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Noise characteristics of thin ferromagnetic film devices

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NOISE CHARACTERISTICS OF THIN FERROMAGNETIC FILM DEVICES

by

Robert Lynn Samuels

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I. INTRODUCTION

The possible application of thin single-domain ferromagnetic film inductors as time varying parametric elements in electronic devices has been proposed by Head (1). Of the proposed devices, the balanced modulator, the upconverter and the parametric amplifier exhibit the possibility of power gain for small signals. Read and Pohm (2) have demonstrated the possibility of signal gain for the parametric amplifier. The purpose of this investigation was to study the noise characteristics of these three devices.

A. The Noise Problem

The smallest signal that can be interpreted by an electronic detection system is determined by three factors. One of these factors is the ratio of signal power to noise power required at the input of the final detector to achieve a satisfactory output. The signal-to-noise power ratio required varies greatly from one type of system to another. High fidelity reception of music, for instance, requires a much higher ratio than is necessary for barely understandable speech. Systems for which the form of the expected or desired signal is simpler than human speech or for which the characteristics of the expected signal are known, however complex that signal might be, can function satisfactorily with an even smaller signal-to-noise ratio. The second determining factor is the noise power that is added to or combined with the signal at the input of the electronic detection system. Every known sensor delivers noise power along with its output signal, for example, microphones deliver noise power as the result of gas molecule bombardment and the thermal noise of their loss resistances while
antennas deliver noise power resulting from cosmic radiation and the thermal noise of their loss resistances. The third factor is the excess noise power added to the signal by the amplifier(s) between the sensor and the detector. If there were a perfect or noiseless sensor, the system would still be limited in the smallest signal it could detect satisfactorily by the excess noise generated within the amplifying system. If, on the other hand, there were a perfect, noiseless amplifier, the detection system would be limited by the noise power delivered by the sensor to the input termination of the amplifier.

For a system having a given signal-to-noise ratio requirement at the final detector, the ultimate system sensitivity is determined by the excess noise power of the amplifier when the sensor's excess noise power is considered to be unavoidable. If the amplifier's excess noise power could be decreased, the required input signal power could be decreased. This would mean that the distance from transmitter to receiver could be increased for a radio communication system, or perhaps the size and gain of either the transmitting or the receiving antenna could be decreased. If the transmitter were a satellite or any portable transmitter, the required transmission power could be decreased.

If thin ferromagnetic film devices are to be useful as amplifying devices, then, it is necessary that they not be excessively noisy. The noisiness of amplifiers is characterized by a noise factor or noise figure to be defined in the next section. In later sections, a midband noise factor for the thin ferromagnetic film upconverter and parametric amplifier will be derived for the minimum noise possible. This minimum noise is the
thermal or Johnson noise generated by the resistances of the devices' cir-
cuitry. Other noise power(s) such as that produced by Barkhausen jumps,
the pump source, the bias source, and stray external magnetic fields will
be present in the devices. The amount of this excess noise must be de-
termined experimentally.

B. Noise Definitions

Noise in its broadest sense, can be defined as any undesired signal at
the output terminals of an amplifier. Van der Ziel (3) points out that
"spontaneous fluctuations" is a more descriptive term, but that the term
noise is commonly used as a result of the acoustical effects accompanying
spontaneous fluctuations in radio receivers.

In 1928, Johnson and Nyquist (4) showed noise to be a significant prob-
lem in the design of sensitive amplifiers. Nyquist's theorem states that
the noise in any circuit kept at a uniform temperature T can be described
by a noise electromotive force \( \left( \frac{e^2}{2} \right)^{1/2} \) in series with each resistance R
of the circuit such that for a small frequency interval \( df \) between the
frequencies \( f \) and \( f + df \)

\[
\overline{e^2} = 4kT p(f) df
\]

(1)

where \( \overline{e^2} \) is the mean square of the noise voltage, \( k \) is Boltzmann's constant,
\( T \) is the temperature in degrees Kelvin and \( p(f) \) is the Planck factor defined
by

\[
p(f) = \frac{hf/kt}{\left[ \exp \frac{hf}{kt} - 1 \right]}
\]

(2)
where $h$ is Planck's constant. At normal room temperature, $hf/kT \ll 1$ even at microwave frequencies. Then $p(f)$ is approximately equal to unity so that

$$
\overline{e^2} = 4kTRdf . \tag{3}
$$

A useful expression of Nyquist's theorem is the average available noise power $P_n$ from any resistor in the frequency interval $df$ given as

$$
P_n = \frac{\overline{e^2}}{4R} = kTdf . \tag{4}
$$

In amplifying devices there is a minimum level below which the noise cannot be reduced. This is the noise of the conductors of the device as expressed by Eq. 3. It is called Johnson noise or thermal noise. Even a "perfect" amplifier which added no noise to the input signal would be limited in the smallest detectable signal it could amplify by the thermal noise from its input termination. Amplifiers are classified as to their small signal capability by a noise factor or noise figure. These will be defined below.

Equation 4 is related to amplifiers by a noise bandwidth defined

$$
\Delta f = \frac{\int_{0}^{\infty} G(f)df}{G_c} . \tag{5}
$$

where $G(f)$ is the power gain of the amplifying system as a function of frequency, and $G_c$ is the midband power gain.

The noise power available at the output terminals of a perfect amplifier would be that which resulted from the available thermal noise power at the amplifier's input terminals. The noise power in a frequency interval $df$ would be
\[ d[N_p(f)] = G(f)kTdf \] (6)

where \( d[N_p(f)] \) is the available output noise power from a perfect amplifier at a frequency \( f \). The total noise output power \( N_{po} \) would then be the integral of Eq. 6, i.e.,

\[ N_{po} = \int_{0}^{\infty} G(f)kTdf = G_c kT \int_{0}^{\infty} \frac{G(f)df}{G_c} = G_c kTdf. \] (7)

In the noise analysis of a tuned amplifier it is conventional to refer all noise at the output to the input and to consider the amplifier to be a perfect, noiseless amplifier with an equivalent noise power generator at the input terminals. This equivalent noise power generator \( n_e \) has the magnitude \( n_e = N/G_c \) where \( N \) is the output noise power. An equivalent noise generator for the case of a perfect amplifier with an impedance matched input termination has, by Eq. 7, a noise power of

\[ n_p = \frac{N_{po}}{G_c} = kTdf. \] (8)

A definition of the noise factor of an amplifier is

\[ F = \frac{S_i/N_i}{S_o/N_o} \] (9)

where:

- \( S_i = \) Power of the input signal, excluding noise
- \( N_i = \) Power of the input noise
- \( S_o = \) Power of the output signal, excluding noise
- \( N_o = \) Power of the output noise.
Since the ratio $S_0/S_1$ is defined as the power gain $G_c$ for a tuned amplifier, $F$ can be written

$$F = \frac{N_o}{G_c N_i}.$$  \hspace{1cm} (10)

The output noise power $N_o$ is equal to $G_c N_i$ plus the excess noise power $N_{ex}$ added by the components of the amplifier. When this excess output-noise power is referred to the input, one can write the equivalent excess input noise power $n_{ex}$ as

$$n_{ex} = \frac{N_{ex}}{G_c}.$$  \hspace{1cm} (11)

Then one can write Eq. 10 as

$$F = \frac{G_c (N_i + n_{ex})}{G_c N_i} = 1 + \frac{n_{ex}}{N_i}.$$  \hspace{1cm} (12)

Equation 12 is a useful relationship for comparing amplifying systems because it involves only the input noise power and the equivalent excess noise power. When one knows the noise factor of an amplifier, he can compare its small-signal capability with almost any other amplifier. Parametric amplifiers are a notable exception, however.

It is common practice to state the noise factor $F$ in terms of a noise figure $NF$ as

$$NF = 10 \log_{10} F$$ \hspace{1cm} (13)

where $NF$ is in decibels and $F$ is dimensionless.

C. The Sources of Noise in Thin Ferromagnetic Film Devices

The principle types of noise in thin ferromagnetic film devices are
the thermal or Johnson noise generated by the circuit elements and the Barkhausen noise generated by the discontinuous and irreversible domain wall motions that occur in materials having multiple magnetic domains.

Thermal noise can be caused by many different mechanisms, for example, shot effect, the random interaction between the free electrons and the phonons in a conductor, and black body radiation. It would be wrong to conclude that all the thermal noise is caused by any single effect even though a given one may predominate in a particular system. The classification of the sources of thermal noise is relatively unimportant; the concern is for the amount of noise generated, and this is expressed by Eq. 3.

