RF NOTE #19

214-0020-01-01_00

November 1977
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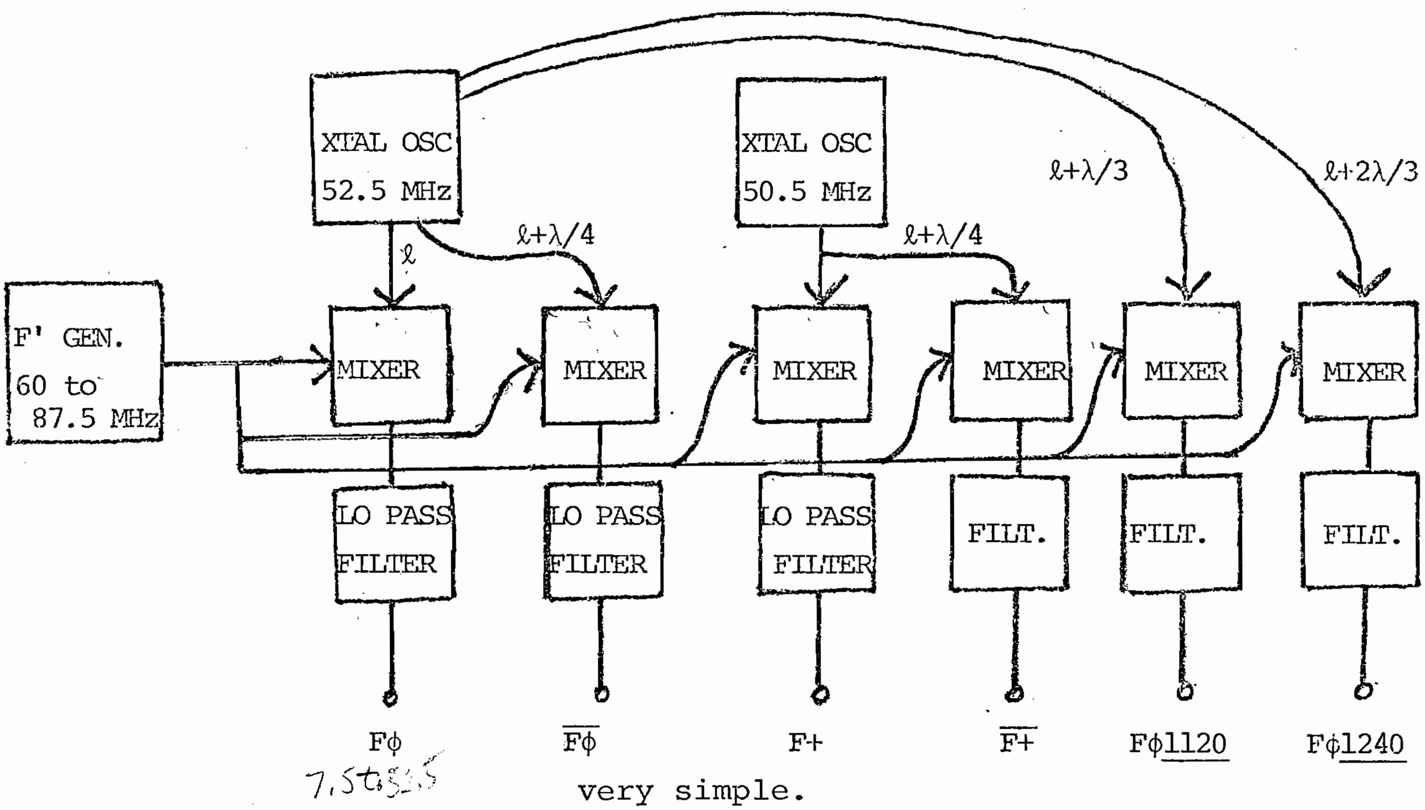
Conceptual Design for Frequency, Phase and Amplitude Control

Now that we have had "no" answers to our request for bids for the "special purpose synthesizer" that embodied multiple outputs of $F\phi$, $\overline{F\phi}$, F+ and $\overline{F+}$ (see RF Note#16 for definitions), we can proceed to do all these and more things right!

