TABLE II
TYPICAL AMPLITUDES FOR SPIN WAVE PROPAGATION

$m_x = 3.4\times10^3$ amperes/meter	$p_{em} = 6.0\times10^{-9}$ watts/cm²
$b_x = 4.3\times10^{-3}$ webers/meter	$p_{mag} = 3.5\times10^{-2}$ watts/cm²
$h_x = 4.7\times10^{-6}$ 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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[3] Dr. Soohoo's present address is the University of California, Davis.

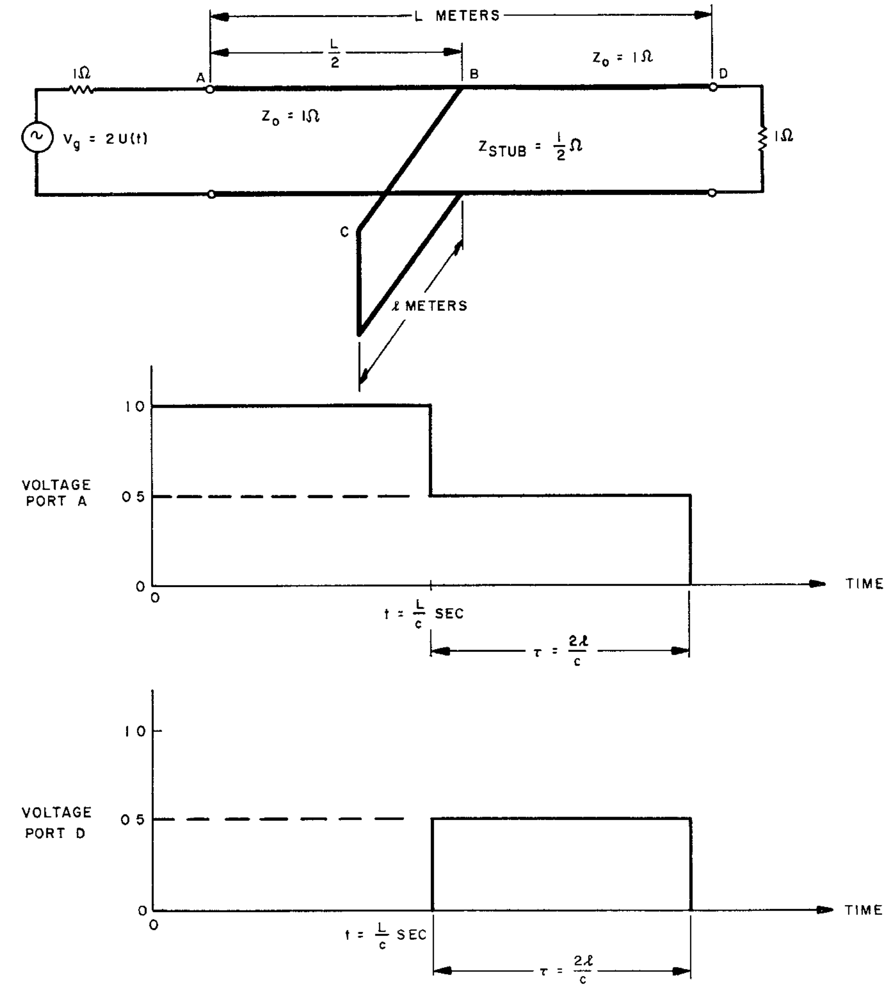

Fig. 1. Step response of a TEM mode T network.

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.

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 ½ 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

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[1] G. Ross, L. Susman, and J. Hanley, "Transient behavior of large arrays," Final Rept. AF30(602)-3348, TR-64-581, June 1965. This work was sponsored by the Air Force Systems Command, Research & Technology Division, RADC; John Potenza, Project Engineer.