Fanslow (5, pp. 3-7) gives an explanation for the Barkhausen noise which will be repeated here in the interest of completeness:

Barkhausen noise is associated with the discontinuous and irreversible domain boundary movements that occur when the direction of magnetization is varied in materials that have many magnetic domains. This may be illustrated by using the model of the domain structure of a single crystal of iron as shown in Fig. 1. In Fig. 1 the position of the domain boundaries with zero field is shown by the solid line. The domain structure has been formed to minimize the energy associated with the localized fields due to the effect of free poles induced in the specimen. Since the energy associated with free poles is relatively large, it is necessary that the component of magnetization normal to the domain boundaries be continuous as far as possible. For the minimum energy condition then there are seen to be no free poles on the 180° (vertical) boundary. The closure domains at the ends of the crystal also form in such a way as to minimize the energy due to free poles. Since the area of the closure domain is generally small compared with the major domains, the effects of the closure domains may be assumed to be negligible. A change in the magnetic field causes the movement of the 180° boundary across the crystal to position x. If the crystal is perfect, without any non-magnetic inclusions or other defects, the movement will be smooth and no Barkhausen noise will be generated. When the crystal is strained or has non-magnetic inclusions present at a boundary, the direction of the component of magnetization normal to the domain boundary is no longer continuous and free poles
Figure 1. Model of the domain structure in a single crystal of iron (5, p. 6)

(A) Position of $180^\circ$ (vertical) boundary in a field $H$ as shown. Solid line shows boundary for $H = 0$

(B) Retardation of a section of the boundary with the production of free poles
are produced. The presence of inclusions results in a retardation of the movement of sections of the domain wall, Fig. 1. The motion of the retarded sections of the boundary will be discontinuous and will proceed in jumps producing Barkhausen noise. The Barkhausen effect is small and it is not often noticed, but its presence would be apparent on a hysteresis curve if the curve were pictured in sufficient detail. On the curve, changes in magnetization as a function of field applied would proceed in a series of steps rather than smoothly as it is generally shown. Tebble (6) shows that these step changes in magnetization are dependent upon the position of the inclusions, the width of the inclusions, and the length of the section of the domain wall that is retarded. Biorci and Pescetti (7) say that the Barkhausen effect produces a pulse that may be represented by an exponential curve having a time constant of approximately $10^{-4}$ second. Their study utilized the random superposition of these pulses in determining that the spectral density of Barkhausen noise is constant up to 1 kc and that it decreases rapidly thereafter. Since the distribution of sizes of Barkhausen discontinuities in bulk and thin film magnetic materials is similar, as has been shown by Ford and Pugh (8), one would expect the same type of noise from each. Williams and Noble (9) indicate that Barkhausen noise is the limiting factor in the signal level that may be amplified by present magnetic amplifier methods.

In addition to the thermal and Barkhausen noise, there will be interfering output voltages as the result of the capacitive coupling between the film and the signal winding or windings and as the result of misalignment. These sources of noise are of particular interest in the operation of the balanced modulator since they both contribute output noise power at the pump frequency. Theoretically, any interfering signal at a single frequency can be removed. Practically, however, a large interfering signal having nearly the same frequency as a small desired signal can make detection of the small signal extremely difficult. One method for reducing the effect of this interfering signal is discussed in the balanced modulator section.

Thin films are also extremely sensitive to extraneous magnetic fields. This sensitivity was demonstrated by Fanslow (5, p. 55) and investigated
further in the experimental portion of this study.

Another source of noise in thin ferromagnetic film devices is the variation of the magnetization with temperature as discussed by Dekker (10). The problem of producing a perfect or nearly perfect single magnetic domain film is probably more pressing than the variation of the magnetization; however, the effect of this fluctuation will be discussed qualitatively in a later section after the characteristics of the devices have been established.
II. THIN FERROMAGNETIC FILM DEVICE MODELS

A. The Model Derived by Read

The model derived by Read\(^1\) for thin ferromagnetic film devices is shown in Fig. 2. The constants are:

\[
\begin{align*}
\mu_0 &= \text{the permeability of free space} = 4\pi \times 10^7 \text{ henries per meter} \\
\delta &= \text{the phenomenological damping constant} \approx 0.02 \\
k_s &= \text{the ratio of magnetic field intensity in the film to the current in the signal winding} \\
\gamma &= 221 \times 10^3 \text{ radians per second per ampere turn per meter} \\
M &= \text{the magnitude of the saturation magnetization in ferromagnetic materials} \approx 1 \text{ weber per square meter} \\
\lambda_{ps} &= \text{the maximum flux linkage of the signal winding due to the magnetization} \\
H_K &= \text{the anisotropy constant of the film} \approx 220 \text{ ampere turns per meter} \\
H_b &= \text{the bias magnetic field intensity} \\
H_p &= \text{the peak pumping magnetic field intensity} \\
L_a &= \text{the air inductance of the signal winding.}
\end{align*}
\]

The small signal resonant frequency of this model with \(H_p = 0\) yields the Kittel ferromagnetic resonant frequency

\[
f_o = \frac{1}{2\pi} \left( \frac{1}{L_0 C_0} \right)^{1/2} = \left[ \frac{\gamma M (H_K + H_b)}{4\pi^2 \mu_0 (1 + \delta^2)} \right]^{1/2}.
\]

For the 80-20 nickel-iron alloy films under consideration, this resonant

---

Figure 2. The model derived by Read for the thin ferromagnetic film devices
\[ C_0 = \frac{\mu_0 (1 + \delta^2)}{K_s \gamma^2 M \lambda_{ps}} \]
\[ R_0 = \frac{K_s \gamma \lambda_{ps}}{\delta} \]
\[ L_0 = \frac{A_0 K_s \lambda_{ps}}{2 (H_K + H_b)} \]
\[ \alpha = \frac{H_p}{H_K + H_b} \]
\[ A_n = \frac{2}{\sqrt{1 + \alpha^2}} \left[ \frac{1 - \sqrt{1 - \alpha^2}}{\alpha} \right]^n ; \quad n = 0, 1, 2, 3, \ldots \]
\[ L(t) = L_0 \sum_{n=1}^{\infty} (-1)^n A_n \cos n \omega_p t \]
frequency is approximately 620 megacycles per second for the case $H_k = H_b = 220$ ampere turns per meter.

The films used in this study were electrodeposited on 5 mil diameter beryllium-copper wire with their easy direction circumferential to the axis of the wire and their hard direction parallel to the axis of the wire. This configuration has the advantage of presenting a closed flux path to the magnetization of the film, thereby decreasing the probability of forming closure domains caused by the demagnetizing fields present in flat thin films. However, this process of deposition increases the probability of nonmagnetic substances being left within the body of the film. These inclusions can be a source of the Barkhausen noise common to magnetic devices. This effect was discussed in an earlier section.

For the frequencies investigated in this thesis, the capacitance $C_0$ in Fig. 2 is negligibly small (approximately $10^{-15}$ farads for the estimated coefficients) and will be neglected. The remaining elements can be converted from the parallel arrangement shown to an approximately equivalent series impedance by assuming the pumping parameter $\alpha$ to be small and converting $R_0$ and $L_0$ to their equivalent series values. The equivalent series resistances and inductances are

$$R_e = \frac{R_0 \omega^2 L_0}{R_0^2 + \omega^2 L_0^2}$$

and

$$L_e = \frac{R_0 L_0}{R_0^2 + \omega^2 L_0^2}$$

1The permalloy coated wire was donated by Dr. T. N. Long of The Bell Telephone Laboratories.
where \( \omega \) is any angular velocity of interest. Now, if the inductor is to be an efficient amplifier, \( R_0 \gg \omega L_0 \). Then \( R_e \) and \( L_e \) become

\[
R_e \approx \frac{\omega^2 L_0}{R_0}
\]  
\[ (17) \]

and

\[
L_e \approx L_0 \cdot
\]  
\[ (18) \]

The thin ferromagnetic film balanced modulator, upconverter and parametric amplifier are very similar in their operation. In the next section, the principle of superposition will be used to divide the time invariant elements \( L_a, R_e \) and \( L_e \) into two branches to yield a conceptual two-port model for the purpose of determining the impedances and thermal noise characteristics of these devices.

If in Fig. 2 a periodic signal current \( i_s(t) \) flows through the parametric inductance such that

\[
i_s(t) = \sum_{s=0}^{\infty} I_s \cos (\omega_s t + \theta_s)
\]  
\[ (19) \]

where Eq. 19 could represent either a timewise complex current or the Fourier series of such a current, the voltage induced will be the derivative with respect to time of the product \( L(t)i_s(t) \). Thus,

\[
e(t) = \frac{d[L(t)i_s(t)]}{dt} = -\frac{L_0}{2} \sum_{s=0}^{\infty} \sum_{n=1}^{\infty} (-1)^n I_s A_n [\frac{n\omega_p + \omega_s}{p_s}]
\]

\[
\sin(\frac{n\omega_p t + \omega_s t + \theta_s}{p_s}) + (\frac{n\omega_p - \omega_s}{p_s})\sin(\frac{n\omega_p t - \omega_s t - \theta_s}{p_s})].
\]  
\[ (20) \]
Figure 3. The model used in the derivations of the impedances, gain, and noise factors of the thin film devices considered.
Equation 20 shows that the voltage contains frequencies at \((nw_p + ws)\) for \(n = 1, 2, 3, \cdots\), and \(s = 0, 1, 2, 3, \cdots\). These voltages will give rise to currents at any frequency passed by the filters \(F_s\) and \(F_u\), and these currents in turn induce voltages, among others, with the frequencies of the original input current, yielding a feedback impedance. In the next section, a single frequency input signal current will be assumed and the resultant impedances determined, for these impedances must be known if one is to compute the effect of the thermal noise generated by the elements of the system.

B. The Noise Model

Equation 20 shows that a current having angular velocity \(w_s\) will interact with the time varying inductance \(L(t)\) to produce voltages with frequencies \((nf_p + f_s)\) for \(n = 1, 2, 3 \cdots\). It will be shown in later sections that the balanced modulator and the upconverter achieve gain by increasing the signal frequency to another higher frequency and that the parametric amplifier's optimum noise performance occurs for the case of the signal frequency less than half the pump frequency. For the purpose of analysis, it will be assumed that the signal frequency is less than half the pump frequency and that the significant term of Eq. 20 for power gain is the term for which \(n = 1\).