1. Frequency: $F\phi$, $\overline{F\phi}$, F+, $\overline{F+}$, $F\phi\angle 120^\circ$, $F\phi\angle 240^\circ$

After considerable mental stumbling, and with a good idea from P. Marchand, I now realize the way to do all this is as follows:

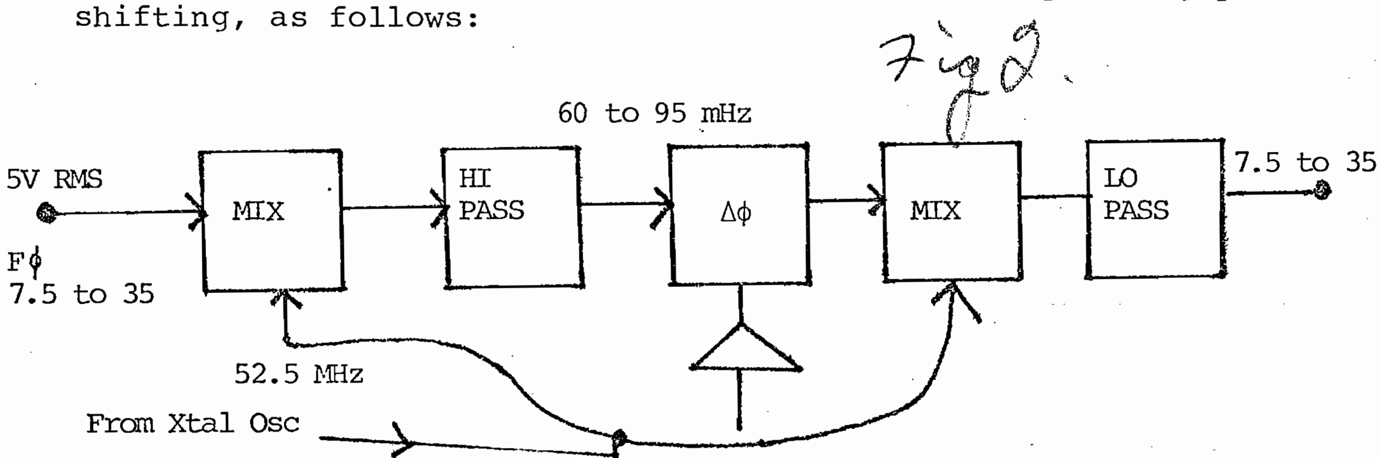


very simple.

The reason for choosing the F' generator frequencies is that the laboratory presently possesses a 50 to 100 MHz Government surplus property generator of that frequency range which is built like a xxx -- no -- like a, Chinese pagoda; namely, stable. Its output is a very pure sine wave and although its long time stability is probably no better than 100 PPM, I suggest using it instead of a $5000 synthesizer. The long time stability of 1 PPM can be achieved via a counter and computer control, both of which we will have.

2. Phase shifters.

Referring to RF Note #11, one will note that manual and electronic phase shifters are part of the conceptual design of the rf system. It had been thought that phase shifters of the type used on the Orsay storage ring project, copied from NAL, would be used. However, these are not very satisfactory over our large frequency range; the phase shift being proportional to frequency. So we have come up with a superior method of accomplishing phase shifting, as follows:



Obviously, this is a perfect way to accomplish phase shifting! The Δϕ is the same kind of animal as before, but with only a 30% ΔF/F. It will work fine. The present phase shifters need only be modified to move the cut-off frequency up to 100 mHz. This means different inductors and varicaps. For the constant K line we are using we must have

