larger the gain and the phase shift of an amplifier, the higher the levels of AM and PM noise. This was found to be true experimentally. CE amplifiers, which have the potential for high gains and large phase shifts, had the highest noise, but, by following the design criteria, we were able to reduce their 1/f noise considerably with only a modest reduction in the gain. Our 5 MHz CE amplifier, with a 25 dB gain, had AM and PM noise less than -160 dBc/Hz at 10 Hz from the carrier. We also had a 100 MHz CE amplifier (18.5 dB

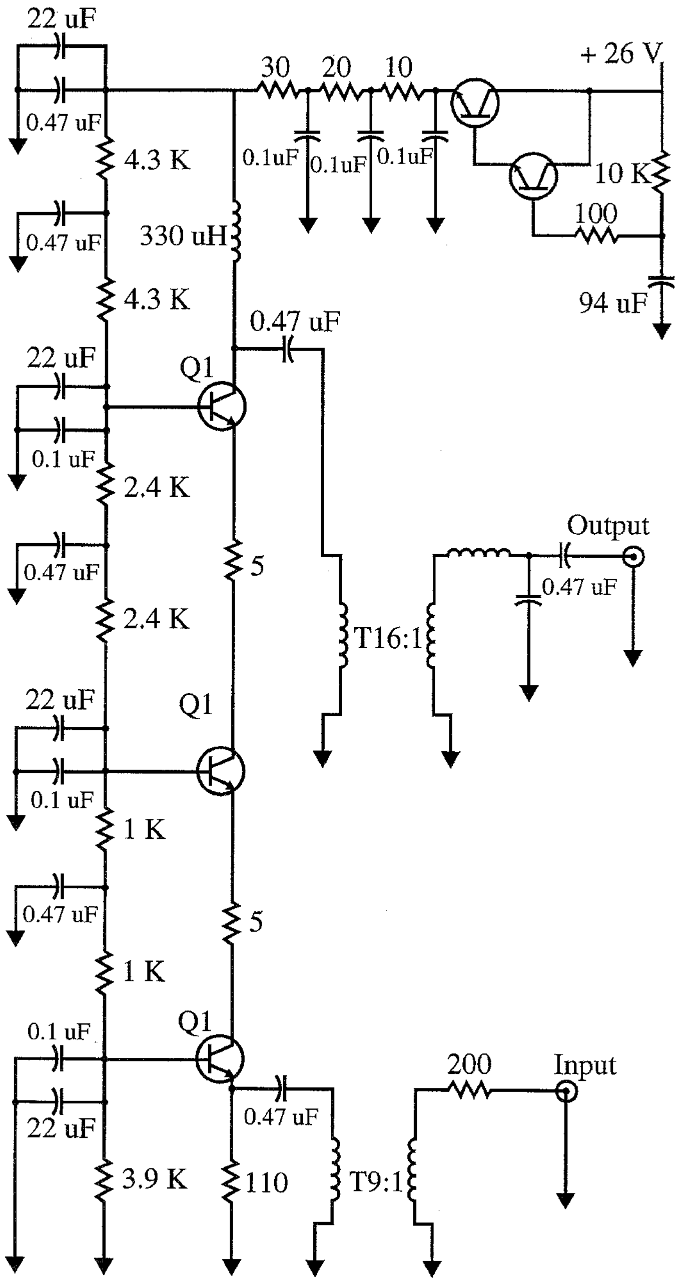

Fig. 12. A 5 MHz isolation amplifier.