In Fig. 3:

\[
\begin{align*}
F_s &= \text{a filter which passes currents with frequencies near the signal frequency } f_s \\
F_u &= \text{a filter which passes currents with frequencies near either or both the upper frequencies } (f_p + f_s)
\end{align*}
\]
\( R_s \) = the loss resistance of the signal frequency circuit including the circuit element losses and the loss of the film at the signal frequency

\( X_s \) = the reactance of the signal frequency circuit including that of the filter and the film

\( R_g \) = the generator resistance

\( R_u \) = the loss resistance of the upper frequency circuit element losses and the loss of the film at the upper frequency of frequencies

\( X_u \) = the reactance of the upper frequency circuit including that of the filter and the film

\( R_L \) = the load resistance

\( L(t) \) = the parametric inductance as in Fig. 2

\( e_s(t) \) = the voltage induced by the film at the signal frequency

\( e_u(t) \) = the voltage induced by the film at the upper frequency or frequencies

\( i_s(t) = I_s \sin (\omega_s t + \theta_s) \)

and \( Z_s, Z_{in}, Z_u, \) and \( Z_d \) are the impedances looking in the directions indicated.

Although only the first harmonic of the parametric inductance is the significant term for the purpose of power gain, other harmonics of the parametric inductance will contribute to the total noise of the devices to be analyzed. Therefore, it is convenient to consider current \( i_1(t) \) (not shown in Fig. 3) flowing into \( L(t) \) and to consider the complete Fourier series for \( L(t) \) assuming only that the pumping parameter \( \alpha \) is sufficiently small that the series converges rapidly. Let \( L_0 A_n = L_n \) for \( n = 1, 2, 3, \ldots \), and let
\[ i_1(t) = I_1 \sin(\omega_1 t + \theta_1) \quad (21) \]

where \( \theta_1 \) is the phase of the current with respect to the time varying inductance. The time varying inductance is

\[ L(t) = \sum_{n=1}^{\infty} (-1)^n L_n \cos \frac{n\omega t}{p}. \quad (22) \]

The voltage induced will be

\[ e(t) = \frac{d[i_1(t)L(t)]}{dt} = \frac{I_1}{2} \sum_{n=1}^{\infty} (-1)^n L_n [\omega_{2n} \cos(\omega_{2n} t + \theta_1) \]
\[ -\omega_{2n} \cos(\omega_{2n} t - \theta_1)] \quad (23) \]

where \( \omega_{2n} = (n\omega - \omega_1) \), and \( \omega_{2n} = (n\omega + \omega_1) \). Current will then flow through some appropriate impedance for each frequency involved. Let

\[ Z_{ln}(\omega_{ln}) = |Z_{ln}| e^{j\theta_{ln}} \text{ and } Z_{2n}(\omega_{2n}) = |Z_{2n}| e^{j\theta_{2n}} \text{ for } n = 1, 2, 3, \ldots. \]

The total current can then be written

\[ i_1'(t) = \frac{I_1}{2} \sum_{n=1}^{\infty} (-1)^n L_n \left[ -\frac{\omega_{2n} \cos(\omega_{2n} t + \theta_1 - \theta_{2n})}{|Z_{2n}|} \right. \]
\[ + \frac{\omega_{ln} \cos(\omega_{ln} t - \theta_1 - \theta_{ln})}{|Z_{ln}|} \right] \quad (24) \]

where the sign of the current has been reversed to indicate that the current \( i_1'(t) \) has been assumed positive in the same direction as \( i_1(t) \).

This current will induce a voltage
\[ e_i(t) = \frac{d[I(t)i_i(t)]}{dt} \]

\[ = \frac{1}{4} \sum_{n=1}^{\infty} \sum_{m=1}^{\infty} (-1)^{m+n} L_m L_n \left[ \frac{\omega_{2n}(\omega_p + \omega_{2n})\sin(\omega_{p} t + \omega_{2n} t + \theta_{2n})}{|Z_{2n}|} \right. \]

\[ + \frac{\omega_{2n}(\omega_p - \omega_{2n})\sin(\omega_{p} t - \omega_{2n} t + \theta_{2n})}{|Z_{2n}|} \]

\[ \left. - \frac{\omega_{ln}(\omega_p + \omega_{ln})\sin(\omega_{p} t + \omega_{ln} t + \theta_{ln})}{|Z_{ln}|} \right] - \frac{\omega_{ln}(\omega_p - \omega_{ln})\sin(\omega_{p} t - \omega_{ln} t + \theta_{ln})}{|Z_{ln}|}. \]  

(25)

The terms of Eq. 25 that are of interest here are those for which \( n = m = 1 \) and for which \( \omega_i = \omega_s \), \( \omega_i = \omega_p - \omega_s \), or \( \omega_i = \omega_p + \omega_s \). When one lets \( \omega_i = \omega_s \) and compares the voltage \( e_i(t) \) with the input current as expressed by Eq. 21, he finds that

\[ Z_{in} = \frac{\omega L_i^2}{4} \left( \frac{\omega_{21}}{Z_{21}} - \frac{\omega_{11}}{Z_{11}} \right) \]  

(26)

where \( Z_{11}^* \) is the complex conjugate of \( Z_{11} \).

When one lets \( \omega_i = \omega_p + \omega_s \) and compares the voltage \( e_i(t) \) with the input current \( i_i(t) \), he finds

\[ Z_{d2} = \frac{\omega_s \omega L_i^2}{4 Z_s} \]  

(27)

for the case \(|Z_{11}| \) very large.

When one lets \( \omega_i = \omega_p - \omega_s \) and again compares \( e_i(t) \) with the input current \( i_i(t) \), he finds

\[ Z_{dl} = -\frac{\omega_s \omega L_i^2}{4 Z_s^*} \]  

(28)
for the case $|Z_{21}|$ very large.

These impedances describe the necessary first order impedances for the analysis of the thermal noise characteristics of the upconverter and the parametric amplifier. The voltage $e(t)$ as expressed by Eq. 23 for the case $n = 1$ is

$$e(t) = e_u(t) = \frac{I_1 I_s}{2} \left[ w_{21}\cos(w_{21}t + \theta_s) - w_{11}\cos(w_{11}t - \theta_s) \right]. \quad (29)$$

Equation 29 defines the voltage induced for the balanced modulator and the upconverter.

The voltage $e_i'(t)$ for the case $m = n = 1$, considering only the terms with an angular velocity $\omega_i = \omega_s$, is

$$e_i'(t) = e_s(t) = \frac{I_s L_2^2}{4} \left[ \frac{w_{21}\sin(w_s t + \theta_s - \theta_{21})}{|Z_{21}|} - \frac{w_{11}\sin(w_s t + \theta_s + \theta_{11})}{|Z_{11}|} \right]. \quad (30)$$

This voltage is the "feedback" voltage for the balanced modulator, the upconverter and the parametric amplifier.
III. THE DEVICES

A. The Balanced Modulator

The thin ferromagnetic film balanced modulator has been investigated by Read (1), Fanslow (5), and this writer (11). Its noise characteristics were investigated by Fanslow. When the filter $F_u$ of Fig. 3 passes both the upper and lower side bands, the input impedance is as expressed by Eq. 26

$$Z_{\text{in}} = \frac{L_1 l^2 s}{4} \left( \frac{w_{21}}{Z_{21}} - \frac{w_{11}^*}{Z_{11}} \right). \quad (31)$$

If it is assumed that the output filter is tuned to $w_p$, then $Z_{11}^* \approx Z_{21}$ and

$$Z_{\text{in}} = \frac{w_{21}^2 Z_{21}^*}{2 |Z_{21}|^2} \quad (32)$$

where $|Z_{21}|$ is the magnitude of $Z_{21}$.

At first thought, this may appear to present no problem, but when one considers that the balanced modulator's primary use is modulating or up-converting small signal frequencies and that thin ferromagnetic films do not give a very large parametric inductance coefficient $L_1$ per turn of signal winding, it becomes clear that one must use a coil with many turns to couple to the film if he is to avoid a vanishingly small input or output impedance.

The resistance of the coupling winding(s), however, will dissipate much of the signal power if the loss resistance is a significant fraction of the real part of the input impedance. It would be possible to use a material which would become a superconductor at low temperatures to avoid
the losses in the signal winding; however, it has been found by Mitchell (12) that the rest direction of thin ferromagnetic films changes with a change of temperature. In a previous investigation (11), this writer has demonstrated that misalignment causes feedthrough of the pump frequency. Another cause of pump frequency feedthrough is the capacitive coupling between the film and the input and output winding(s). This capacitive coupling can be reduced by shielding or by a cancellation scheme such as that used by Fanslow, but these methods cannot be more than partially effective. Assuming the pump frequency feedthrough is not sufficient to saturate the amplifier following the modulator, a possible solution would be to use a coherent or synchronous detection technique as described by Costas (13). The system required here would be much less elaborate than required for a remote receiver, but would have the same advantage. Briefly, the system proposed here would require that the balanced modulator's output voltage \( e_u(t) \) as given by Eq. 29 be detected with a phase-locked product detector with a voltage gain of

\[
g_d(t) = G_d \cos \omega_p t. \tag{33}
\]

The product of Eqs. 29 and 33 is

\[
v_o(t)=g_d(t)e_u(t)=K[w_{21}\cos(\omega_{21} t+\omega_p t+\theta_s)+w_{21}\cos(\omega_{21} t-\omega_p t+\theta_s)]
- w_{11}\cos(\omega_{11} t+\omega_p t-\theta_s)w_{11}\cos(\omega_{11} t-\omega_p t-\theta_s)
= 2K[w_{11}\cos(\omega_p t+\theta_s)] + \text{voltages at frequencies near } 2f_p \tag{34}
\]

where \( K \) is a proportionality constant. In this system, a simple RC coupling at the output of the detector would effectively block the pump
frequency feedthrough, since the product of \( g_d(t) \) with the pump frequency voltage would yield frequencies at zero and \( 2f_p \) but would also make any zero-frequency amplification impossible.