$$\sqrt{L/C} = 50$$

$$\sqrt{LC} = 1/2\pi x 10^8$$

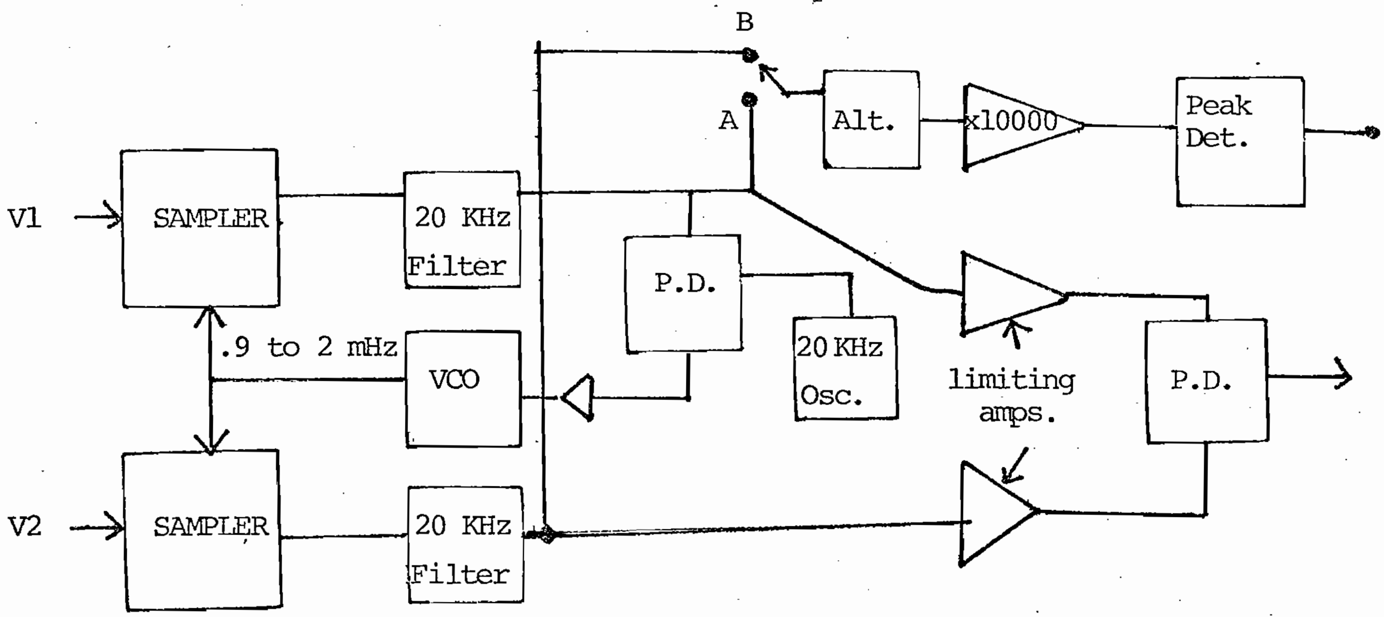
whence L ≈ .1μH, C_{nom} = 30 pf.
The delay per section is 1.6 ns and 20 sections give us 32 ns delay. If C varies by ±20% then the delay can vary by ±10% or a total variation of 6 ns or by about <u>120°</u> which is more than adequate.

After sample components are purchased and measured, the performance will be calculated on the program "SHIFT". All this is considered trivial.

3. <u>Phase detector</u>

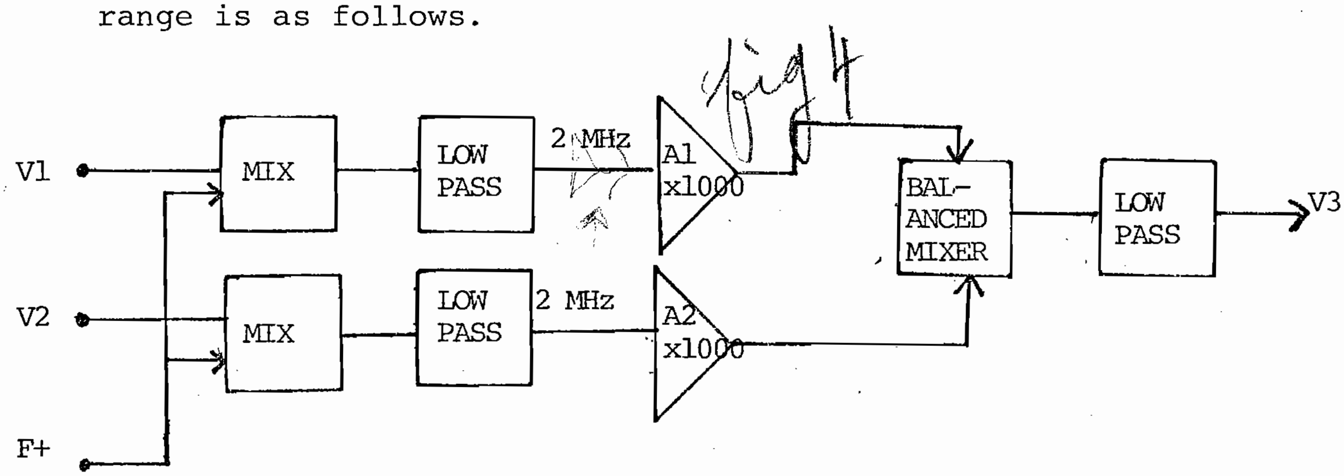
The simple type of phase detector described in RF Note #10 has many virtues: it is simple, very stable, and fast. But its output is proportional to the amplitude of the product of the two input signals, and then only if one of the inputs is greater than 1 volt peak. So although it gives a good output over a 5 to 1 amplitude range, it becomes excessively nonlinear for inputs over a 1000 to 1 amplitude range. So some other method must be found.