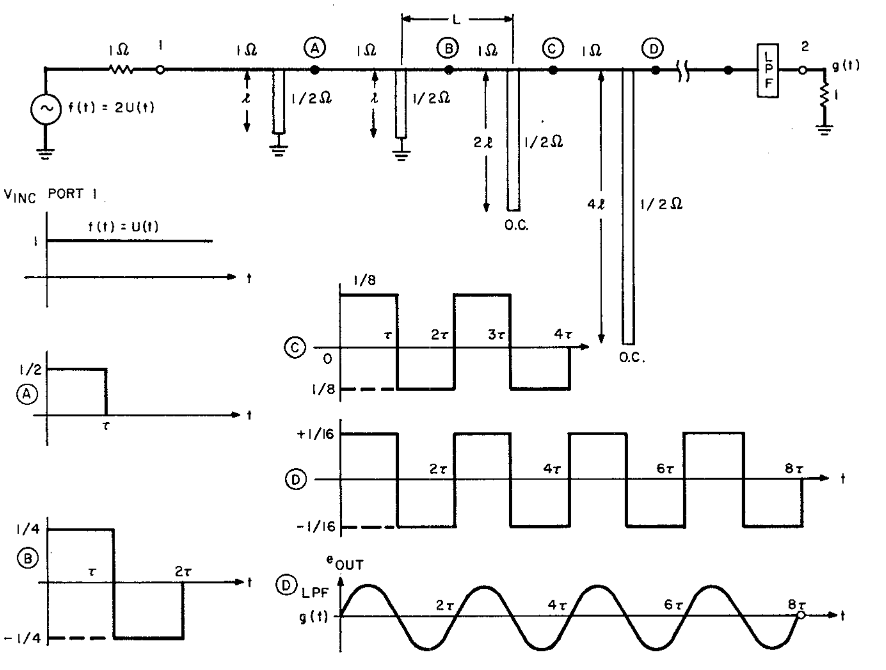

Fig. 2. The generation of synchronized RF signals, synthetically.

in the medium. At junction B the waveform senses a surge impedance of a one-ohm line in parallel with a $\frac{1}{2}$-ohm line, or a reflection coefficient of

$$\Gamma_B = \frac{z_{eq.} - 1}{z_{eq.} + 1} = \frac{\frac{1}{3} - 1}{\frac{1}{3} + 1} = -\frac{1}{2}. \quad (2)$$

Hence, $1+\Gamma_B$ or $+\frac{1}{2}$ volts is transmitted to port C and D, while $-\frac{1}{2}$ volts is reflected toward port A. The signals arriving at ports D and A are completely absorbed in matched terminations; however, the signal arriving at port C is totally reflected with opposite sign because of the short circuit. Thus, a signal of $-\frac{1}{2}$ volt is returned to junction B where it senses a perfect match (e.g., the surge impedance of lines $\overline{AB}$ and $\overline{BD}$ in parallel constitute a match with respect to the characteristic impedance of the stub), and is, therefore, transmitted back to port A and forward to port D without reflective loss. The leading edge of the $-\frac{1}{2}$ volt signal arriving at port A is completely absorbed reducing the total voltage (i.e., the sum of the incident plus reflected voltage) to zero. When the $-\frac{1}{2}$ volt signal propagating in the forward direction arrives at port D the total voltage is also reduced to zero, thus resulting in the formation of a rectangular pulse of amplitude $+\frac{1}{2}$ volt and duration τ-$2l/c$ sec, where τ is the time for the leading edge of the incident signal to propagate from junction B to port C and return.

The suggested generator is shown in Fig. 2. It consists of a cascade connection of TEM mode T networks similar to the one shown in Fig. 1 and previously described. The first two stubs are short-circuited and of equal length. Succeeding stubs are open-circuited, and are of increasing length, the output from the last T section is fed into a low-pass filter and then to the network under test.

The operation of the generator is as follows: application of a unit step function at port 1 results at point A in a rectangular pulse of amplitude $+\frac{1}{2}$ volts and duration $\tau = 2l/c$ sec, where l is the length of the first stub. When this pulse is fed into the second T section, a delayed and attenuated replica of the pulse is transmitted directly to point B. τ second later a similar but inverted pulse (returning from the short-circuited stub) joins the first waveform at point B. The result is a single "square" cycle of duration 2τ sec and $\pm\frac{1}{4}$ volts in amplitude.