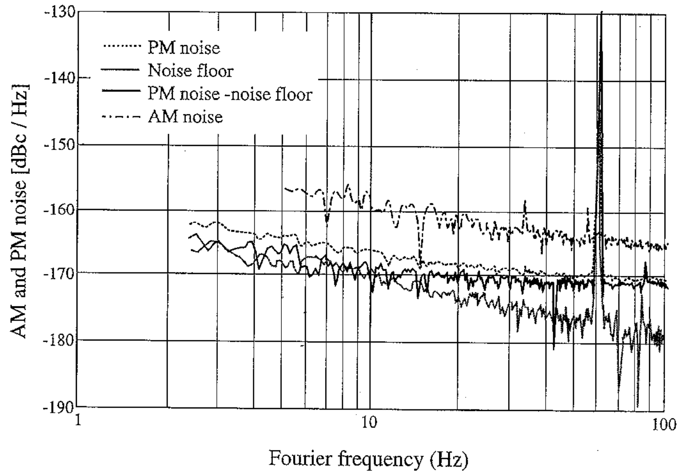

Fig. 13. PM and AM noise for an isolation amplifier at 5 MHz.

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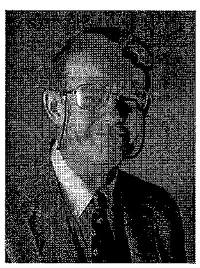
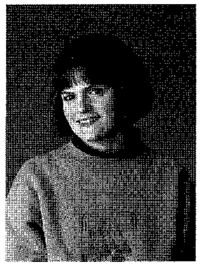
gain) with PM noise of $\mathcal{L}(10 \text{ Hz}) \leq -162$ dBc/Hz. AM and PM noise in CC and CB configurations was also studied. Due to their small phase shift and small gain, these stages have very low 1/f AM and PM noise. Our 5 MHz CC amplifier had noise levels similar to the noise floor of the measurement systems [$\mathcal{L}(10 \text{ Hz}) \leq -169$ dBc/Hz, $1/2S_a(10 \text{ Hz}) \leq -162$ dBc/Hz]. In addition, our 5 MHz distribution amplifier (composed of 3 CB stages) had noise levels of $\mathcal{L}(10 \text{ Hz}) \leq -168$ dBc/Hz, and $1/2S_a(10 \text{ Hz}) \leq -160$ dBc/Hz.



Fred L. Walls (A'93–SM'94) was born in Portland, OR, on October 29, 1940. He received B.S., M.S., and Ph.D. degrees in physics from the University of Washington, Seattle, in 1962, 1964, and 1970, respectively. His Ph.D. dissertation was on the development of long-term storage and nondestructive detection techniques for electrons stored in Penning traps and the first measurements of the anomalous magnetic (g-2) moment of low energy electrons. From 1970 to 1973 he was a Postdoctoral Fellow at the Joint Institute for Laboratory Astrophysics in Boulder, CO. This work focused on developing techniques for long-term storage and nondestructive detection of fragile atomic ions stored in Penning traps for low energy collision studies. Since 1973 he has been a Staff Member of the Time and Frequency Division of the National Institute of Standards and Technology, formerly the National Bureau of Standards in Boulder. He is presently Leader of the Phase Noise Measurement Group and is engaged in research and development of ultra-stable clocks, crystal-controlled oscillators with improved short- and long-term stability, low-noise microwave oscillators, frequency synthesis from RF to infrared, low-noise frequency stability measurement systems, and accurate phase and amplitude noise metrology. He has published more than 110 scientific papers and holds 5 patents. He received the 1995 European "Time and Frequency" award from the Societe Francaise des Microtechniques et de Chromometrie "for outstanding work in ion storage physics, design and development of passive hydrogen masers, measurements of phase noise in passive resonators, very low noise electronics and phase noise metrology." He is the recipient of the 1995 IEEE Rabi award for "major contributions to the characterization of noise and other instabilities of local oscillators and their effects on atomic frequency standards". He has also received two silver medals from the US Department of Commerce for fundamental advances in high resolution spectroscopy and frequency standards, and the development of passive hydrogen masers. Dr. Walls is a member of the American Physical Society, a senior member of the IEEE, a member of the Technical Program Committee of the IEEE Frequency Control Symposium and also a member of the Scientific Committee of the European Time and Frequency Forum.



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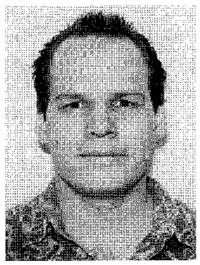

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