This qualitative discussion has assumed that thin ferromagnetic films will function while supercooled. The effect of small temperature changes was investigated in the experimental part of this project and will be discussed in the experimental results section.

The balanced modulator does present the possibility of very high power gain. It is a combination of the upconverter and the parametric amplifier in that both the upper and lower sideband currents are allowed to flow. As will be shown, the upconverter has a theoretical maximum midband gain of \( G_c = \frac{\omega_2}{\omega_s} \), while the parametric amplifier is a potentially unstable device capable of operating as an oscillator. The balanced modulator should then show a combination of the gains of these two devices. Transducer gain is defined as the ratio of the actual output power to the available input power. If the film and winding were lossless and the output circuit sufficiently broadband such that \( Z_{21} = R_L \), then

\[
R_{in} = \frac{\omega_s^2 L_1^2}{2R_L},
\]

and the available input power would be, assuming \( R_{in} = R_s \),

\[
P_{in} = \frac{I_s^2 R_{in}}{2} = \frac{I_s^2 \omega_s^2 L_1^2}{4R_L}.
\]

The output power would be, from Eq. 29,
The problems of film winding losses, capacitive coupling, and misalignment, however, are still formidable. Fanslow has given an analysis of the losses in the balanced modulator.

B. The Upconverter

If the device in Fig. 3 is a thin ferromagnetic film upconverter, the filter \( F_u \) passes currents with angular velocity \( \omega_{21} \) and stops those with angular velocity \( \omega_{11} \). The input impedance at the film inductance \( L(t) \) becomes

\[
Z_{in} = \frac{\omega \omega_{21} L}{4Z_{21}}
\]

where \( Z_{21} = R_u + R_L + jX_u \). Voltages will be generated at many frequencies, but the voltage of principle interest in the upper frequency circuit is

\[
e_u(t) = \frac{I_s \omega_{21} L}{2} \cos \left( \omega_{21} t + \theta_s \right).
\]

If the film and circuitry were lossless, and if the input and output circuits were tuned, the input impedance would be real, i.e.,

\[
R_{in} = \frac{\omega \omega_{21} L^2}{4R_L}.
\]

Maximum power gain would occur for \( R_{in} = R_g \). Transducer gain is again defined as the ratio of power output to the available power at the input.
The available input power would be
\[
\frac{I_s R_{in}^2}{2} = \frac{I_s^2 w^2 L_2}{8 R_L}
\]  \hspace{1cm} (42)

and the output power would be
\[
\frac{e_u^{(\text{max})}}{2 R_L} = \frac{I_s^2 w_{21}^2 L_2}{8 R_L}
\]  \hspace{1cm} (43)

The theoretical midband gain would then be
\[
G_c = \frac{w_{21}}{w_s}
\]  \hspace{1cm} (44)

This corresponds to the theoretical gain predicted by the Manley-Rowe Relations.

It is interesting to note that this result is true regardless of the magnitude of \( R_L \). This presents the possibility of avoiding the problem of the vanishingly small input impedance by cascading two upconverters such that the load resistance of the first upconverter would be the small input impedance of the second upconverter. Thus, the input impedance of the first upconverter and the load impedance of the second upconverter could be of reasonable magnitude.

It can be shown by this same reasoning that the lossless power "gain" for a signal downconverted from the angular velocity \( w_{21} \) to the angular velocity \( w_s \) would be
\[
\"G_c\" = \frac{w_s}{w_{21}}
\]  \hspace{1cm} (45)

This relationship will be useful for referring the noise output of the upconverter to its input terminals to establish a theoretical midband noise
factor for the upconverter.

The input impedance expressed by Eq. 39 shows that the upconverter is potentially a wideband device, since the impedance inverting property of the time varying inductance indicates the possibility of obtaining a conjugate impedance match for maximum power transfer. This conjugate impedance match will occur when \( Z_{in} = Z_s^* \), where \( Z_s \) is indicated in Fig. 3. Then

\[
Z_s^* Z_{21} = \frac{w_s w_{21} L_1}{4},
\]

and, inasmuch as the right side of Eq. 46 is real, one can write \( Z_s = |Z_s| e^{j\theta_s} \) and \( Z_{21} = |Z_{21}| e^{j\theta_{21}} \) to obtain

\[
Z_s^* Z_{21} = |Z_s| \cdot |Z_{21}| \cdot e^{j(\theta_{21} - \theta_s)} = \frac{w_s w_{21} L_1^2}{4}.
\]

Maximum power transfer and power gain will then occur for \( \theta_{21} = \theta_s \).

The filter required to make \( \theta_{21} = \theta_s \) over the passband is restricted to have impedance zeros at \( \omega_s \) and \( \omega_{21} \), and it must have the series inductance \( L_o + L_a \) as the first element. In a manner similar to that of Kuh (14) for upconverters employing capacitive reactances, this can be accomplished by setting

\[
|Z_s| = |Z_{21}| = \frac{(w_s w_{21})^{1/2} L_1}{2}
\]

and using a low-pass to multiband-pass transformation,

\[
p = \frac{(s^2 + \omega_s^2)(s^2 + \omega_{21}^2)}{s(s^2 + \omega_s^2)}
\]

with

\[
\omega_s^2 = \frac{w_s^2 + w_{21}^2}{2}
\]

\[
(50)
\]
where \( p \) is the low-pass variable and \( s \) is the band-pass variable. Such a filter is shown in Fig. 4 with

\[
L = L_a + L_o
\]

\[
L_a = \frac{L(w_s^2 + w_{21}^2)^2}{(w_s^2 - w_{21}^2)^2}
\]

\[
C_A = \frac{2(w_s^2 + w_{21}^2)^2}{21w_s^2w_{21}(w_s^2 + w_{21}^2)}
\]

\[
C_B = \frac{2}{L(w_s^2 + w_{21}^2)}
\]

The cutoff angular velocity for the low-pass filter would be \( \omega_c = \frac{R_l}{(L_a + L_o)} \). Then, by conservation of bandwidth, the high frequency bandwidths will be \( B = \frac{\omega_c}{2} \), or the fractional bandwidth

\[
\frac{B}{\omega_s} = \frac{1}{2} \left( \frac{\omega_{21}}{\omega_s} \right)^{1/2} \frac{L_1}{L_a + L_o} \cdot (52)
\]

The theoretical voltage gain-fractional bandwidth product is then

\[
(G_c)^{1/2} \frac{B}{\omega_s} = \frac{\omega_{21}}{2\omega_s} \frac{L_1}{L_a + L_o} \cdot (53)
\]

If the lower sideband \( \omega_{11} \) is not stopped sufficiently by the impedance pole of the filter in Fig. 4, however, the networks denoted by \( F_s \) and \( F_u \) in Fig. 3 must stop current at that frequency.

Equation 12 will be used to obtain a midband thermal noise factor for the upconverter. The excess noise generator in the signal circuit is \( R_s \), and the thermal noise voltage generated is
Figure 4. Illustration of a filter for attaining the broadband capability of the thin ferromagnetic film upconverter where the elements \( L, L_A, C_A \) and \( C_B \) have the values stated in Eq. 51.
\[ E_{ns} = \left( \frac{2}{e_{ns}} \right)^{1/2} = (4kTR_s \Delta f)^{1/2} \]  

(54)

Other noise, such as Barkhausen noise, will be generated by the "resistance" \( R_{in} \), but that noise will be omitted at this point. \( R_{in} \) is generated reactively and is theoretically noiseless. The excess thermally generated noise current in the signal circuit is then

\[ I_{ns} = \frac{(4kTR_s \Delta f)^{1/2}}{R_g + R_s + R_{in}} \]  

(55)

and the noise power delivered into the upper frequency circuit will be

\[ \frac{w_1}{w_2} I_{ns}^2 R_{in} = \frac{w_1}{w_2} \frac{4kTR_s R_{in} \Delta f}{(R_g + R_s + R_{in})^2} \]  

(56)

Of this power, only the fraction \( R_L/(R_u + R_L) \) will be delivered to the load. Denote this power as \( P_{ns} \).