The square cycle travels toward the third T section where the entire cycle is transmitted to point C after losing half of its amplitude. The square cycle propagating in the stub encounters an open circuit at a distance of $2l$ meters from the junction. Hence, this signal reflects with the same sign and joins the waveform at point C, 2τ second and joins the waveform at point C, 2τ sec later. The result is two square cycles of 4τ overall duration and of $\pm\frac{1}{8}$-volt amplitude.

When this signal is fed into the fourth section consisting of an open-circuited stub of length $4l$, the output at point D consists of four square cycles of 4τ duration and $\pm\frac{1}{16}$-volt amplitude.

In this manner, it can be seen that the number of square cycles N at the output of the kth stub, is given by

$$N = 2^{k-2} \quad (3)$$

while the output voltage amplitude V_0 is given by

$$V_0 = 2^{-k} \text{ volts.} \quad (4)$$

The overall pulse duration T is

$$T = N2\tau = 2^{k-1}\tau \text{ second} \quad (5)$$

and since

$$\tau = \frac{2l}{c}$$

$$T = \frac{2^k l}{c}. \quad (6)$$

The period of the generated waveform is 2τ second, or the fundamental frequency f_0 of the wave is

$$f_0 = \frac{1}{2\tau} \text{ c/s.} \quad (7)$$

If the output from the last T section drives an ideal low-pass filter whose cutoff frequency is approximately $2f_0$, then all higher harmonics (e.g., only odd harmonics of f_0 exist) of the N square cycle signal are attenuated and the resultant waveform $g(t)$ approaches

$$g(t) \doteq V_0[u(t) - u(t - T)] \sin \omega_0 t \rightarrow V_0 p_T(t) \quad (8)$$

or the required signal (except for a scale factor) described in (1). Actually, the ideal low-pass filter has a build-up time given by

$$T_{\text{LPF}} = \frac{\pi}{\omega_c} = \frac{1}{4f_0} = \frac{\tau}{2} \text{ sec,} \quad (9)$$

or $\frac{1}{4}$ of an RF period. (For example, at $f_0 = 1000$ Mc/s, the build-up time of the filter is approximately $\frac{1}{4}$ ns).

It should be noted that in the network shown in Fig. 2 there are reflections between junctions due to waves traveling in the back direction that eventually are re-reflected toward the output terminal. If the line lengths between stubs, however, are made "long" with respect to the length of the last stub, these undesired reflections are far removed from the required output, $g(t)$. For example, if f_0 equals 1000 Mc/s the length of the last (longest) stub for an 8-cycle signal is given by

$$l_5 = 2^{k-2}l = 2^{k-4}c(2\tau)$$
$$= 2(3 \times 10^8)(10^{-9}) = 0.6 \text{ meters}$$

where $k = 5$. Hence, the line lengths L between stubs must be greater than 0.6 meter in order to keep secondary reflections from distorting the desired signal. One meter between stubs would place the closest undesirable signal

$$\frac{2\Delta L}{c} = \frac{2(1 - 0.6)}{c} = 2.7 \text{ ns}$$

away.

The factor 2 in the preceding equation accounts for the two-way travel of the reflection.

A strip-line model of the generator shown in Fig. 2 was constructed for use at 1.3 Gc/s. A six-stub network was employed resulting in a 16-cycle pulse packet of microwave energy. The network was excited by a unit step function generator whose rise time was approximately 0.15 ns and the output was viewed on a sensitive 3.5 Gc/s sampling oscilloscope which also served as the low-pass filter. The output of the network is shown in Fig. 3. Note that there is virtually no discernible rise time of the output waveform: The polarity of the waveform was reversed by the camera. Work is continuing in an effort to reduce the small amount of amplitude modulation present. This modulation

