pressure windows, was 1.28 over the band of interest.

The measured stop band insertion loss is shown in Fig. 4. The insertion loss of the TE_{10} mode was measured using smooth tapered sections from waveguide connected directly to the input and output of the filter. A continuously swept signal generator was connected at the input and the filter output was monitored on a scope. A TWT amplifier operating from 4–11 Gs/c provided additional sensitivity throughout this band. The limit of measurements for the other frequencies of interest was 55 dB. Point-to-point receiver measurements were taken throughout the band of limited sensitivity.

A point of further interest concerned the location of the waffle-iron filter assembly relative to the radar system duplexer. Figure 5 illustrates the results of dynamic co-operative tests performed by the radar and a microwave link station operating at the subject radar's third harmonic frequency. The objectives of the test was to ascertain: 1) the effectiveness of this waffle-iron filter in reducing interference levels beyond the television fade margin of the microwave link and 2) to determine the optimum location for the waffle-iron filter assembly relative to the duplexer in the radar system.

It can be seen from the graph that the range of levels received with the waffle-iron filter before the duplexer were 5 dB below the fade margin for teletype but still within the television fade margin. Placing the filter on the antenna side of the duplexer resulted in mean signal levels 17 dBm below the previous reading. This data furnishes conclusive proof of TR tube harmonic radiation and the necessity for packaging such waffle-iron filters on the antenna side of the duplexers.

This correspondence has shown that waffle-iron filters can be utilized at moderate power levels at L-band. Higher levels could also be achieved by utilizing a technique of paralleling filters [3] using binary symmetrical power dividers. Thus, a double layer filter could be packaged within the dimensions of L-band waveguide. By dividing the power among two filters and then recombining their outputs, the power handling capacity is doubled.

Heavy wall construction was utilized (Fig. 1) to provide distortion free operation with high internal pressures. A considerable weight savings can be achieved if a ribbed construction were used with a thin wall thickness.

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Correction to "Exact Design of Band-Stop Microwave Filters"

In the above paper[1] the authors have called the following to the attention of the Editor:

The formula for Z_4, for case $n=5$ in Table II (page 8), should have read

$$Z_4 = \frac{Z_A}{g_0}\left(\frac{1}{1+\Lambda g_5 g_6} + \frac{g_6}{\Lambda g_4(1+\Lambda g_5 g_6)^2}\right).$$

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[1] B. M. Schiffman and G. L. Matthaei, *IEEE Trans. on Microwave Theory and Techniques*, vol. MTT-12, pp. 6–15, January 1964.

The Electric Field and Wave Impedance in Spin Wave Propagation

The propagation of electromagnetic waves in magnetic materials, allowing exchange coupling, has been treated by Auld [1] and by Soohoo [2]. Their calculations involved the simultaneous solution of Maxwell's equations and the equation of motion of the magnetization, for plane waves.

According to the well known results of Auld [1] for ferrimagnetic materials, the susceptibility tensor is of the same form as the well-known Polder tensor for the uniform mode, except that the internal static magnetic fields are augmented by an exchange field in the case of magnetic waves with a finite wave vector $\boldsymbol{k}$. Depending on the magnitude of $\boldsymbol{k}$, the various plane waves could be classified into electromagnetic, magnetostatic, and exchange types.

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In the interpretation of these results, the statement has been made that as the spin wavelength becomes shorter and shorter, the RF electric field becomes increasingly smaller compared with the RF magnetic field and can eventually be neglected.

We suggest that this statement is perhaps somewhat too strong, for it is desirable to examine not only the ratios of absolute quantities of electric and magnetic fields, but such quantities as the transverse wave impedance. Here we find the general rule that as the wavelength decreases, the wave impedance increases. This is seen by substitution of a magnetic field intensity of form $[\hat{x}h_0+\hat{y}h_1][\exp\{j(\omega t-kz)\}]$ into the Maxwell curl equation, in the absence of conduction current. The result is

$$\boldsymbol{e} = [\hat{x}(h_1)(k/\omega\epsilon_r\epsilon_0) + \hat{y}(-h_0)(k/\omega\epsilon_r\epsilon_0)] \cdot \exp[j(\omega t - kz)] \qquad (1)$$

so that

$$Z_{\text{transverse}} = |e_x/h_y| = |e_y/h_x| = k/\omega\epsilon_r\epsilon_0. \qquad (2)$$

As the wavelength gets shorter, k becomes larger; so that the ratio $|e/h|$ increases. At 10 Gc, for example, for $\epsilon_r=15$, and for wavelength $\lambda_g=(5)(10^{-7})$ meters, $|e/h|=(1.5)(10^6)$ ohms. Direct substitution of the appropriate quantities into Auld's expressions yields essentially the same result.[1]

[1] In private communications, Dr. Auld stated that he was aware of this and of the field structure. We appreciate Dr. Auld's kindness in communicating with us on this subject.