Additional noise power will be generated in the upper frequency circuit. By convention, the noise power generated in the load resistance is not attributed to an amplifying device. This convention will be followed in this section; however, in the discussion of the parametric amplifier, the noise generated in the load resistance will be included. For the up-converter, then, the noise voltage to be considered in the upper frequency circuit is

\[ E_{nu} = (4kTR_u \Delta f)^{1/2} \]  

(57)

The impedance \( Z_{u2} \) expressed by Eq. 27 is included in the current path for this noise voltage. At midband this becomes
The noise power delivered to the load resistance from the noise voltage of Eq. 57 is then

\[ P_{nu} = \frac{4kT R_u \Delta f}{R_d + R_u + R_L}^2, \]  

and the total noise power at the load resulting from the excess thermal noise generated in the upconverter is the sum of the powers \( P_{ns} + P_{nu} \):

\[ P_{nt} = \frac{4kT R_u \Delta f}{(R_d + R_u + R_L)^2} \left[ \frac{2l_s}{R_g + R_u + R_L} \right] \frac{4kT R_u R_R \Delta f}{(R_d + R_u + R_L)^2} \]  

To complete the noise factor calculation, the output noise power must be referred to the input terminals by dividing by the midband gain. When one accounts for all the losses, he can show that the midband signal gain is

\[ G_s = \frac{2l_s}{R_g + R_u + R_L} \left[ \frac{R_s}{R_g + R_u + R_L} \right] \]  

This expression reduces to the theoretical gain for the case \( R_u = R_g = 0 \) and \( R_{in} = R_g \). The excess output noise power \( P_{nt} \) referred to the input terminals is

\[ P_{ni} = \frac{kT}{R_g} \left[ \frac{R_s}{R_g + R_u + R_L} \right] \left[ \frac{R_s}{R_g + R_u + R_L} \right] \Delta f \]  

Thus, the noise factor for the upconverter is

\[ F = 1 + \frac{R_s}{R_g} \left[ \frac{R_u + R_L}{R_u + R_L} \right] \left[ \frac{R_s}{R_g + R_u + R_L} \right] \left[ \frac{R_s}{R_g + R_u + R_L} \right] \]
In the calculations to this point the noise temperatures of the resistors have been assumed to be the same, but the upconverter's gain could be increased and its noise factor decreased by cooling the device. If one defines:

\[ R_{fs} = \text{the film resistance at angular velocity } \omega_s \]
\[ R_{cs} = \text{the circuit resistance at angular velocity } \omega_s \]
\[ R_{ws} = \text{the coupling winding resistance at angular velocity } \omega_s \]
\[ R_{fu} = \text{the film resistance at angular velocity } \omega_{21} \]
\[ R_{cu} = \text{the circuit resistance at angular velocity } \omega_{21} \]
\[ R_{wu} = \text{the coupling winding resistance at angular velocity } \omega_{21} \]

and identifies the absolute temperature T of each resistance by the subscript of that resistance, he can write Eq. 63 as

\[
F = 1 + \frac{R_{fs} T_{fs} + R_{cs} T_{cs} + R_{ws} T_{ws}}{R_T \frac{T_{fs}}{T}} + \frac{w_s (R_u + R_s + R_{in})^2 (R_u + R_s) (R_{fu} T_{fu} + R_{cu} T_{cu} + R_{wu} T_{wu})}{w_{21} (R_d + R_u + R_L)^2 R_{in} R_T}. \quad (64)
\]

As one can see from this expression, cooling the input circuitry and the film with its winding(s) could reduce the noise factor to nearly the ideal factor of unity. Using a material which would become a superconductor for coupling to the film would increase the gain expressed by Eq. 61 to nearly the theoretical gain, \( \omega_{21}/\omega_s \). The somewhat contradictory experimental results for moderate cooling indicate that a more complete study in this phase of thin film devices is needed.
C. The Parametric Amplifier

The thin ferromagnetic film parametric amplifier has a potential for infinite gain and thus for operation as an oscillator. If the filter $F_u$ in Fig. 3 passes only those currents with angular velocity $\omega_1$, the magnitude of $Z_{21}$ becomes very large. The impedance $Z_{in}$ in Eq. 26 then becomes

$$Z_{in} = -\frac{\omega_s \omega_{11} L_1^2}{4Z_{11}^*} = -\frac{\omega_s \omega_{11} L_1^2 Z_{11}}{4|Z_{11}|^2}.$$  \hspace{1cm} (65)

This apparent negative impedance can be made maximum by shorting out $R_L$ in Fig. 3. The negative real part of this impedance can be utilized by placing a load resistance $R_p$ in the signal circuit.

At midband, when $Z_{11} = R_u$,

$$R_{in} = -\frac{\omega_s \omega_{11} L_1^2}{4R_u} \triangleq -R.$$  \hspace{1cm} (66)

Then the resistance seen from the terminals of the generator will be $R_p + R_s - R$. The threshold of gain for the special case $R_p = R_s$ occurs when $R_g = R$, or

$$R_s = \frac{\omega_s \omega_{11} L_1^2}{4R_u}.$$  \hspace{1cm} (67)

When one recalls the approximate expression implied by Eq. 17 for the portion of $R_s$ and $R_u$ attributed to the film itself, he can show that for the film to overcome its own losses it must be pumped sufficiently hard that

$$L_1 = \frac{2(\omega_s \omega_{11})^{1/2} L_0^2}{R_0}.$$  \hspace{1cm} (68)
This shows the importance of maintaining a high bias field to make $L_o$ small.

Parametric amplifiers are reputed to be low noise devices for two reasons. First, the gain is generated reactively and, therefore, relatively noiselessly. Second, the convention of excluding the noise power generated in the load resistance from computed noise factors and of subtracting that portion generated in the load from measured noise factors yields unrealistically low values for low frequency parametric amplifiers operated without circulators. In low frequency parametric amplifiers the noise power generated by the load resistance is amplified and returned to the load as an output signal. For the circuit of Fig. 3, if the film were considered to be lossless, and if $L(t)$ could generate a perfect negative resistance, the noise factor would be

$$F' = 1 + \frac{R_p}{R_g}$$

(69)

when the noise power generated by the load $R_p$ is considered. In this thesis, the noise power generated by the load will be included for the parametric amplifier. If it should become possible to operate thin film amplifying devices with gain at frequencies where circulators are available, it would be possible to subtract the contribution of the load.

For the parametric amplifier, $R_u$ is the resistance of the film and circuitry at the angular velocity $\omega_{ul}$, while the resistance in the signal frequency circuit is the sum of $R_g$, $R_p$, the circuit losses of the signal circuit and the film loss resistance at the angular velocity $\omega_s$.

The impedance $Z_d$ in Fig. 3 for the parametric amplifier is that
expressed by Eq. 28

$$Z_{dl} = -\frac{\omega L}{4Z_s^*},$$

(70)

where $Z_s^* = R_g + R_p + R_s - jX_s$. At midband this becomes

$$R_d = \frac{-\omega L}{4(R_g + R_p + R_s)}.$$  

(71)

Equation 71 shows that the noise currents in the idler circuit see a negative resistance in their path and are therefore amplified, but, since the denominator contains the relatively large resistances of the generator and the load, the magnitude of $R_d$ will be relatively small.

The general energy relations which govern nonlinear reactor modulators have been derived by Manley and Rowe (15). They show that the average powers at the different frequencies in these reactors are related by two independent equations. Salzberg (16) has developed the Manley-Rowe relations for the special case of three frequencies: the signal frequency, the pump frequency and the sum or difference of the signal and pump frequencies. When the load is responsive to the difference frequency only, these relations give for the present system

$$-\frac{P_{11}}{P_{11}} = \frac{P}{P} = \frac{P_{s}}{-f_s}.$$  

(72)

where the negative sign of a power $P$ indicates power absorption. One sees from Eq. 72 that

$$P_s = \frac{\omega}{\omega_{11}} P_{11}.$$  

(73)
Equation 73 has two interpretations of interest here. If the negative resistance in the idler circuit amplifies noise power, as it must if the resistances in the idler circuit generate noise power, then $\omega_s/\omega_{ll}$ times that power must be delivered to the signal circuit. If the idler circuit elements and the film could be supercooled, the noise factor would be reduced. The effect of cooling will be discussed later in this section.

Equation 73 also shows that when the negative resistance in the signal circuit amplifies signal power, then $\omega_{ll}/\omega_s$ times that power must be generated, and dissipated, in the idler circuit. This excess power might at first thought seem a harmless dividend, but it must be dissipated in some loss element or elements, and, inevitably, two of the loss elements must be the film and the coupling winding. The resultant temperature increase of the film will be small for small signals and narrow bandwidth (the input noise is amplified also), but the increase could be significant for a high gain parametric amplifier in a noisy electronic environment.

The excess noise generating resistances in the signal circuit are $R_s$ and $R_p$. $R_u$ is the noise generating resistance in the idler circuit. The negative resistance $R_d$ amplifies the noise power generated by the resistance $R_u$ to give a noise power in the idler circuit of

$$P_{ni} = \frac{4kT|R_d|R_u\Delta f}{(R_u + R_d)^2}$$  \hspace{1cm} (74)

where $|R_d|$ is the magnitude of $R_d$.