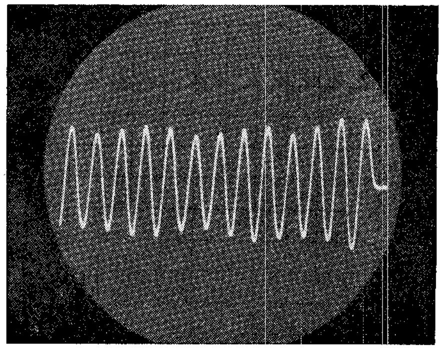

Fig. 3. Generator output B, 16 cycles of RF. Vertical scale: 0.0075 volt/div. Time scale: 1 ns/div. $f_0 = 1300$ Mc/s.

results from the imperfect cancellation occurring at the junctions which is due to construction tolerances and discontinuity capacity.

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Broadbanding Microwave Diode Switches

In the design of microwave switching networks, it is often necessary to design for minimum VSWR and low loss over a broadband. One class of PIN diode TEM microwave switch may be successfully broadbanded by utilizing the band-pass filter designs published by Mumford [1] and Matthaei [2].

Consider the S.P.D.T. diode switch in Fig. 1 which uses two pairs of PIN diodes shunting a TEM stripline circuit. (Diodes must be paired in stripline circuits to obtain maximum forward bias attenuation.) If the diodes in "diode gate" D' are forward biased and the diodes in D are reverse biased, then most of the incident generator power will flow to G_L and vice versa. The canonical form of the Fig. 1 switch in either condition is a four "stub" band-pass filter consisting of $\lambda_0/4$ shunt stubs of characteristic admittance Y_m separated by quarter wave lengths of connecting line with characteristic admittance $Y_{m,m+1}$. For the condition where generator power flows to G_L, the first stub is Y_1, the second is $Y_2 = Y_{23}'$, the third "stub" (normally denoted by Y_3) is the parallel resonant circuit at D obtained by tuning out the diodes' reverse biasc apacitance with the short inductive stub Y_α, and the fourth is Y_4. The filter will exhibit maximally flat or Tschebyscheff band-pass response if certain prescribed values of characteristic admittance are assigned to Y_m and $Y_{m,m+1}$.

The problem arises as to what value of equivalent quarter wave stub characteristic admittance (Y_3) one must assign to the D and Y_α resonant circuit. This is properly called a "quasi-lumped" stub, since the inductive stub Y_α is loaded by the lumped capacitance of D. One method, found to give good practical results, is to equate the mid-band susceptance slope of the quasi-lumped stub to that of an actual $\lambda_0/4$ stub of characteristic admittance Y_3. This may be done by considering Fig. 2, where

$$B = j\omega C - jY_\alpha \cot\frac{\omega d_\alpha}{V}$$

where

$$V = \frac{3\times 10^8}{\sqrt{\epsilon_r}}\ \text{meters/s}$$

$$\frac{dB}{d\omega} = j\left(C + Y_\alpha\frac{d_\alpha}{V}\csc^2\frac{\omega d_\alpha}{V}\right).$$

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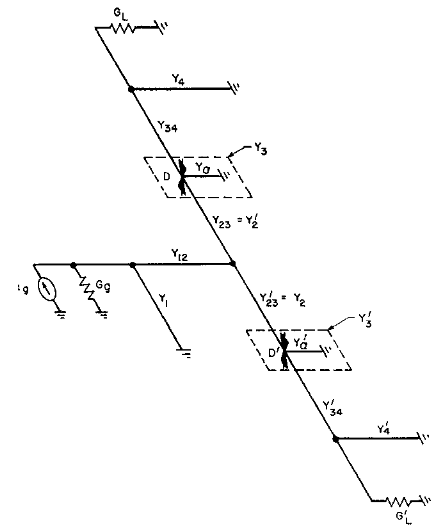

Fig. 1. Schematic diagram of stripline S.P.D.T. diode switch.

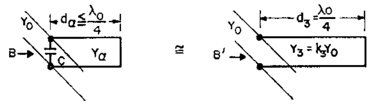

Fig. 2. Quasi-lumped stub and equivalent quarter-wave stub.