Now HP in their vector voltmeter proceeds as follows:



This method is quite satisfactory over a 1000/1 amplitude range, but is slow in that an operator must adjust the gain. It is slow also because the detection is done at 20 KHz and thus the output is limited to a frequency response of about 5 KHz.

A good way to accomplish linear phase detection over a 1000 to 1 range is as follows.



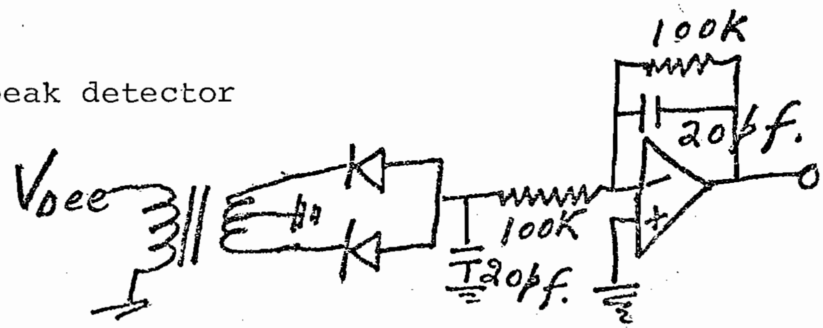
where A1 and A2 are limiting amplifiers such that the output square waves are not phase shifted relative to V1 and V2 as a function of the amplitude of V1 and V2. Development work is required to determine if this is possible. Q. Kerns built such amplifiers in 1950 using only vacuum tubes and 1N345 at a frequency of 14 mHz. So I say it is possible to do it now at 2 MHz. It is done with nonlinear feedback. If, as I have heard, one can buy 1GHz opamps, it will be a cinch.

Amplitude detection

If we assume that the phase detection system works -- and we might as well so assume, because it has to be made to work well; then amplitude detection is a snap. We have the following ways of doing this, or a combination of them, i.e. one way for slow but accurate detection, and another for the fast.

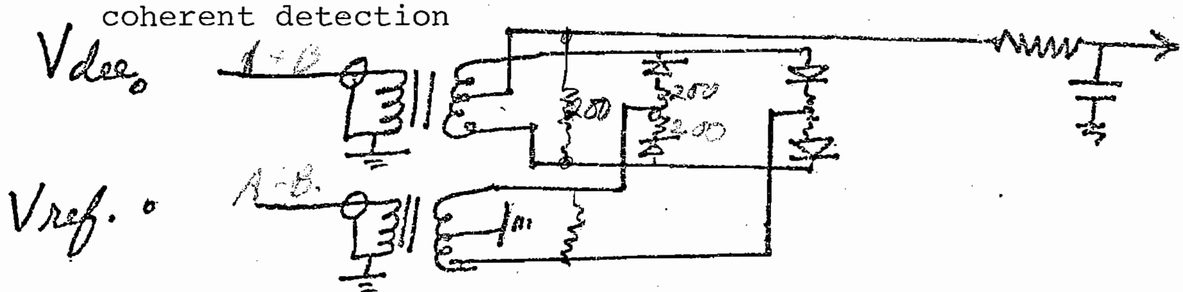
Detector A.

simple peak detector



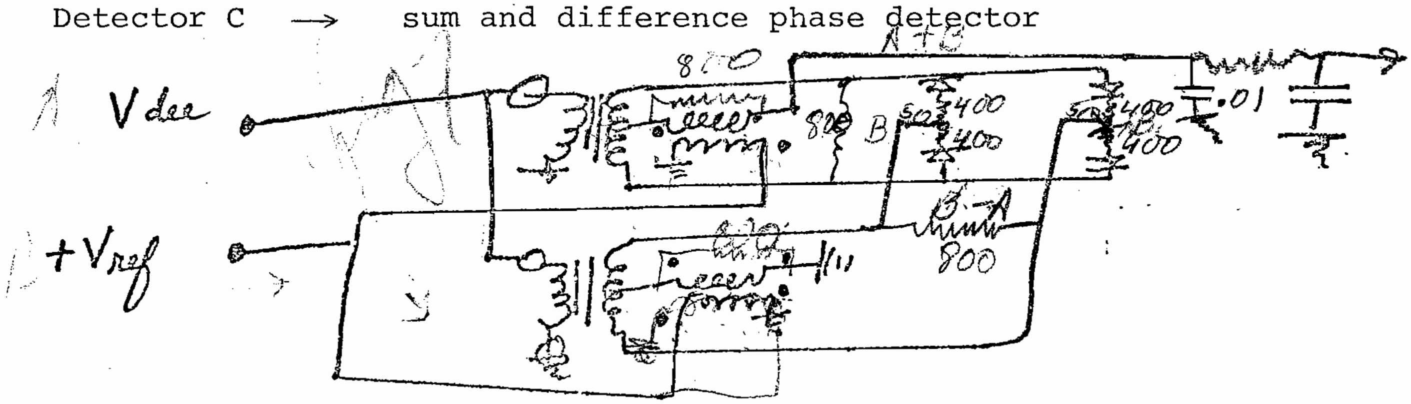
Detector B

coherent detection



V_{ref} is a constant 5 V rms signal of the same phase as V_{dee}.