This power, multiplied by the factor $\omega_s/\omega_{ll}$, will be delivered to the signal circuit. Of that total power, the fraction $R_p/(R_g + R_p + R_s)$ will be delivered to the load resistance. The noise contribution to the load
The noise power delivered to the load resistance by the excess noise generators in the signal circuit is

\[ P_{ns} = \frac{4kT(R_p + R_s)R_Af}{(R_g + R_p + R_s - R)^2} \]  \hspace{1cm} (76)

To compute the amplifier's noise factor the total output excess noise power obtained by adding Eqs. 75 and 76 must be transferred to the input terminals by dividing by the parametric amplifier's gain

\[ G_p = \frac{\text{Actual Output Power}}{\text{Available Input Power}} = \frac{4R_g^2R_p}{(R_g + R_p + R_s - R)^2} \]  \hspace{1cm} (77)

This gives the thin film parametric amplifier a noise factor of

\[ F = 1 + \frac{R_p + R_s}{R_g} + \frac{\omega_s R_p}{\omega_{ll} R_g} \frac{R_d}{(R + R_p + R_s - R)^2} \]  \hspace{1cm} (78)

The third term of Eq. 78 reduces to a much more convenient form, but will be retained in this form to illustrate the effect reducing the temperature may have on the noise factor. If one defines:

- \( R_{fs} \) = the film resistance at angular velocity \( \omega_s \)
- \( R_{cs} \) = the circuit resistance at angular velocity \( \omega_s \)
- \( R_{ws} \) = the coupling winding resistance at angular velocity \( \omega_s \)
- \( R_{fi} \) = the film resistance at angular velocity \( \omega_i \)
\[ R_{ci} = \text{the circuit resistance at angular velocity } \omega_i \]
\[ R_{wi} = \text{the coupling winding resistance at angular velocity } \omega_i \]
\[ R_i = \text{the resistance one might place in the idler circuit to increase the bandwidth} \]
\[ A = \frac{|R_d|}{(R_p + R_s + R)^2/(R_p + R_s + R)(R_p + R_d)^2} \]

and identifies the absolute temperature \( T \) of each resistance by the subscript of that resistance, he can write the equivalent of Eq. 78 as

\[ F = 1 + \frac{R_{fs}T_{fs} + R_{cs}T_{cs} + R_{ws}T_{ws} + R_{p}T_{p}}{R_{g}T_{g}} \]
\[ + \frac{\omega_s A(R_iT_i + R_{ci}T_{ci} + R_{wi}T_{wi} + R_{li}T_{li})}{R_{ll}R_{g}T_{g}} \]  

(79)

As one can see from this expression, a circulator would reduce the noise factor considerably by removing the \( R_{p}T_{p} \) term from Eq. 79, but cooling the input circuitry would have little effect unless the film and its coupling winding could be cooled also. The electrodeposited films tested exhibited interesting but inconclusive changes with temperature. Experimentally, the noise factor decreased with a small degree of cooling for some films and increased for others. These tests were made on films deposited on beryllium-copper wire, and the changes observed may have been more the result of strain than temperature change.

When one recalls the definitions of \( R \) and \( R_{d} \) from Eq. 66 and Eq. 71, respectively, he can reduce Eq. 78 to the more simple form

\[ F = 1 + \frac{R_p + R_s}{R_g} + \frac{\omega_s R}{\omega_{ll}R_g} \]  

(80)

by assuming that all the elements have the same temperature. When the
noise factor is expressed in this form, it becomes clear that the idler frequency must be higher than the signal frequency to obtain a low noise device, particularly if the magnitude of the negative resistance generated is as large as or larger than $R_g$. 
IV. ADDITIONAL NOISE CONSIDERATIONS

As was mentioned in the discussion of the sources of noise in thin ferromagnetic film devices, the magnetization of thin films is dependent on temperature, and the fluctuations will be a source of additional thermal noise. There are several frequencies at which the variations could occur to generate noise power output in the balanced modulator, the upconverter and the parametric amplifier. The distribution of the random fluctuations of the magnetization is not known, but it can be written as equivalent noise voltages whose values cannot be stated until the physical process of their generation is found. Also, the fluctuations of the magnetization will cause fluctuations in the magnitude of the parametric inductance coefficient $L_1$.

This discussion will be limited to the effects the fluctuations in the magnetization will have on the parametric amplifier.

First, there will be a fluctuation in the magnetization that will induce a voltage at the signal frequency. This voltage will be denoted as $\left(\frac{e_s}{2}\right)^{1/2}$. By analogy with the effect of the thermal noise voltages generated in the loss resistances of the signal circuit, the output noise power is

$$N_{OS} = \frac{e_s^2 R_p}{(R_g + R_p + R_s - R)^2}. \quad (81)$$

This adds to the noise factor of Eq. 80 a term

$$F_S = \frac{e_s^2}{4kT R\Delta f}. \quad (82)$$
Also, there will be a fluctuation in the magnetization which will produce a voltage at the idler frequency. This voltage will be denoted as \((\bar{e}_i^2)^{1/2}\) and can be treated by analogy with the thermal noise voltages generated by the loss resistances in the idler circuit. The contribution of this fluctuation to the noise factor of Eq. 80 is

\[
F_i = \frac{\omega_S e_i^2 R}{\omega I_{id} 4kT R R A f}
\]

(83)

In the paragraphs to follow, it will be assumed that the fluctuations in the magnetization will give rise to a perturbation in \(L(t)\) of the form

\[
\Delta L(t) = \epsilon_n L \cos(\omega n t + \theta_n)
\]

(84)

where \(\epsilon_n\) and \(\theta_n\) are random functions of time, and the \(n\)'s denote frequencies that will cause output noise at the signal frequency. In the development \(\epsilon_n\) is considered small so that all \(\epsilon_n^2\) terms in the current expressions can be neglected.

First, fluctuations in the inductance at the pump frequency denoted by \(n = p\) will produce noise currents at the signal frequency. A coherent signal current in the signal circuit will mix with the incoherent fluctuations of \(L(t)\) to produce a current in the idler circuit just as the mixing action of the signal current with the coherent fluctuation of \(L(t)\) produces current in the idler circuit, and, in the same manner as in the desired amplifying action, this noise current flowing in the idler will react with the coherent variation of \(L(t)\) to produce a noise current in the signal circuit. Also, the idler current caused by the signal being amplified will mix with the incoherent fluctuation of \(L(t)\) to produce a noise
current at the signal frequency. Equation 30 can be modified to account for the $\epsilon_p$ factor that occurs in the incoherent fluctuation in $L(t)$ to yield the "feedback" voltage for the first contribution to the noise current in the idler circuit by multiplying the equation by $\epsilon_p$, adding the phase angle $\Theta_p$ and taking only the term involving the idler frequency impedance. The noise current produced in the signal circuit is

$$i_{pl} = -\frac{I_{l1}^2 \epsilon_p \omega_{sl} \sin(\omega_{st} + \Theta_s + \Theta_p)}{4R_u(R_g + R_p + R_s - R)} \quad (85)$$

where midband tuning has been assumed. For midband tuning, the coherent current in the idler circuit due to the signal frequency current can be obtained from Eq. 24 as

$$i_i(t) = -\frac{I_{l1}^2 \omega_{ll} \cos(\omega_{ll} t - \Theta_s)}{2R_u} \quad (86)$$

This current, when interacting with the incoherent fluctuation of $L(t)$ will yield a noise current in the signal circuit

$$i_{p2} = -\frac{I_{l1}^2 \epsilon_p \omega_{sl} \sin(\omega_{st} + \Theta_s - \Theta_p)}{4R_u(R_g + R_p + R_s - R)} \quad (87)$$

These two currents have the same incoherent source and must be added directly to give

$$i_p = -\frac{I_{l1}^2 \epsilon_p \omega_{sl} \left[\sin(\omega_{st} + \Theta_s + \Theta_p) + \sin(\omega_{st} + \Theta_s - \Theta_p)\right]}{4R_u(R_g + R_p + R_s - R)^2} \quad (88)$$

The mean square of this current is
The noise power delivered to the load resistance $R_p$ is then

$$\overline{i_p^2} = \frac{I_s^2 R_p^2 \epsilon_p^2}{(R_p + R_p + R_s - R)^2} . \quad (89)$$

and the contribution to the noise factor becomes

$$F_p = \frac{I_s^2 R_p^2 \epsilon_p^2}{4kTR_g \Delta f} . \quad (91)$$

Another source of noise at the output will be the incoherent fluctuations of $L(t)$ at twice the signal frequency, since the signal frequency current will interact with that fluctuation to yield an output power at $2\omega_s - \omega_s = \omega_s$. If that incoherent fluctuation is written

$$L_{e2}(t) = \epsilon_{e2} L_{\perp} \cos(2\omega_s t + \Theta_2), \quad (92)$$

the resulting noise current can be computed just as the signal current was computed for the coherent fluctuation of $L(t)$. When the output noise power from this incoherent fluctuation is computed and the result is divided by the parametric amplifier's gain and by the input noise power, the contribution to the noise factor becomes

$$F_{e2} = \frac{\omega_s I_s^2 R_{Rb} \epsilon_{e2}^2}{\omega_{\perp} 4kTR_g \Delta f} . \quad (93)$$

Yet another source of noise power at the output will be the incoherent fluctuations of $L(t)$ at the angular velocity $2\omega_{\perp}$, since the
coherent idler frequency currents will interact with this fluctuation to produce noise voltages at the angular velocity \(2\omega_{11} - \omega_{11} = \omega_{11}\). These noise voltages will be downconverted into the signal frequency circuit and produce output noise power at the load resistance. If that incoherent fluctuation is written

\[
L_{e3}(t) = e_{3}L_{1} \cos(2\omega_{11}t + \theta_{3}),
\]

the contribution to the noise factor can be computed just as that for \(e_{1}^{2}\) was computed for Eq. 83. The term to be added to the noise factor as the result of this incoherent fluctuation is

\[
F_{e3} = \frac{\omega_{11} I_{s} R \epsilon_{3}^{2}}{4kT R_{u} \Delta f}.
\]