TABLE I

Wave Components for Spin Wave Propagation, for Circularly Polarized RF Magnetization

The dc magnetic field is taken in the z direction. Propagation is in the Y—Z Plane. Note that although the RF magnetization is circularly polarized, $\boldsymbol{e}$, $\boldsymbol{h}$, and $\boldsymbol{b}$ are circularly polarized only for $\theta=0°$. For this table, the units are MKS rationalized, with $\boldsymbol{b}=\mu_0(\boldsymbol{h}+\boldsymbol{m})$; $\epsilon=\epsilon_r\epsilon_0$; $k_f=\omega(\epsilon\mu_0)^{1/2}$.

	$\theta_k = 0°$	$\theta_k = 90°$
$\boldsymbol{m}$	$m_0\begin{bmatrix}-1\\ j\\ 0\end{bmatrix} e^{j[\omega t-kz]}$	$m_0\begin{bmatrix}-1\\ j\\ 0\end{bmatrix} e^{j[\omega t-ky]}$
$\boldsymbol{e}$	$-m_0\left(\frac{k_f}{k}\right)\left(\frac{\mu_0}{\epsilon}\right)^{1/2}\begin{bmatrix}-j\\ 1\\ 0\end{bmatrix} e^{j[\omega t-kz]}$	$m_0\left(\frac{k_f}{k}\right)\left(\frac{\mu_0}{\epsilon}\right)^{1/2}\begin{bmatrix}0\\ 0\\ -1\end{bmatrix} e^{j[\omega t-ky]}$
$\boldsymbol{h}$	$m_0\left(\frac{k_f}{k}\right)^2\begin{bmatrix}1\\ j\\ 0\end{bmatrix} e^{j[\omega t-kz]}$	$m_0\begin{bmatrix}-(k_f/k)^2\\ -j\\ 0\end{bmatrix} e^{j[\omega t-ky]}$
$\boldsymbol{b}$	$\simeq\mu_0 m_0\begin{bmatrix}-1\\ j\\ 0\end{bmatrix} e^{j[\omega t-kz]}$	$\mu_0 m_0\begin{bmatrix}-1\\ 0\\ 0\end{bmatrix} e^{j[\omega t-ky]}$

TABLE II
TYPICAL AMPLITUDES FOR SPIN WAVE PROPAGATION

$m_x = 3.4 \times 10^3$ amperes/meter	$p_{em} = 6.0 \times 10^{-9}$ watts/cm^2
$b_x = 4.3 \times 10^{-3}$ webers/meter	$p_{mag} = 3.5 \times 10^{-2}$ watts/cm^2
$h_x = 4.7 \times 10^{-6}$ amperes/meter	
$e_x = 12.8$ volts/meter	

Since spin waves are a magnetic, rather than an electric phenomenon, this is rather surprising. However, the mystery is cleared up by computing the magnetization. The results, summarized in Table I, show that the RF magnetization $\boldsymbol{m}$ dominates the picture; the ratio $|m/e|$ also varies directly as k.

It is of interest to compute the field intensities at a particular level of power density. These are given in Table 2, for z-directed spin wave propagation in a typical case ($4\pi M_s = 2439$ gauss; exchange constant $A = (4.9)(10^{-7})$ ergs/cm; $\epsilon_r = 14$; precession angle of 1°; 3000 Mc/s). Included in this table are electromagnetic power density p_{em}, computed as $\boldsymbol{E} \times \boldsymbol{H}$, and the magnetization power density p_{mag}. The latter was computed by multiplying the energy density stored in the magnetization by the group velocity.

We note that the energy of the magnetic system far exceeds that of the electromagnetic field. Consequently, the approximation of neglecting the electromagnetic field in all those calculations in which spin wave propagation is to be analyzed appears justified. We *do* notice, however, that the electric field is of appreciable magnitude.

In view of the presence of an E field of reasonable amplitude, it may be possible to employ this field in some useful manner in the future. One possible application is in the excitation of spin waves by electric fields. Another is in the coherent interaction with charged particles, provided a low loss ferrimagnetic material that is also semiconducting can be found.

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The Synthetic Generation of Phase-Coherent Microwave Signals for Transient Behavior Measurements

The purpose of this correspondence is to describe a new technique for generating a microwave signal for use as a diagnostic tool in the investigation of the transient behavior of wide-band networks and dispersive media.[1] The test signal is a periodic, phase-coherent, pulse-modulated microwave waveform which has a negligible transient build-up time. It is phase-coherent in the sense that the phase of the carrier is exactly the same at the start of each pulse. This condition is essential if the individual cycles of the response of a system to the test waveform are to be observed on a sampling oscilloscope.