Detector C → sum and difference phase detector

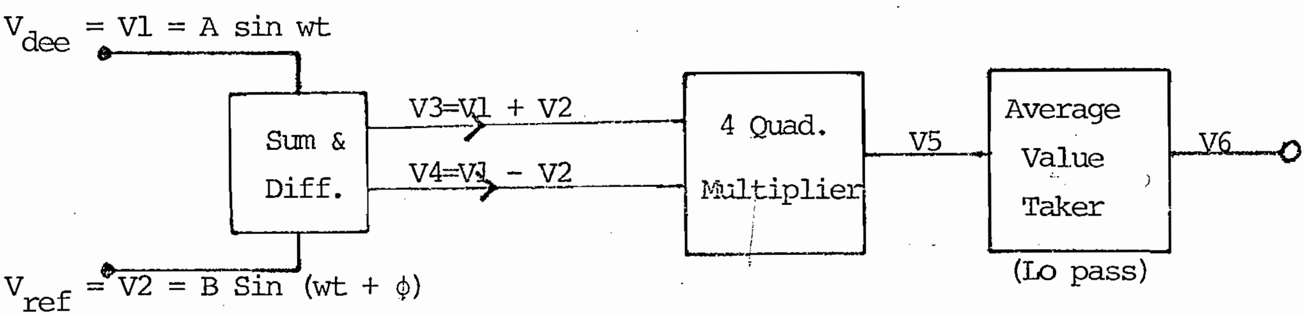


Detector scheme A, which is standard, suffers from having phase delay problems above 100 KHz, and in practice (i.e. at Princeton and Orsay) we have had to limit our unity gain amplitude response to 36 KHz, which means that at 360 Hz we can only have an open loop gain of 100, which is insufficient to reduce 360 Hz amplitude ripple to 10 PPM. Also it becomes very nonlinear below about .2 volts so it isn't very useful over a 1000 to 1 amplitude range. Also it is not very stable unless the diodes are temperature regulated, as conduction takes place in the highly temperature dependent region of conduction of the diodes.

Detector B is fast and linear and only has the drawback that it is not independent of the phase of V_{dee}. For example, if the dee phase is in error by 1^o then the error in voltage is $1-\text{Cos } 1^o = 2 \times 10^{-4}$.

So we will try scheme C, proposed by Q. Kerns and modified as follows: One starts with V_{dee} (varying from 5 mv to 5 volts) and a reference voltage of arbitrary phase, but a fixed amplitude. We insure that this reference rf voltage is stable by regulating its amplitude to be constant to 10 PPM in a manner to be described later.

For the benefit of the mathematically inclined we will now illustrate the principle of operation by analyzing exactly the following circuit.



$$V3 = (A+B \text{ Cos}\phi) \text{ Sin wt} + B \text{ cos wt Sin}\phi$$

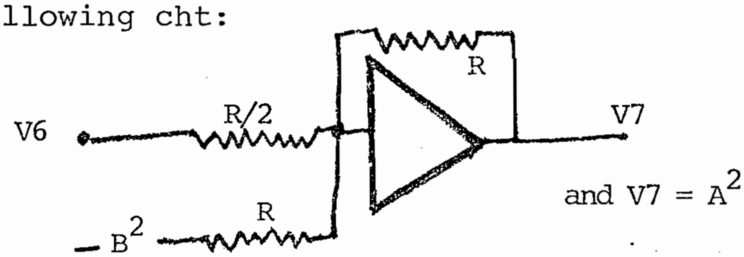
$$V4 = (A-B \text{ Cos}\phi) \text{ Sin wt} - B \text{ Cos wt Sin}\phi$$

$$V5 = V3 \times V4 = \sout{\text{Sin}^2\text{wt}(A^2-B^2) + \frac{B^2}{2} \text{ Sin } 2 \text{ wt Sin } 2\phi}\ \frac{A^2-B^2}{2} + \frac{B^2 \text{Cos}2\phi}{2} - \frac{A^2}{2}\text{Cos}2a$$