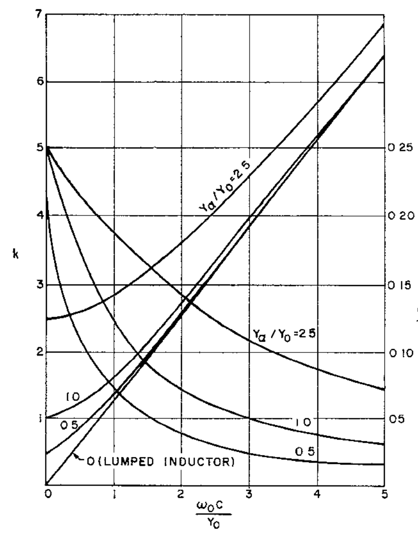

Fig. 3. Equivalent k of quasi-lumped stub (positive slope curves) and length d_α of quasi-lumped stub (negative slope curves), all vs. $\omega_0 C/Y_0$.

Also, for the quarter wave stub

$$B' = -jY_3 \cot\frac{\omega d_3}{V}$$

$$\frac{dB'}{d\omega} = jY_3\frac{d_3}{V}\csc^2\frac{\omega d_3}{V}.$$

Equating the two derivatives since we wish the selectivities to be equal near ω_0.

$$C + Y_\alpha\frac{d_\alpha}{V}\csc^2\frac{\omega d_\alpha}{V} = Y_3\frac{d_3}{V}\csc^2\frac{\omega d_3}{V};$$

let

$$\omega = \omega_0, \qquad d_3 = \frac{\lambda_0}{4} \qquad k_3 = \frac{Y_3}{Y_0}:$$

then

$$\csc^2\frac{\omega_0 d_3}{V} = 1$$

and

$$k_3 = \frac{4}{\pi}\cdot\frac{\omega_0 C}{2Y_0} + \frac{4}{\pi}\cdot\frac{\omega_0 d_\alpha}{2V}\cdot\frac{Y_\alpha}{Y_0}\csc^2\frac{\omega_0 d_\alpha}{V};$$

also for $B=0$ at $\omega=\omega_0$

$$\omega_0 C = Y_\alpha \cot\frac{\omega_0 d_\alpha}{V}$$

whence

$$k_3 = \frac{2}{\pi Y_0}\left\{\omega_0 C + Y_\alpha\left[\text{arc cot}\frac{\omega_0 C}{Y_\alpha}\right]\cdot\left[1+\left(\frac{\omega_0 C}{Y_\alpha}\right)^2\right]\right\}. \qquad (1)$$

k_3 may be considered as the normalized characteristic admittance of the equivalent quarter wave stub Y_3. Figure 3 plots k vs $\omega_0 C/Y_0$ for various values of Y_α, and may be used for designing a switch or any other bandpass structure with one or more quasi-lumped stubs.

One further constraint faces the designer of the Fig. 1 switch: for a symmetrical configuration (through this is not necessary) $Y_{23} = Y_{23}'$, and when Y_{23} is being used as a coupling line, Y_{23}' must serve as a shorted stub, and vice versa. Hence, the filter design equations must be examined to find if any cases exist where $Y_{23} = Y_{23}'$. In this respect Mumford's [1] equations for maximally flat filters are particularly useful since they have the constraint that $Y_{m',m+1} = Y_0$ for all cases. Therefore, the designer need only choose the case where $Y_{23} = Y_{23}' = Y_0$. Then, for the 4-stub filter $Y_3 = Y_0$ and $Y_1 = Y_4$. However, if the designer must use a diode capacitance C such that $Y_3 > Y_0$, then he should build a filter with an odd number of sections, and place D in the center since for maximally flat filters the center stub always has the highest characteristic admittance.

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The author wishes to acknowledge the helpful guidance received from W. W. Mumford who first proposed this method of attack

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Organic Superconductor and Dielectric Infrared Waveguide, Resonator, and Antenna Models of Insects' Sensory Organs

With reference to Little's recent paper [1], the following comments may be of value:

Nature may have long ago discovered the facts concerning the feasibility of organic

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