The total noise figure for the thin ferromagnetic film parametric amplifier is then

\[
F = 1 + \frac{R_{p} + R_{s}}{R_{g}} + \frac{\omega_{s} R}{\omega_{11} R_{g}} + \frac{1}{4kT R_{u} \Delta f} \left[ \epsilon_{2}^{2} + \frac{\omega_{s} R}{\omega_{11} R_{u}} \epsilon_{1}^{2} \right] + \frac{\omega_{s} I_{s}}{4kTR_{u} \Delta f} \left[ \frac{\omega_{11} R}{\omega_{s}} \epsilon_{2}^{2} + \frac{\omega_{11}}{R_{u}} \epsilon_{2}^{2} + \frac{R^{2}}{R_{u}} \right].
\]

Since all the extra terms in this equation over those of Eq. 80 are the result of the thermal agitation of the magnetization, it would appear that the thin ferromagnetic film parametric amplifier would be much less noisy if it could be cooled. The films available to the writer did not however, consistently follow this pattern.
V. EXPERIMENTAL RESULTS

Of the engineering problems involved in the construction of a thin ferromagnetic film device, perhaps the greatest was that of constructing a coil with sufficient magnetic coupling to the film and yet one with a minimum capacitive coupling to the pump circuit. The writer attempted various schemes of shielding and canceling but found that the best results were obtained by using a relatively large diameter coil and accepting the reduced magnetic coupling in exchange for a reduced capacitive coupling. The coil used in these experiments was wound on a glass capillary tube whose inside diameter was much greater than the 5 mil diameter of the film coated wires. The No. 37 wire turns were spaced slightly to reduce the interturn capacitance. The turn spacing was such that there were approximately 60 turns per centimeter. The coil had an outside diameter of approximately 1.5 millimeter and an inductance of 1.82 microhenries. The coil was mounted on a bakelite block with teflon discs in slots to facilitate centering and aligning the ferromagnetic film wires. The ferromagnetic films were then mounted solidly on wire pegs so they would not be moved or strained during the alignment procedure. The easy direction of magnetization for the films used in this study was nominally circumferential.

At the beginning of each test run a film was aligned and centered within the coil while the system was being operated as a balanced modulator. This procedure was one developed earlier by the writer in his investigation of ferromagnetic film balanced modulators (11). Although this method is one of trial-and-error, it is effective and will be described briefly. During alignment the output of the modulator is observed on an oscilloscope
synchronized with the modulating frequency. Misalignment causes the dis-
symmetry illustrated in Fig. 5b. The output is observed with the bias
current applied in a given direction. The coil is then adjusted to pro-
duce maximum symmetry. Then the bias current is reversed and the coil re-
adjusted to again produce maximum symmetry. This process is repeated un-
til no change can be detected in the output wave form when the bias cur-
rent is reversed. The element must be magnetically shielded from the
earth's field during this procedure.

It was found that the easy direction of magnetization of film C was
far from circumferential to the axis of the wire, since it was obvious the
film axis was not common with the coil axis after a few attempts to follow
the alignment procedure described above. Since the individual filaments
were cut from the same wire, one might logically deduce that the other
filaments' rest directions may have deviated from circumferential by a
smaller angle. It is also possible that film C was damaged during prepar-
ation and mounting within the coil; however, each of the films was mounted
several times and only film C exhibited this characteristic.

A. Impedance Measurements

Since the model proposed by Read had not been verified experimentally,
and since there was concern that the film electrodeposited on wire might
be inferior to that vacuum deposited on glass, impedance measurements were
made on several films. It was not possible to use the usual impedance
measuring devices since their "test signals" were sufficient to destroy the
single domain character of the films. The measurements were made by series
tuning the element with a capacitor and measuring the current and voltages with a Hewlett Packard Model 411A RF Millivoltmeter. The series resistance and inductance were then converted to their equivalent parallel resistance and inductance. During these measurements, the elements were shielded from the earth's magnetic field by a mu metal box. The results are plotted in the A parts of Figs. 6 through 12. The films were cut in the sequence shown from a single longer film. It should be noted that the measurements of the resistances were very approximate at the high values of bias current and the measurements of both resistance and inductance were very approximate for the small values of bias current. These films were then tested in devices to determine their susceptibility to temperature change and to external fields. These tests will be discussed in the following paragraphs.

B. Balanced Modulator Tests

Temperature and stray field tests were made on a balanced modulator. The films were at first shielded from the high frequency fields of the other circuit elements by a small brass case and shielded from the earth's magnetic field by a large mu metal box. The output signal was attenuated by a 47 ohm resistor, amplified by the 75A3 Collins receiver, and detected by a Hewlett Packard 300A harmonic wave analyzer. This was not a noise measurement, except in the sense that pump frequency feedthrough can be considered noise if it impairs detection of the modulation signal. No attempt was made to improve the alignment over that previously described, because the films were not accessible. There was capacitive coupling
between the signal winding and the pump conductor. The films were biased with a current of one ampere and the output power versus pump power observed. This established the approximate level of pump frequency feedthrough due to capacitive coupling.

The original intention was to make controlled temperature tests on the films; however, the initial tests proved to be sufficient to establish their temperature sensitivity, and the controlled temperature tests were deemed unnecessary. Figures 8b and 9b show the variation in pump-frequency feedthrough for a given pump input as a ratio of output power \( P \) to output power \( P_0 \) with a bias current of one ampere versus bias current. This output power could result from angular dispersion, misalignment, discontinuous domain wall motion, and the capacitive coupling. Also, the pump frequency currents flowing in the signal winding will interact with the second harmonic component of the parametric inductance to give a voltage at the pump frequency. The magnitude of the pump-frequency feedthrough power was relatively high because the pump power was maintained at a high level during these tests; the purpose was not to minimize the pump-frequency feedthrough but to observe its variation with temperature. These same figures show the pump-frequency output power variation with the magnetic shield removed. The temperatures for these tests were not measured.

For the heat tests, a lead heat sink was attached to the film's ground terminal and heated with a soldering iron. The film was then the only circuit element subjected to temperature increase. For the cold tests, a pan of dry ice was placed within the radio frequency shield containing the film and the signal winding. In this case, the film, the signal winding and,
to some extent, the adjacent circuit elements were cooled. The above mentioned figures show the pump-frequency output power variation just after the heat sink had been heated and the shields assembled. This assembly required approximately one minute. For the cold tests, the films were cooled approximately thirty minutes before the pump-frequency feedthrough variations were measured.

The variation of pump-frequency feedthrough power versus time for films F and G is shown in Fig. 13. Both the heated films and the cooled films were cooling with increasing time. These temperature tests were made after the noise measurements had been completed. The purpose of these temperature tests was to determine the reason for the noise figure increase with cooling for film F as opposed to the noise figure decrease with cooling for film G. The normalizing value $P_1$ for these curves is the room temperature pump-frequency feedthrough with a bias current of 150 milliamperes. It appears that the variations in the noise figure and feedthrough measurements are the result of strain caused by the cooling.

C. Noise Measurements

Because of the variations observed in the film's impedance properties, it seemed desirable to use them as elements in a device for the purpose of a noise measurement to compare their noise characteristics with their impedance characteristics. The coupling winding and films were too small to give measurable gains when used as elements in either a balanced modulator or an upconverter but were sufficiently large for use in a parametric amplifier. The negative impedance characteristic of a parametric amplifier,
however, presented the problem of impedance matching. The noise factor of the following amplifier would have been altered had it "looked-back" into an impedance other than fifty ohms, and there would have been no accurate method to measure the signal power into the device for other than a fifty ohm input impedance. These problems were avoided by placing a resistor in the signal path to cancel the negative resistance generated by the element. This resistor is shown in Fig. 14 as a variable resistor, but one-quarter watt carbon resistors were used.

The accuracy of the noise measurement was only nominal, because the calibrated noise measurement equipment necessary for absolute accuracy was not available. Since the method used was not standard, it will be described in complete detail in order that the reader may evaluate the results.

Figure 15 is a block diagram of the equipment used for the noise measurement. A fifty ohm composition resistor network was first placed inside the device at the input terminal to replace the actual device circuit. A Boonton 250A R-X impedance bridge was then adjusted to a null on this resistor. The impedance bridge had been previously loaded to reduce the amplitude of its test signal; this loading reduced the measurement accuracy and made this comparison technique necessary. Then the resistor was removed and the actual device circuit was connected to the input terminal. The device's output terminal was then connected to the following amplifier through an attenuator to avoid saturating that amplifier. With the bias current set at the desired value, the circuit was tuned and the pump power adjusted to null the impedance bridge. The device was then overcoming its
own losses and the loss of the resistor R placed in the signal input path. The device gain was then unity.