To take the average value we take

$$V6 = \frac{1}{2\pi} \int_0^{2\pi/\omega} V5 \, dt = \frac{A^2-B^2}{2}$$ so to make this an amplitude detector

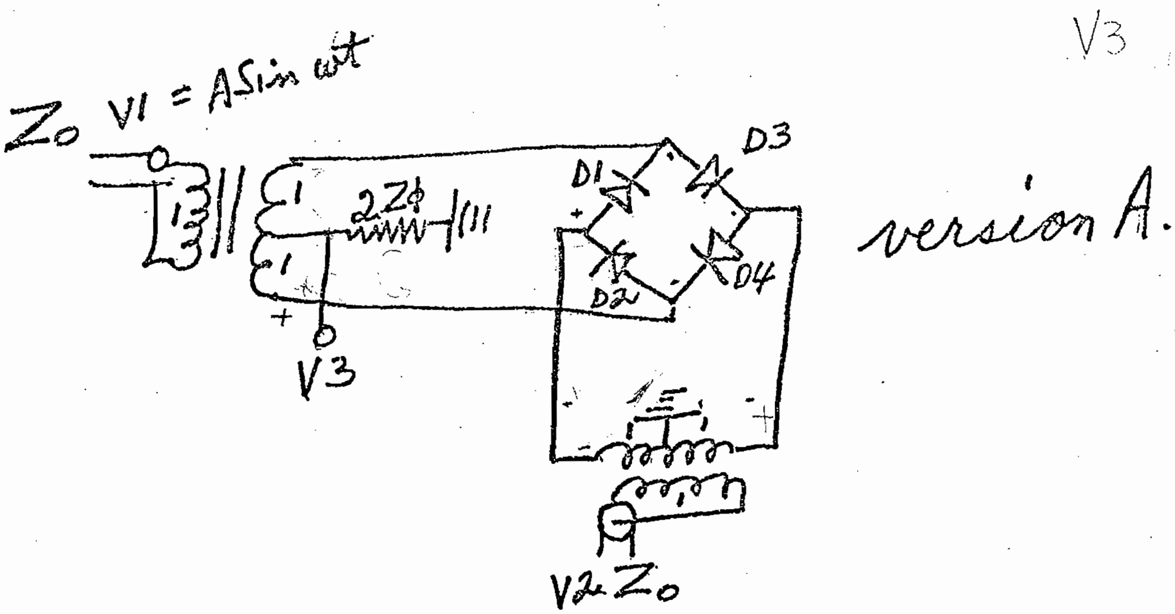
we add the following cht:



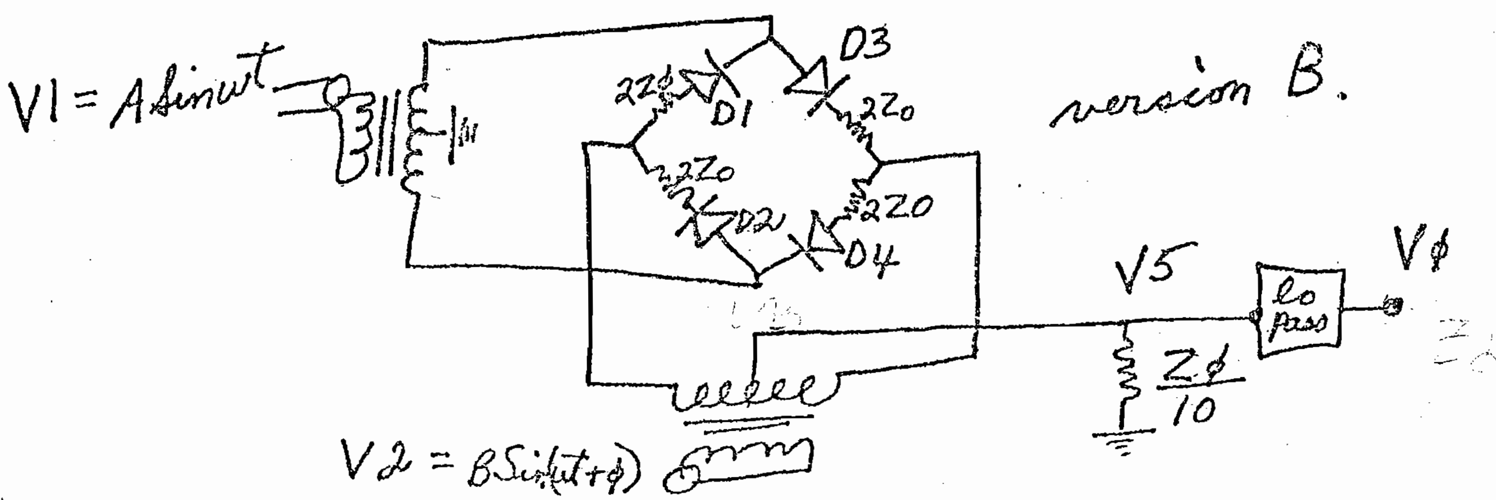
It is a square law detector.