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When the duration T of the test signal is long compared to the transient "settling time" of a given network, the output voltage during the time $0 < t < T$ is essentially the step-modulated response of the system. The response to a RF pulse of arbitrary duration can then be found by appropriately shifting and subtracting replicas of the step-modulated response. Other test signals, such as the unit step function or the narrow video pulse, have only limited utility because the response to these signals must be further manipulated, mathematically, (e.g., the Duhamel integral for linear, fixed parameter systems) to determine the response at the nominal system operating frequency.

Mathematically, the required test waveform is of the form

$$\rho_T(t) = [u(t) - u(t-T)] \cdot \sin \omega_0 t * \sum_{n=-\infty}^{\infty} \delta(t - n\hat{T}) \quad (1a)$$

where

T is the pulse width
$\hat{T}$ is the pulse repetition period
ω_0 is the microwave radian frequency
$u(t)$ is the unit step function
$\delta(t)$ is the Direc delta function
$*$ indicates time domain convolution,

and where the signal within any given period is

$$\rho_T(t) = [u(t) - u(t-T)] \sin \omega_0 t. \quad (1b)$$

The technique presented below generates the pulse-modulated signal $\rho_T(t)$ "synthetically" by appropriately weighting, delaying, combining and filtering "video" nanosecond pulses.

The operation of the generator can be described best by first considering the "T" connection of distortionless TEM mode transmission lines shown in Fig. 1. Here the characteristic impedance of the forward transmission line and the short-circuited stub are normalized and are one ohm and $\frac{1}{2}$ ohm, respectively. If the unit step function is incident on port A at time $t=0$ second, then at $t = L/2c$ second the step function appears "undistorted" at junction B: L is the length of line $\overline{AB}$, and c is the speed of light

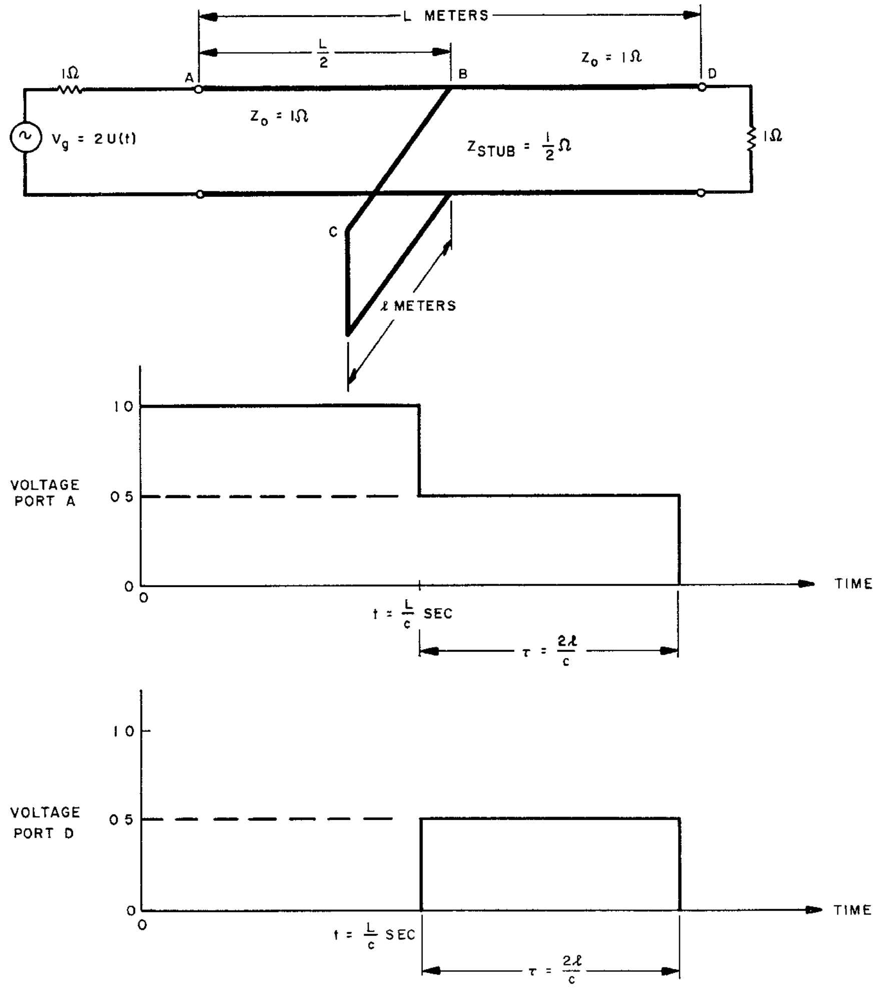

Fig. 1. Step response of a TEM mode T network.

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