Next, the input terminal of the device was connected to the Hewlett Packard 608C signal generator, and the attenuator between the device and the preamplifier was removed. The input signal power $P_1$ was then adjusted to yield a value of output power $P_0$ measured on the thermocouple meter. This output power divided by the system gain $G_c$ was

$$\frac{P_0}{G_c} = kT\alpha f + n_1 + n_2 + P_1$$

(97)

where $kT\alpha f$ is the input noise power from a matched source impedance, $n_1$ is the device's excess noise power, $n_2$ is the system's excess noise power less the device's excess noise power and $P_1$ is the signal power from the signal generator. This value of output power was then compared to a previously prepared calibration curve for the system without the device in the circuit. This same value of output power was for that case

$$\frac{P_0}{G_c} = kT\alpha f + n_2 + P_2$$

(98)

where $P_2$ is the signal power input required to achieve the output power $P_0$ when the device was not in the circuit. Taking the difference of Eq. 97 and Eq. 98 gives

$$n_1 = P_2 - P_1.$$  

(99)

Then

$$F = 1 + \frac{P_2 - P_1}{kT\alpha f}$$

(100)
where $F$ is the device's noise factor.

It was critical that the device gain be equal to unity in these measurements. Had the gain been greater than unity, the signal gain would have appeared to be excess noise generated within the device. Conversely, signal degradation would have made the device's excess noise appear to be less than its actual value. To guard against the possibility of signal gain or loss, the output signal was observed on the Hewlett Packard Model 300A harmonic wave analyzer indicated in Fig. 15 with its noise bandwidth set at approximately seventy cycles per second. The input signal was adjusted to give an output signal level well above the residual noise level when the device was not in the circuit; then, with the device in the circuit, the output signal was observed to be approximately the same as it had been without the device and to be well above the residual noise level for the films with the smaller noise factors. The output signal was also applied to a filter with a noise bandwidth of approximately one hundred cycles per second to facilitate observation on the Tektronix 565 oscilloscope.

These methods for impedance matching and checking required that the device be operating at relatively high signal levels so that if it had been saturating at these high signal levels there would have been some gain at the smaller signal levels used for the noise factor calculations. The tendency toward error, then, would have been for the noise factors to be greater than their true value. This procedure was repeated at least once for each point and several times for the points with small noise factors. For the smaller noise factors, the maximum ratio in the measured excess
device noise power for any two measurements at a single point was 1.24.
The noise figures shown are computed from the averages of the excess noise powers measured at each point. The measurements of the heated films were made quickly so they are probably in considerable error; however, the increase in excess noise with an increase of film temperature is clear despite some error. It is interesting to note that the excess noise of film F, Fig. 11, increased when cooled while the excess noise of film G, Fig. 12, decreased when cooled. This would appear to be the result of straining the magnetic material and may be a characteristic of films deposited on wire. No tests were made on flat films deposited on glass.

The noise figures shown in the B parts of Figs. 6, 7, 10, 11 and 12 are for \( R = 5.6 \) ohms and for the seemingly trivial case of \( R = 0 \) ohms. The elements were quite small, however, and required a large pump power to overcome their own losses and the losses of the circuit.

The noise figures should decrease for increasing bias current for two reasons. First, the film's losses should decrease for increasing bias current, and, second, the film should tend to form a more nearly perfect single magnetic domain for increasing bias current.
Figure 5. Illustration of the input and output waveforms of a thin film balanced modulator

(A) The waveforms for a modulator with an aligned film.
(B) The waveforms for a modulator with a misaligned film.
(A)

(B)
Figure 6. Measured characteristics of film A

(A) The inductance $L_0$ and resistance $R_0$ of film A corresponding to those illustrated in Fig. 2.

(B) The noise figure versus bias current $I_b$ for film A operating in the circuit illustrated in Fig. 14.
Figure 7. Measured characteristics of film B

(A) The inductance $L_0$ and resistance $R_0$ of film B corresponding to those illustrated in Fig. 2.

(B) The noise figure versus bias current $I_B$ for film B operating in the circuit illustrated in Fig. 14.
\[ L_0 \approx \frac{0.985}{1 + 0.0107 I_b} \]

- ○ 14.5 Mc
- □ 27 Mc

(A)

ROOM TEMPERATURE \( R = 0 \)

COOLED \( R = 0 \)

(B)
Figure 8. Measured characteristics of film C

(A) The inductance \( L_0 \) and resistance \( R_0 \) of film C corresponding to those illustrated in Fig. 2.

(B) The pump-frequency feedthrough power versus bias current \( I_B \) for film C. The reference power \( P_0 \) is the pump-frequency feedthrough power with a bias current of one ampere.
Figure 9. Measured characteristics of film D

(A) The inductance $L_0$ and resistance $R_0$ of film D corresponding to those illustrated in Fig. 2.

(B) The pump-frequency feedthrough power versus bias current $I_b$ for film D. The reference power $P_0$ is the pump-frequency feedthrough power with a bias current of one ampere.
\[ L_0 \approx \frac{1.03}{1 + 0.011 I_b} \]

(A) \( L_0 \) (\( \mu \text{H} \)) vs. \( I_b \) (mA)

- \( 14.5 \) Mc
- \( 27 \) Mc

(B) \( 10 \log \frac{P}{P_0} \) vs. \( I_b \) (mA)

- No Magnetic Shield
- Heated
- Cool
- Room Temperature
Figure 10. Measured characteristics of film E

(A) The inductance $L_0$ and resistance $R_0$ of film E corresponding to those illustrated in Fig. 2.

(B) The noise figure versus bias current $I_b$ for film E operating in the circuit illustrated in Fig. 14.
Figure 11. Measured characteristics of film F

(A) The inductance \( L_0 \) and resistance \( R_0 \) of film F corresponding to those illustrated in Fig. 2.

(B) The noise figure versus bias current \( I_b \) for film F operating in the circuit illustrated in Fig. 14.
\[ L_0 \approx \frac{1.07}{1 + 0.011 I_b} \]

- ○ 14.5 Mc
- □ 27 Mc

(A) NO MAGNETIC SHIELD \( R = 0 \)

(B) COOLED \( R = 0 \)

\[ R = 5.6 \]

ROOM TEMPERATURE \( R = 0 \)
Figure 12. Measured characteristics of film G

(A) The inductance $L_0$ and resistance $R_0$ of film G corresponding to those illustrated in Fig. 2.

(B) The noise figure versus bias current $I_b$ for film G operating in the circuit illustrated in Fig. 14.
\[ L_o \propto \frac{1.13}{1 + 0.011 I_b} \]

- **14.5 Mc**
- **27 Mc**

(A) 

(B) 

- **NO MAGNETIC SHIELD** \( R = 0 \)
- **ROOM TEMPERATURE** \( R = 0 \)
- **COOLED** \( R = 0 \)
- **\( R = 5.6 \)**
Figure 13. The variation of pump-frequency feedthrough power for films F and G as the heated and cooled films cooled with increasing time

(The normalizing value $P_1$ is the feedthrough power at normal room temperature.)
Figure 14. Schematic drawing of the parametric amplifier circuit used for the noise measurements
Figure 15. Block diagram of the equipment used for the noise measurements
VI. DISCUSSION

The results of this investigation indicate that thin ferromagnetic film devices are not necessarily noisy, but they also indicate that many problems remain to be solved before they will be useful for low level amplifiers.

The very small feedback impedances of the balanced modulator and the upconverter, present a serious problem in that a significant fraction of the available signal power at their input terminals is dissipated in the coupling winding(s) when conventional coils are used to provide the magnetic coupling to the films. This loss could be reduced by using a material for the coupling winding(s) which would become a superconductor when adequately cooled; however, the effects of straining the film and associated circuitry may offset the advantage gained from such cooling.

Seven films, all cut from the same parent film, were tested in this investigation. It should be noted that much of the variation in their characteristics could have been the result of damage between the time of deposition and the time of final mounting. Much of this interim history is not known.

The encouraging results of the noise measurements indicate that thin ferromagnetic film devices are potential competitors to other parametric devices. For low noise operation, thin film devices require careful shielding from the high frequency magnetic fields of the associated circuit elements and external sources such as the earth's magnetic field and the magnetic fields around power lines.

Some of the unknown factors in the noise measuring equipment used in
this investigation are sufficiently important to cause the writer to hesitate in estimating the measurement error. Since the excess over the minimum Johnson noise of many resistors is considerable, the output impedance of the signal generator used may have been delivering a large excess noise power; however, for the method used, this generator's output impedance was the source impedance for the system when the device was in the circuit and when it was not in the circuit. Therefore, the excess noise of the signal generator should have been effectively subtracted out.

The necessity for taking the difference of two power readings on the signal generator's attenuator dial could be another source of error. The generator's output power was calibrated at a relatively high level, but the noise measurement readings were taken at a much lower level. A small non-linearity in the attenuator, therefore, could have caused some error. The possibility of this error was minimized by taking measurements at several different power levels for each bias point. Finally, the complex procedure described in the noise measurement section could be the source of considerable error. The fact that it was possible to repeat a particular measurement of excess noise with only a small difference in results, however, indicates to the writer that the measurement was probably accurate to within 3 decibels.

To the best knowledge of the writer, the noise measurements described here and those described by Fanslow (5) are the first noise measurements made on thin ferromagnetic film devices. Although more elaborate measuring equipment could reduce the probable error in the measurement, it appears to this writer that the major problem at present in thin
ferromagnetic film devices is one of device fabrication rather than the excess noise of the film. As far as amplifying devices are concerned, the next effort should be toward achieving better fabrication techniques.
VII. BIBLIOGRAPHY


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