However, in practice, instead of a true four quadrant multiplier we will use Riedel's version of a balanced mixer, or as some call it, ring modulator. But first, a few words on ring modulators seem in order.

Analysis of a balanced ring modulator



If the diodes are perfect, and V1 = V2, then both signals are are terminated and conduction only takes place through diodes D2 and D3, the others being unnecessary and V3 = a full wave rectified signal of amplitude 2A and of average value 4A/π. If V2 = 0 then D1 and D2 or D3 and D4 conduct and V1 is short circuited, so for a varying B it is obvious that the above modulator is no good, in fact stupid. So we go to Riedel's balanced, terminated, ring modulator



Now if B and ϕ vary, and A is a constant large enough to guarantee turning D1 and D2 or D3 and D4 on hard then both inputs are terminated with a SWVR of less than .05 and we can use it as a mixer, phase detector, or amplitude detector.

Mixer if A = B > 2 volts rms, then version A is the best choice, and a commercial balanced mixer should be used.

Amplitude detector.

V3 = A Sin wt is a reference signal where A is about 5 volts rms and very stable (10 PPM) for example.

V4 = B Sin wt + ϕ is the signal to be amplitude detected where B may vary from 5 mv to 5 volts. and ϕ from $-\pi/2$ to $\pi/2$.
we take the sum and difference with transformers so that

$$V1 = V3 + V4 = (A + B \cos\phi) \sin wt + B \cos wt \sin\phi$$

$$V2 = V3 - V4 = (A - B \cos\phi) \sin wt - B \cos wt \sin\phi$$

and analyze the output appearing as $V\phi$ in modulator version B. The extremes are easy to analyze:

if B = 0 then $I = \frac{2A}{2Z\phi}$ and $V\phi = \frac{4}{\pi} \times \frac{A}{10}$

if B = A, $\phi=0$, then V2= 0 and $V\phi = 0$

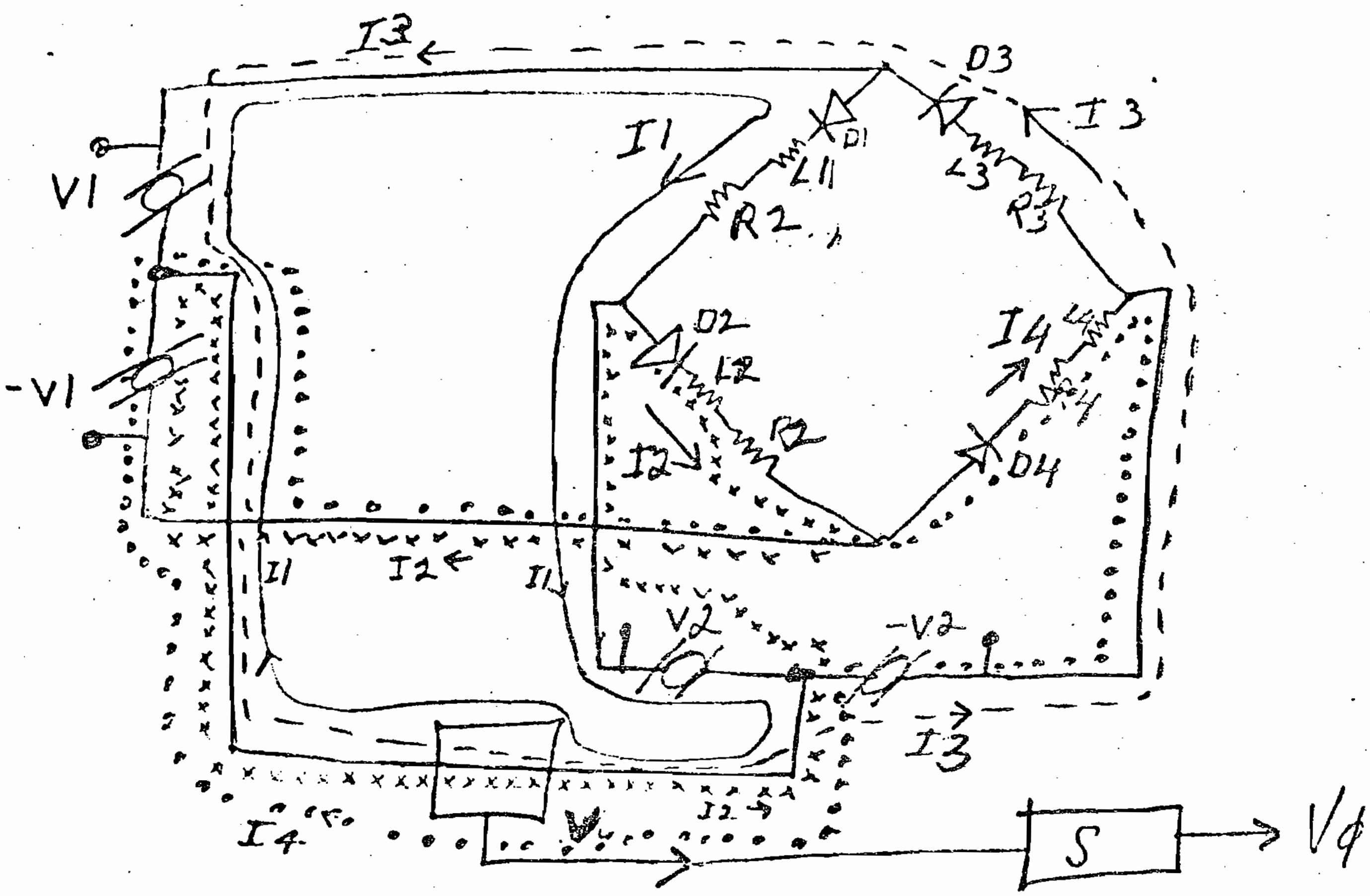
if B = A, $\phi=\pi/2$, then V2= A Cos wt and $V\phi = 0$

if B = A, $\phi= -\pi/2$, $V\phi = 0$

Let us pause to point out two important virtues of this circuit. First, during most of each cycle the diodes are either off or hard on, where their forward resistance is very much lower than the 200 ohms in series with them and they are operating in a region where the temperature coefficient of this resistance is small. This is true, even when A=0, because B alone will turn them on. This is the virtue of coherent detection. Secondly, the output response time is fast and the same for upgoing and downgoing A's, limited only by the delay of the Chebechev filter.

Now in order to get a complete and exact understanding of this circuit it seems advisable to write a simple computer program to analyze it. In this program we can actually put in the true value of the forward and reverse resistance of the diodes, as well as their temperature dependence. We could also put in the effects of carrier lifetime and delay due to mobility limitations. However we should have something for someone else to do -- so we will content ourselves with an approximation to the forward resistance, and the temperature coefficients and for simplicity assume that the output works into a zero impedance current to voltage converter.

Analysis of Riedel's Ring Modulator
(this is undoubtedly not original, but I have never seen it properly analyzed.)



We have 4 voltages and 4 currents, as shown, and a current to voltage converter V = 50 (I1 + I2 + I3 + I4) followed by an average taker.

$$V1 = A \sin wt$$
$$V2 = B \sin (wt + \phi)$$

and for mixer applications, $\phi = f(t)$.

R1 = R2 = R3 = R4 = 100 ohms

D1 = D2 = D3 = D4 = perfect diodes.

L1 = L2 = L3 = L4 = forward resistance

of the diodes, which we will approximate as f(V') where V' is the forward drop.

To analyze this circuit with a computer one can first draw a flow diagram, or simply state the logic functions, as follows:

1. if V1>V2 then I1 = (V1-V2)/(R1+L1)

   else I1 = 0 (or the reverse resistance)

2. if V2 > -V1 then I2 = (V2+V1)/(R2+L2)

   else I2 = 0

3. if V1 > -V2 then I4 = -(V2+V1)/(R4+L4)

   else I4 = 0

4. if -V1 > -V2 then I3 = (-V1 + V2)/(R3+L3)

5. I5 = I1 + I2 + I3 + I4

and then we set, for N = 1 to 4, $L(N) = \frac{\text{appropriate V}}{I(N)}$

and iterate twice to get a second order correction for the diode drops, where the L(N) are approximated by a logarithmic function of $\frac{V(+)}{I}$ from the manufacturer's data sheets.

This note has not come to a logical conclusion -- but it is only a note and therefore it is possible to just stop, so I do just that! It is obvious that at this time we need to do more work and less writing.